Coaxial SIW Filters – *When compactness meets flexibility*

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Substrate Integrated Waveguide (SIW) technology [1], [2] is nowadays a well-established and successful approach for implementing planar microwave and mm-wave filters with very stringent requirements in terms of Q-factor and integrability. Furthermore, packaging and EM shielding, power handling capabilities and low-cost batch manufacturing are also broadly recognized strengths of this technology. However, SIW filters are still larger than most of their planar counterparts, while advanced topologies are not always easy to accommodate, and filter reconfigurability usually leads to very complex implementations [3], [4], [5].

Many other SIW structures have been developed to overcome these drawbacks. For instance, more compact implementations can be obtained based on folded SIW cavities [6] or by loading SIW filters with dielectric rods [7], both of them at the expense of less standard manufacturing processes. Other techniques have exploited the EM field symmetry to create a virtual magnetic wall without requiring a multi-layer manufacturing technology, as it is shown in half-mode [8], quarter-mode [9] and even eight-mode [10] SIW structures. The loading of the SIW cavity with complementary split-ring resonators (CSRRs) [11] has also been a successful approach for further miniaturization. However, the former approaches can lead to a relevant Q-factor degradation [12] due to radiation issues at opened boundaries. On the other hand, different structures have been developed for reducing the dielectric losses with the sacrifice of compactness, fabrication simplicity and integration potential. In this sense, empty (ESIW) [13] and air-filled (AFSIW) [14] SIW are presented as good alternatives for applications where batch manufacturing as well as size can be traded off for very low loss.

Finally, a key feature of SIW technology has always been the ability to adapt classical and well-known metallic waveguide structures and topologies to a dielectric filled planar implementation. As a consequence, ridge evanescent-mode filters [15] in SIW technology were originally proposed as a way to achieve further miniaturization without degrading the intrinsic Q-factor while enabling both narrow and wide-band responses. Thus, in [16] an ultra-wideband filter was designed using ridge waveguide resonators defined by periodic via fences, while in [17] the concept was also extended to folded SIW for extra compactness. Other approaches were based on the introduction of capacitive loading posts inside an SIW cavity resonator [18], [19] or stubs within a below cut-off SIW section [20]. However, multi-layer LTCC technology was required in all previous structures, thus reducing the cost effectiveness of this approach compared to standard PCB technology.

The introduction of vertical capacitive posts into single-layer SIW cavity resonators was initially proposed in [21] as a way of perturbing the electric field distribution, and thus the resonant frequency of square SIW cavity resonators. As a preliminary concept, several posts were inserted and subsequently connected and disconnected from the bottom ground plane for generating different resonant frequencies, showing very interesting discrete tuning capabilities. This approach was latter explored by other groups for implementing reconfigurable SIW filters using p-i-n diodes [22] and RF MEMS [3] as switching elements. However, the focus was mainly put on how to locate the different switchable obstacles inside the cavity in order to obtain the desired set of resonant frequencies, but not in the resonator structure itself.

Following this idea, the well-known coaxial (or combline) cavity resonator [23] can be translated from metallic waveguide technology to a single-layer substrate integrated approach. Thus, a properly designed capacitive post, consisting on a short-circuited via hole capacitively-loaded at the top by an isolated metal patch, was firstly proposed in [25], while equivalent circuit, filter design procedure and measured results were discussed in [25], hence enabling the realization of a single-layer coaxial filters in SIW technology.

The main potential advantages of coaxial SIW structures are the following ones:

- By increasing the length of the post or the loading capacitance, a huge size reduction can be achieved compared to conventional SIW cavity resonators.
- Both inductive and capacitive couplings can be easily achieved on top layer for obtaining filter responses allocating finite transmission zeros [26]. Moreover, other advanced topologies as multi-mode resonators, singlets and doublets can be also implemented by translating their well-known metallic waveguide counterparts.
- The easy access to the loading capacitance enables the direct integration of lumped capacitors as well as tuning elements as varactors, p-i-n diodes or RF MEMS.

In the following sections, we will present the basic coaxial SIW filter topology as well as the latest advances regarding device miniaturization, design flexibility, advanced filtering responses, and implementation of tunable filters based on coaxial SIW structures.

Coaxial SIW Resonator

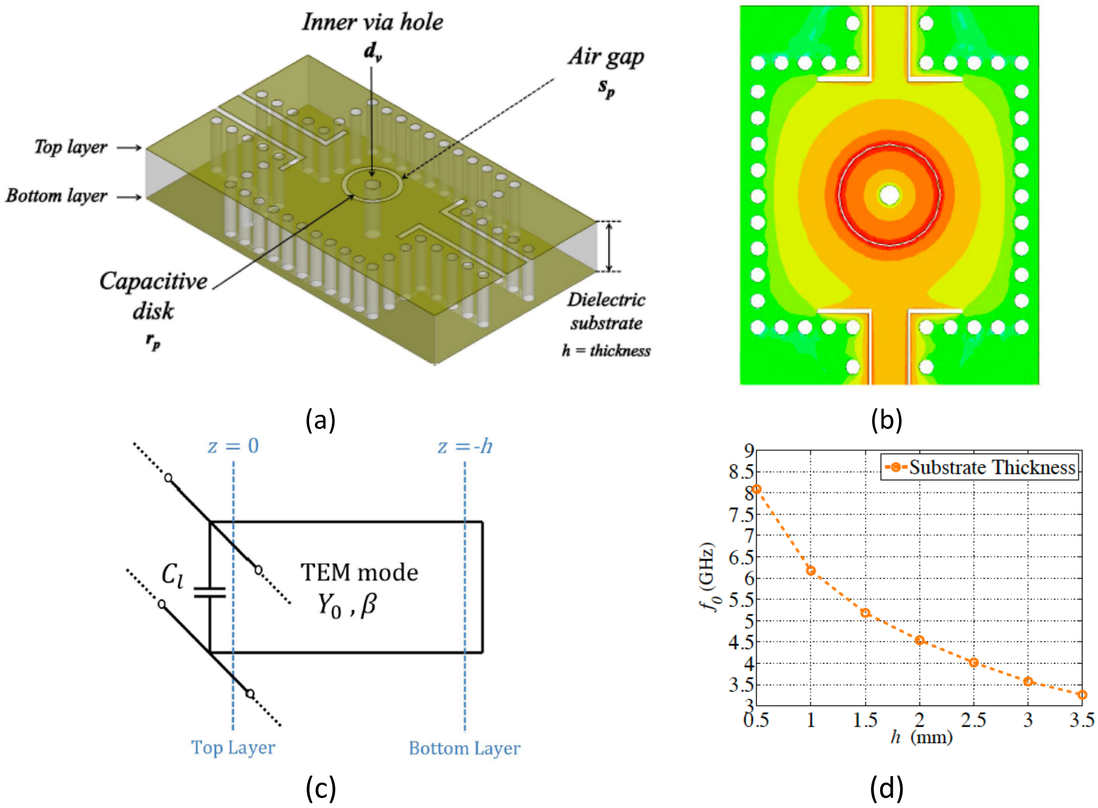


Figure 1. (a) A coaxial SIW resonator, (b) electric field distribution at the top plane, (c) equivalent circuit of the coaxial SIW resonator and (d) resonant frequency as a function of the substrate thickness for a resonator example.

The coaxial SIW concept was firstly presented in [24], where the authors proposed the implementation of a combline resonator by introducing a plated via hole at the center of a square SIW cavity. This plated via hole can be seen as the inner conductor of a short section of TEM-mode coaxial line along the vertical direction (i.e. the substrate thickness), while the outer conductor would be defined by the SIW post wall. In order to define the coaxial resonator, the inner via hole was short-circuited at the bottom ground plane, and ended on a metal patch at the top. A scheme of the structure is shown in Figure 1(a). The metal patch is then isolated from the top ground plane by an isolating gap, hence introducing an important capacitive load through the established fringing fields as can be seen from Figure 1(b).

In [25], a simple circuital model of the structure based on a short-circuited stub in parallel with a lumped loading capacitor was proposed and compared with full-wave EM simulations. The equivalent circuit is shown in Figure 1(c). The main parameters of the coaxial resonator were the characteristic admittance of the inductive section Y_0, given by the ratio between the inner via diameter d_v and the SIW cavity width W, and the value of the loading capacitance C_l. Given these two parameters, the resonant frequency ω_0 can be simply obtained by making the resonator susceptance B equal to zero

$$B(\omega_0) = \omega_0 C_l - Y_0 \cot \beta h \qquad (1)$$

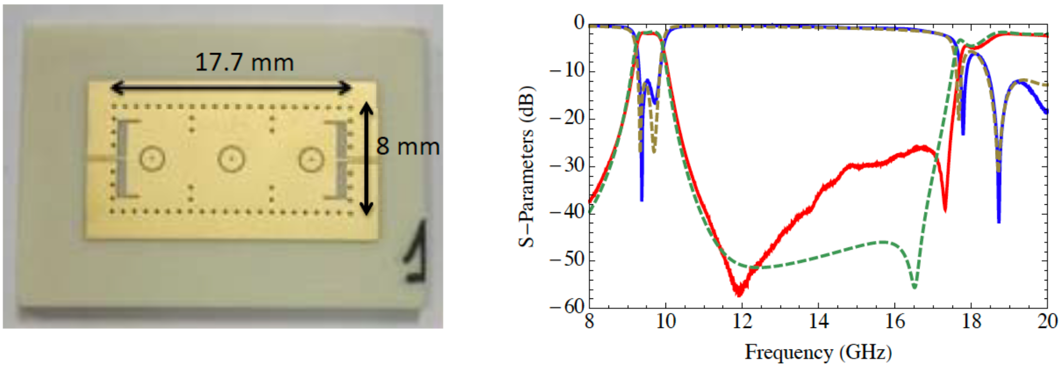
The agreement of this model with the EM simulations, even if a first-order approximation, is still very good and valuable for initial design. The resonant frequency of the structure is very dependent on the substrate thickness h (i.e. the resonator length), as can be seen in Figure 1(d), thus enabling an enormous size reduction for a given resonator layout.

Both the value of Y_0 and C_l are functions of the particular geometry of the structure. For instance, both square [25] and circular [27], [28] outer SIW cavities has been employed. In both cases, approximate expressions for the characteristic impedance of equivalent coaxial transmission lines are easily available [29], [30]. Regarding the loading capacitance, different shapes have been used for the top metal patch including circular [25], rectangular [31] and interdigital [32]. The latter being very interesting when highly loaded resonators are required, even if in this case SMD components could be also employed [33].

The dimensioning of the structure for a given resonant frequency is straightforward. Once the substrate material and thickness have been chosen, a suitable admittance value for the resonator is achieved by properly fixing the ratio (W/d_v). In order to obtain a compact structure, the minimum practical value for the inner via diameter is typically chosen. Next, the loading capacitance is adjusted by means of the metal patch perimeter or the width of the isolating gap s_p, although normally the width of the latter is also fixed to the maximum resolution of the manufacturing process (i.e. typically about 100 – 200 μm for standard PCB technology) for increasing the capacitive effect.

Lastly, the unloaded Q that can be achieved in this structure is close to the one obtained in conventional SIW resonators [25], with additional conductor losses coming from the inner conductor. An unloaded Q-factor about 200 – 300 can be easily obtained with low-cost hydrocarbon ceramic materials, while values up to 500 and greater might be reached using PTFE or low-loss microwave substrates.

Coaxial SIW Bandpass Filters



(a) (b)

Figure 2. (a) Photography of a 3-pole in-line bandpass filter at 9.8 GHz with magnetic coupling between resonators using inductive irises, and (b) measured results (solid) of the device compared with full-wave EM simulations (dashed) [25].

Bandpass filters based on coaxial SIW resonators can be easily implemented by coupling several cascaded resonators. In-line bandpass filters implementing all-pole responses can be realized by using inductive irises between adjacent resonators for controlling the magnetic coupling level as in [25], where a 3-pole 5% FBW Chebyshev filter at 9.8 GHz was designed and manufactured. CPW-to-SIW transitions are usually employed for input/output excitation as they enable the use of thicker substrates. In the example shown in Figure 2(a), the device footprint is less than 3 times the one required in a conventional SIW implementation. Thus, this approach enables to design very compact narrow-band filters with enhanced spurious-free band due to the smaller resonator length, as can be seen from the response in Figure 2(b).

Both positive and negative coupling networks, as well as couplings between non-adjacent resonators, are usually required for the introduction of finite transmission zeros (TZs) due to signal cancellation. Even if couplings between non-adjacent resonators can be easily obtained by means of folded topologies, there is no straightforward way in conventional SIW technology for obtaining both coupling signs.

The first approach reported in the literature consisted on a balanced microstrip line with plated via holes for phase inversion [34]. However, this requires etching the bottom ground plane thus reducing the integrability and self-packaging of the structure. Later, other solutions based on grounded coplanar lines [35], mixed-couplings [36] or over-moded cavities [37] were proposed, although they typically introduce some limitations in terms of coupling level, topology or area required.

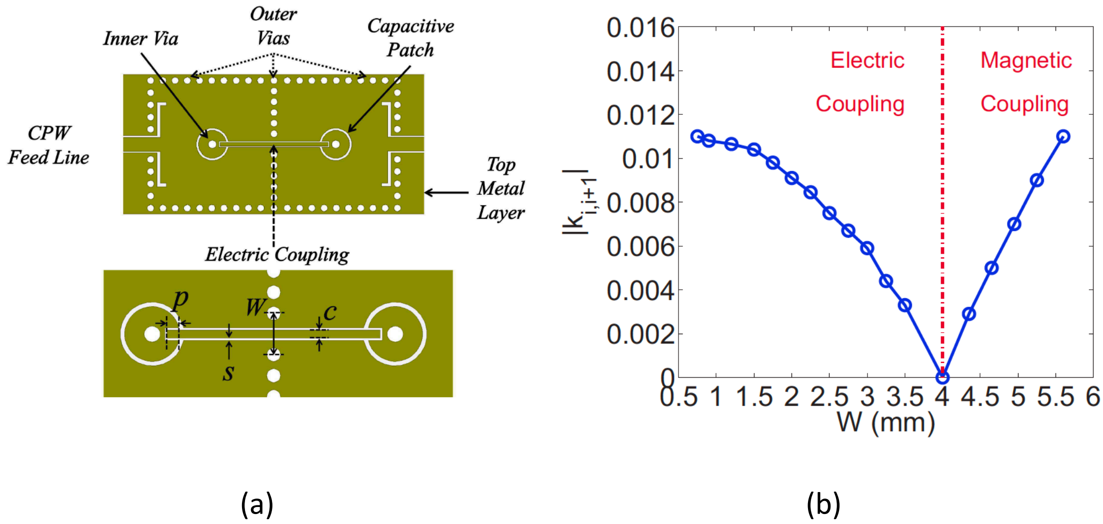
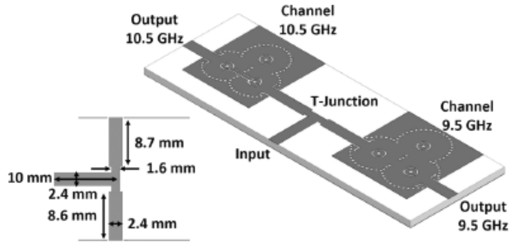


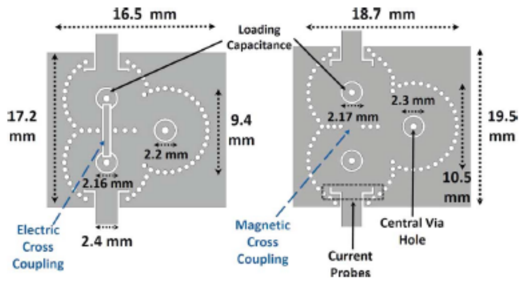
Figure 3. (a) Layout of the electric coupling probe for coaxial SIW resonators and (b) coupling coefficient variation versus the width of the iris window *W* for a given probe dimensions and penetration into the capacitive patch [38].

However, negative (or electric) coupling in coaxial SIW can be easily achieved in a single-layer technology due to the direct access to the capacitive section of the resonator. Thus, a single open-circuited coplanar probe can be used for introducing electric coupling between resonators. Different parameters can be used for controlling the coupling level as the penetration depth into the capacitive metal patch, the probe dimensions or the width of the inductive iris between cavities (i.e. by modifying the intrinsic magnetic coupling between resonators and hence counteracting the electric contribution). A scheme of the electric coupling probe for coaxial SIW is shown in Figure

3(a). As can be seen, the sign of the coupling can be reversed by decreasing the magnetic coupling contribution introduced by the inductive iris between resonators as depicted in Figure 3(b). This approach was extensively studied in [38], and employed for designing in-line filters with electric coupling [38], pseudo-elliptic filters [39] as well as diplexers [27].



(a)



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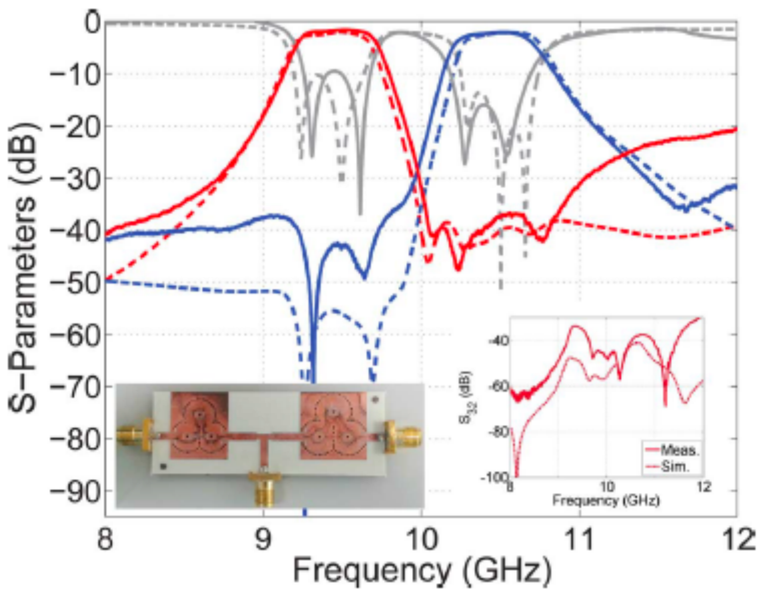
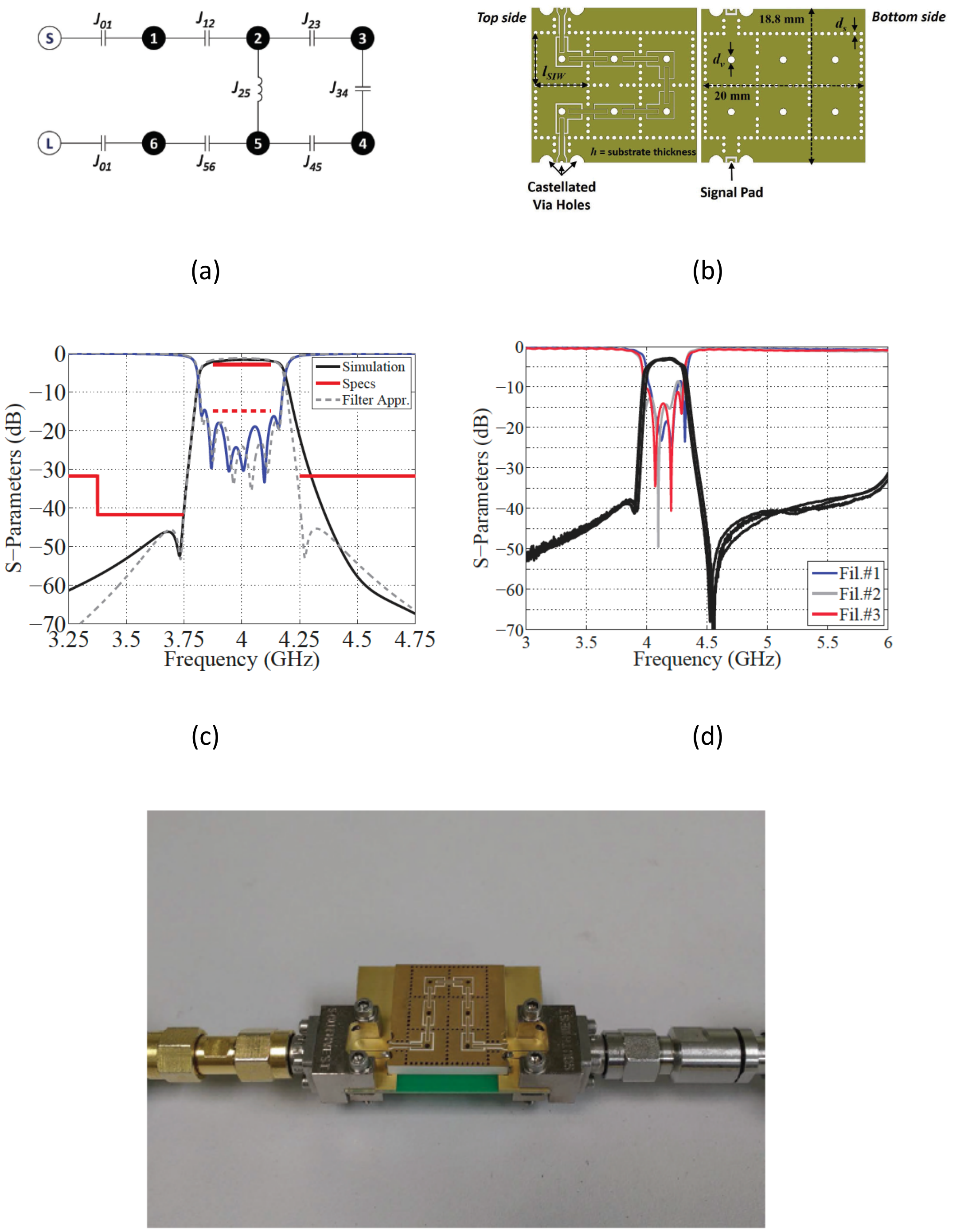


(c)

Figure 4. (a)Scheme of the proposed diplexer, (b) detailed view of the channel filters with negative (left) and positive (right) cross-couplings, and (c) measured and simulated results [27].



(e)

Figure 5. (a) Filter topology, (b) device structure, (c) synthesis and EM simulation results, (d) measured results for three different devices showing good repeatability, and (e) photograph of the manufactured device under measurement [31].

For instance, an X-band diplexer was designed and manufactured in [27] with TZs placed below and above each passband in order to improve rejection and isolation levels. A scheme of the proposed design as well as the layout and measured results are shown in Figure 4. The device was manufactured in 1.52 mm-thick RO4003C. The whole device dimensions are 58.4 x 18.7 mm^2 excluding I/O feeding lines. Maximum IL of 1.6 and 2.1 dB are obtained at the lower and upper channels respectively, while stopband suppression is better than 37.5 dB at both bands.

In order to achieve filter miniaturization, thick ceramic-filled PCB substrates with high permittivity values can be employed. Thus, the impedance of the resonator is reduced and the capacitive loading increased by means of metal patches with greater perimeter. Following this approach, a 6th order filter with two TZs was presented in [31]. The topology of the filter and a scheme of the device are shown in Figure 5(a) and (b). Electric couplings are used for input/output and direct-coupled resonators, while inductive irises are used for introducing a cross-coupling between resonators 2 and 5 thus producing a pseudo-elliptic response. A 2.54 mm-thick Rogers TMM10i substrate (e=9.8, tan d=$2\cdot10^{-3}$) was chosen for implementing the device. The filter included vertical CPW-to-CPW transitions for enabling SMD packaging of the structure and assembly on top of a carrier substrate. The footprint is just 20 x 18.8 mm^2, showing a high degree of miniaturization at C-band.

Even if the design and optimization process resulted in a very satisfactory design, as can be seen in Figure 5(c), the measured results presented a small shift towards higher frequencies due to the trapezoidal shape of the gaps etched on the top substrate, which slightly reduce the effective loading capacitance. However, this can be compensated in a second manufacturing run by slightly reducing the isolating gaps, as the repeatability within the same lot is very good. A photograph of the manufactured filter under measurement is shown in Figure 5(d).

Other examples of very compact coaxial SIW filters have been recently presented. For instance, a compact quasi-elliptic combline filter in single-layer SIW technology using mixed couplings was presented in [40]. Other approaches rely on multi-layer PCB technology for embedding the capacitive post in one of the inner substrate layers, thus enabling a huge increase of the loading capacitance and therefore producing very compact implementations [41]. A similar concept can be further exploited in LTCC technology for achieving miniaturization factors close to 97% [19], [42].

Flexible and Advanced Filtering Responses

Singlets and Cascaded Singlets

The selectivity provided by a filter in the vicinity of the passband can be improved by generating TZs rather than by increasing the number of its poles [43]. The idea of exploiting additional signal paths between the source-load and resonators is not new but it is extremely useful for microwave filter designers [44], [45], [46]. The most common form to generate TZs is to arrange the resonators in such a way that cross-couplings among resonators can be created, leading to signal cancellations at the output as we have discussed in the previous Section. This strategy can produce N TZs with N resonant elements if there is also input-output coupling (otherwise, the maximum number of TZs is $N-2$). An alternative form for TZ generation is by adding additional paths that bypass the resonant element itself, which allows the design of advanced filtering responses also having N TZs. To do that, several singlets (i.e. the most basic building block that contains one resonator and generates one TZ, whose definition was firstly envisaged in [47]) have to be cascaded by using non-resonating nodes (NRN) to attain the modular design of such elliptic filters.

The routing scheme of a singlet is shown in Figure 6(a). This topology can be easily designed in coaxial SIW technology by means of a direct source-load coupling that can be implemented at the top layer. Furthermore, this scheme in coaxial SIW enables simple realizations of different coupling signs, which are needed for a flexible positioning of the TZ.

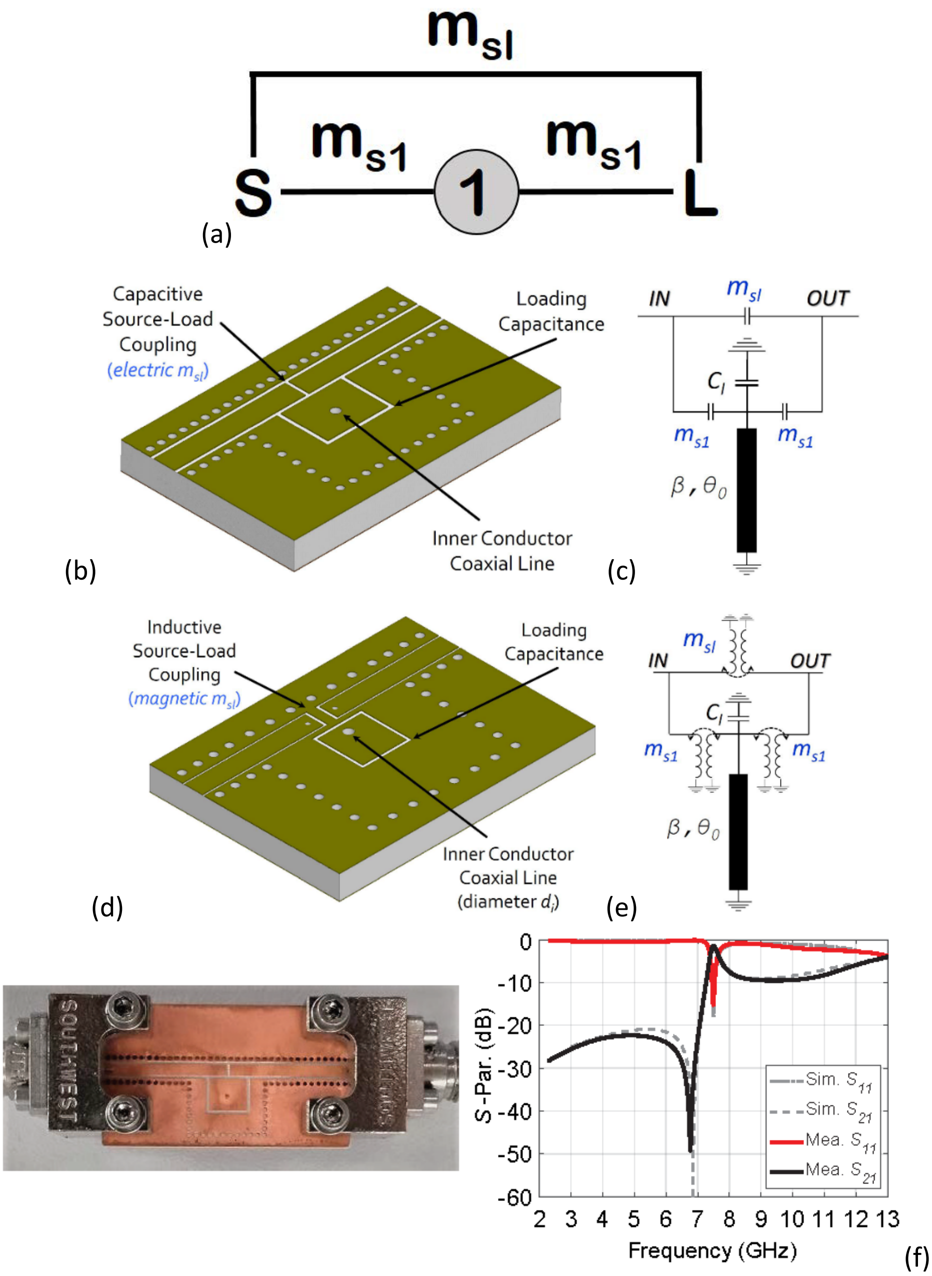


Figure 6: (a) Routing scheme of a singlet. (b) Coaxial SIW singlet with TZ below the passband with (c) its equivalent circuit. (d) Coaxial SIW singlet with TZ above the passband with (e) its equivalent circuit [49]. (f) Practical realization of a singlet in coaxial SIW topology with TZ below the passband [48].

Thus, a fully integrated technique for the design of coaxial SIW singlet was successfully proved in [48] by using a direct electric source-load coupling m_{sl}, which is based on a distributed series capacitance created on the CPW feeding line. The TZ in this singlet is generated below the passband, and its position is set by the spacing and width of the gap that establishes this capacitance. Figures 6(b) and (c) show the layout of such coaxial SIW singlet, and its equivalent circuit, respectively. In addition, Figure 6(e) depicts a practical implementation of a coaxial SIW singlet centered at 7.5 GHz with a TZ located at 6.8 GHz.

An evolution of this coaxial SIW singlet was then proposed in [49] so that the TZ could be generated above the passband. Therefore, the m_{sl} is based on a magnetic coupling mechanism, which has been obtained by means of a standard CPW-to-SIW transition that makes use of a current probe (i.e. a plated via hole is inserted at the end of the CPW feeding line), as shown in Figure 6(d).

The various coaxial SIW singlets can be used as building blocks of passband filters with symmetric quasi-elliptic response, or asymmetric responses with steep filter skirts. Once the singlets have been designed, these can be coupled among them by means of the quarter-wave coplanar waveguide (CPW) NRNs, whose resonant frequencies should quite different from the desired working frequency.

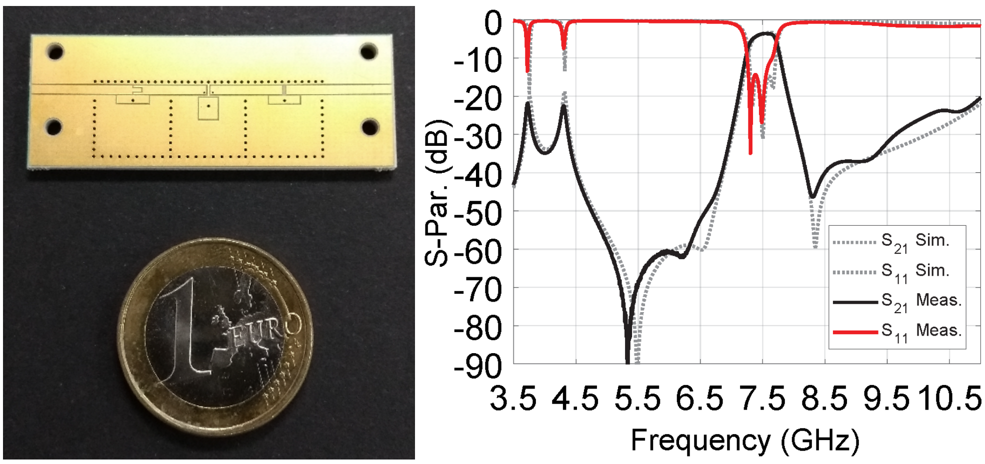
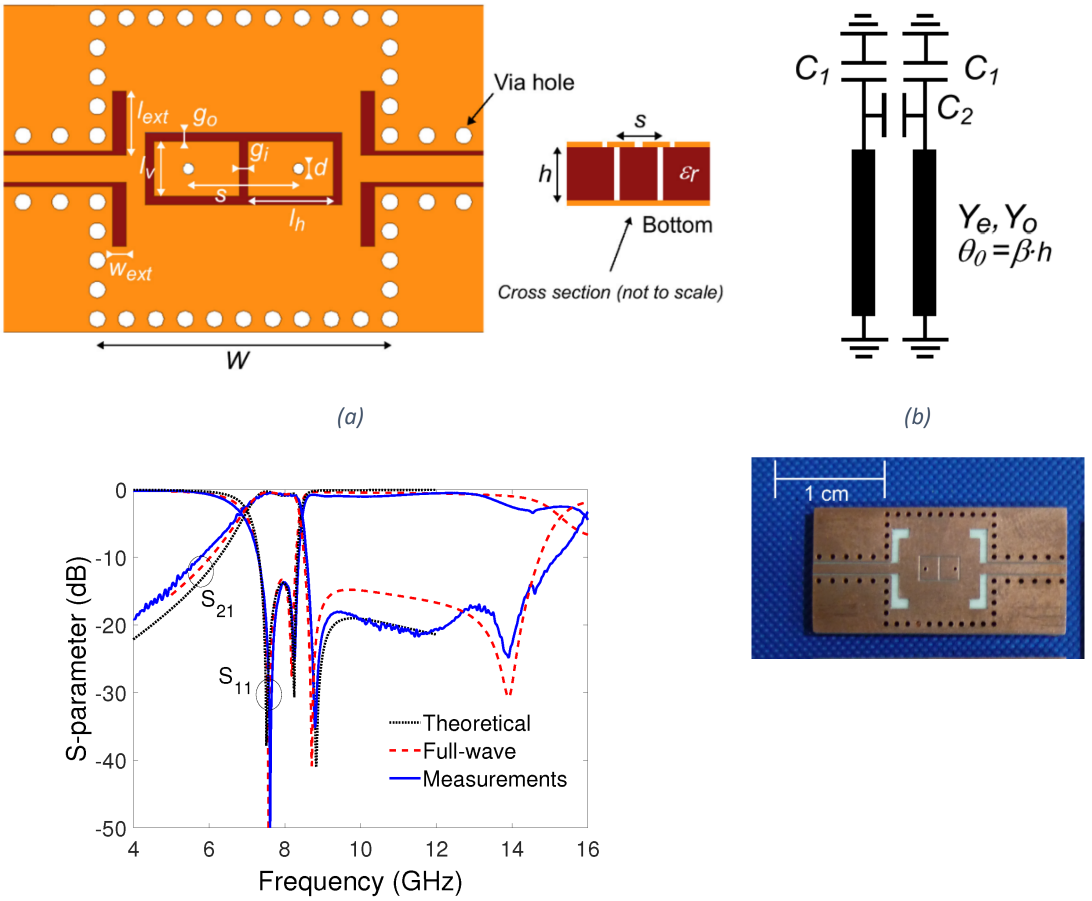
(a)  (b)

Figure 7: Three-pole elliptic BPF based on three cascaded singlets: (a) photograph, and (b) measured and simulated responses.

Figure 7 shows the performance of a third-order elliptic filter, showing two TZs below, and one TZ above the passband, which fully benefits from the design flexibility offered by these building blocks. Fabricated by using a 1.52-thick Rogers R4003C, this BPF has been designed by cascading three singlets with different source-load couplings. Specifically, two singlets show electric bypass coupling (i.e. the first and the third resonators responsible for lower TZs), while the second one has a magnetic bypass coupling, which generates the upper TZ. The filter centre frequency is 7.5 GHz whereas the FBW measured at the level of RL = 15 dB is 5% (i.e. BW = 380 MHz). The filter rejection is extremely good between 4.5 GHz and 6.7 GHz, being better than 48 dB, while its size of 9.8 × 28.5 mm^2 enables a 50% of area reduction with respect to an equivalent TE_{101}-based SIW implementation.

Dual-mode Coaxial SIW

Similarly, to its equivalent waveguide counterpart, the coaxial SIW topology is able to support two propagation modes [50], [51] thanks to its extremely flexible 2.5D geometry that enables the introduction of two (or more) conducting posts connected to their own patches in the very same SIW cavity resonator. Figure 8 shows the layout of a dual-mode coaxial SIW cavity, its cross-sectional section and the corresponding equivalent circuit [50], [51].



(c) (d)

Figure 8: (a) Layout of a dual mode coaxial SIW resonator, (b) equivalent model, (c) theoretical, 3D EM simulated and measured responses of the implemented filter and (d) photograph of the fabricated device [51].

On the one hand, two via holes are symmetrically inserted with respect to the SIW cavity center, and these are then connected to separate capacitive patches at the top layer. On the other hand, there is no difference with the standard coaxial SIW at the bottom layer, where those via holes are short-circuited to ground. Such dual-mode configuration has the clear advantage of enabling a sharp improvement of the achievable BW, out-of-band rejection and compactness. Since there are two separate coupling paths between the source and the load in a doublet, this topology features one TZ in the frequency response.

This simple topology offers the flexibility of designing of both narrowband and wideband compact filters with improved rejection providing a high degree of flexibility. To corroborate the usefulness of this topology, two filters exhibiting wide- and narrow-band responses have been designed, fabricated and measured [51]. The layout, and a photograph of the wideband BPF with a centre frequency f_0 = 8 GHz, and FBW_{3dB} = 20 % that shows a square SIW cavity size of 10.6 mm is visible in Figure 8. Even if a single-layer PCB manufacturing has been used, those designed dual-mode filters do enable a 75% of area reduction, compared to a standard SIW filter implementation. The area reduction might be stretch up to 90% with a proper resonator design. The stopband of the filter is spurious free up to 16 GHz, which is around $2\times f_0$, where the SIW cavity TE_{101} mode is excited.

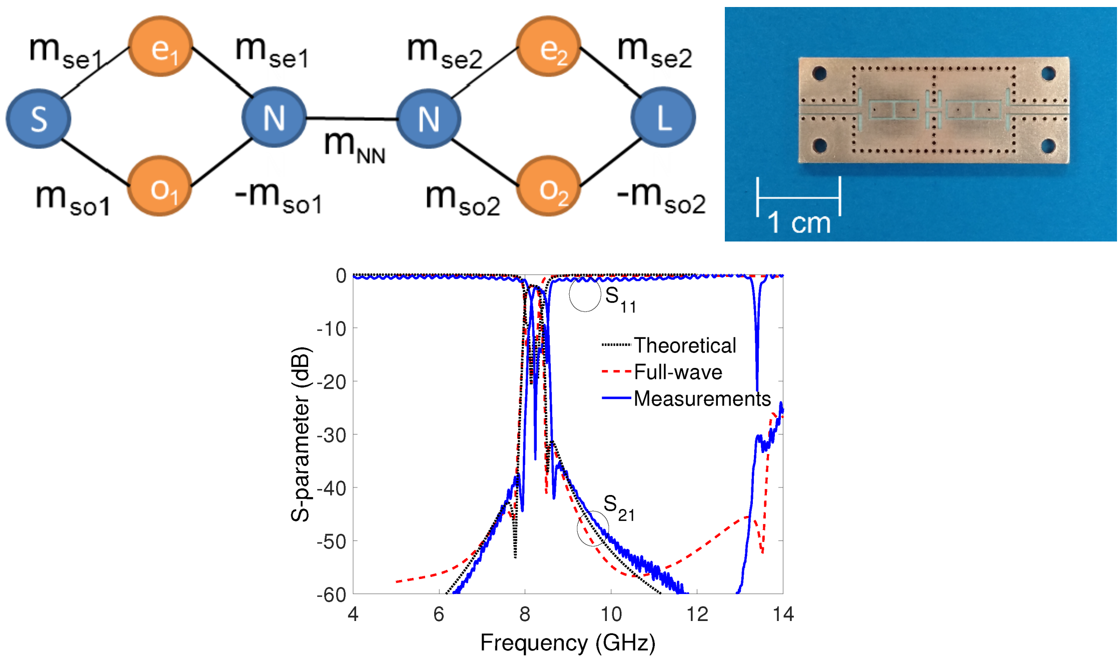


Figure 9: A 4-pole quasi-elliptic type BPF by cascading two dual-mode SIW coaxial cavities. (a) Photograph. (b) Theoretical, full wave simulated and measured responses of the fabricated BPF [50].

Cascading dual-mode building blocks by means of NRNs is a straightforward way to implement complex filtering responses with high-selectivity performance and multiple TZs. For instance, Figure 9 shows the layout of a 4th order narrow-band BPF with FBW_{3dB} = 4% presenting a quasi-elliptic type response since two TZs are generated above and below the passband. Note that, in Figure 11, e_i and o_i indicate the even- and odd-mode resonator node, respectively, whereas $m_{se,i}$ and $m_{so,i}$ are the main direct couplings to the even- and odd-mode, respectively. A clear advantage of this approach is that each doublet can be designed individually.

Such promising scheme can be further developed in order to design dual-band filters that feature two independently controllable bandwidths in just one single SIW cavity resonator [49], as shown in Figure 10, which means a dramatic improvement of the component miniaturization. Indeed, the proposed doublet configuration with electric-based direct bypass coupling is employed to design two dual-band filters with TZs at both sides of the passbands (i.e. four TZs placed at real frequencies), and with different bandwidth requirements. A very interesting property is that both passbands can be independently controlled, since there is no coupling between the upper and lower side of the cavity. The filter shown in Figure 12 as DB2 operates at 5.25 GHz and 7.5 GHz, providing FBW_{1dB} = 12.8% (i.e. BW = 670 MHz) and FBW_{1dB} = 2.7% (i.e. BW = 200 MHz), respectively.

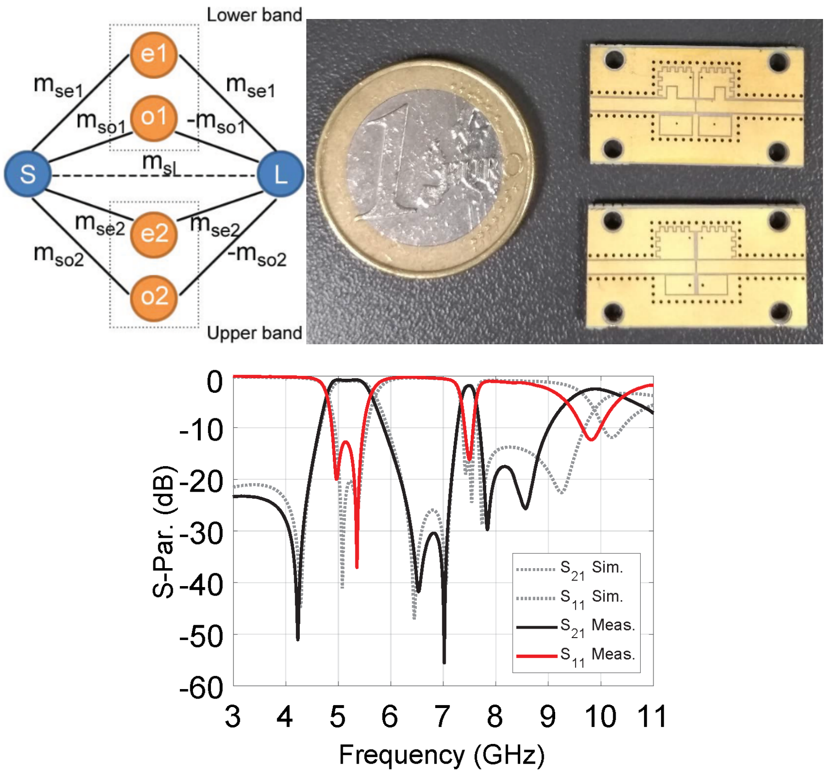


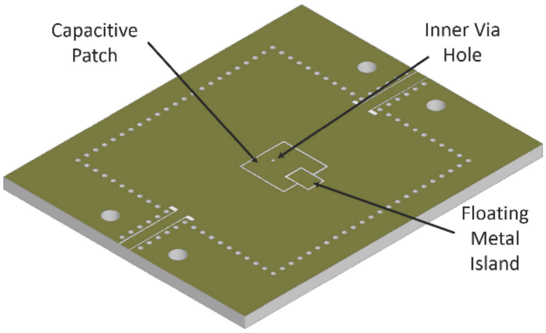
Figure 10: (a) Routing coupling scheme of the proposed dual-band filter based on coaxial SIW doublet. (b) Photograph of two filter prototypes (DB1 below, DB2 above) with different passband configurations, and (c) the measured response of the filter DB2 [49].

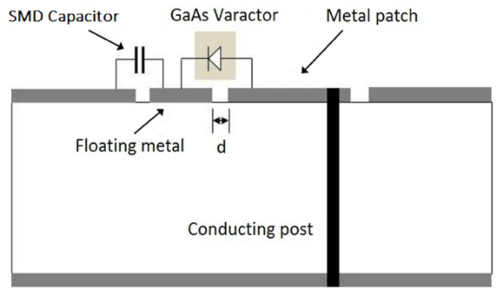
Tunable Coaxial SIW Filters

The availability of moderate-to-high *Q*-factor tunable resonators is of primary concern in the design of low-loss reconfigurable filters for microwave applications since these are going to be of enormous importance in the development of future multi-band multi-standard miniaturized front-ends, both for satellite and terrestrial applications. Nevertheless, when developing reconfigurable BPFs, so far it has been difficult to find an acceptable compromise among fast tuning speed, high Q_u, wide tuning range, low manufacturing cost and reliability, which are very important characteristics to offer in most applications. Likewise, to increase the transceiver reconfigurability, such devices must also offer bandwidth tuning, which is non-trivial feature to be addressed.

The main advantage of the coaxial SIW topology when implementing frequency-agility in SIW components stems from the fact that the integrated loading capacitance is readily accessible at the top metal layer. This means that the frequency tuning can be simply achieved by assembling suitable tuning elements that are called upon to modulate such loading capacitance. Such tuning can be ultimately classified in two categories depending on the manner in which the resonant frequency is changed: analog and discrete tuning. Both mechanisms have been widely demonstrated in the literature to be applicable seamlessly to the coaxial SIW topology.

The first example of tunable coaxial SIW structure was proposed in [52]. Specifically, a continuously tunable coaxial SIW resonator, which made use of GaAs varactor diode MA46H200 from MACOM, has been designed for working at S-band, concretely between 2 and 3 GHz, whose simple layout is shown in Figure 11.





(a)

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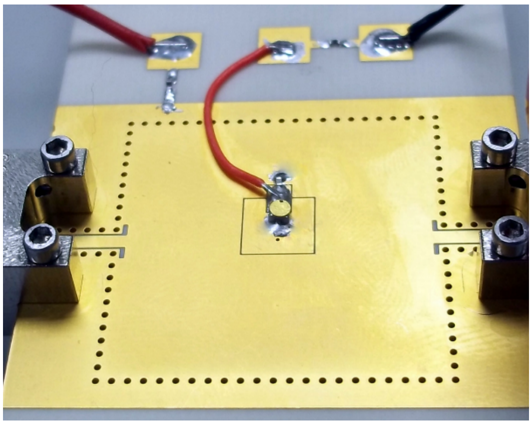
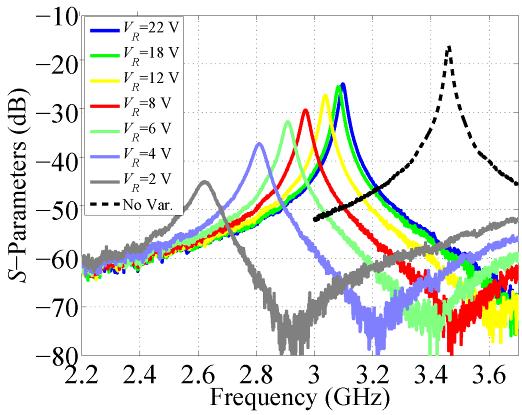




(c)

(d)

Figure 11: (a) S-band varactor-tuned coaxial SIW resonator, (b) its cross-sectional view, (c) manufactured tunable resonator and (d) measured results (only the S21 parameter is shown) [52].

A floating metal patch needs to be introduced between the capacitive patch and the SIW ground plane, being completely isolated from them, in order to allow for the assembly and biasing of the varactor diode, as it can be seen in Figure 11(b). In addition, a fixed SMD capacitor is then inserted in series between the floating metal and ground to increase the tuning range that would be reduced by the small capacitance generated by the floating metal air gap.

Figure 11(c) shows a photograph of the tunable resonator prototype, whereas the measured results are depicted in Figure 11(d). The varactor-tuned resonator shows a wide tuning range of almost 20%, between 2.64 GHz and 3.1 GHz (for reverse bias voltage V_b 2-22 V). In addition, the extracted Q_u ranges from 40 at 2.64 GHz to 160 at 3.1 GHz, while the static resonator presents a Q_u of 220.

Such structure has been successively exploited for the design of reconfigurable filters, as it has been done in [53] where a 2-pole bandpass filter was designed. However, the major drawback of this preliminary filter was the important variation of its fractional bandwidth (FBW) along the tuning range, specifically, from 5% at 2.88 GHz to 2.3% at 2.5 GHz. This issue is due to the short resonator electric length (i.e. $\theta_0 = 11°$) that provokes a wide change of the coupling coefficients over frequency.

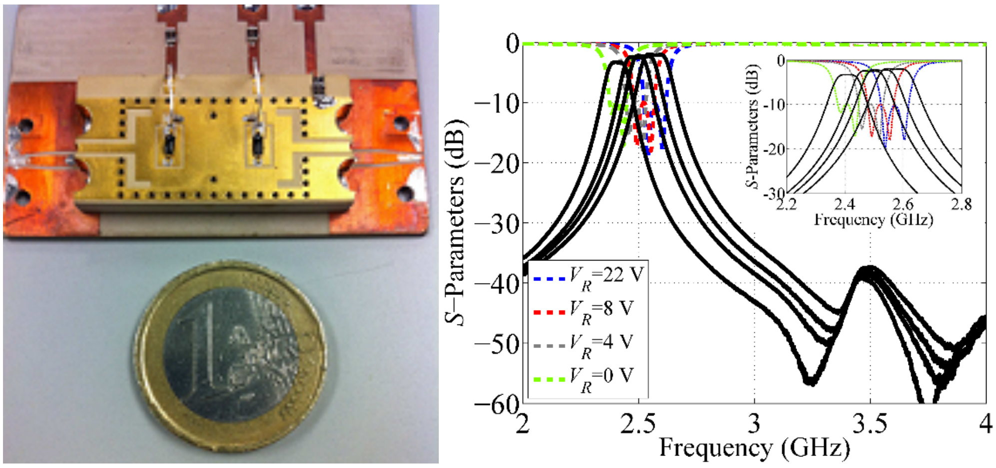


Figure 12: (a) The manufactured 2-pole varactor tunable filter and (b) its measured results [56].

To overcome such BW variation, the use of tunable coupling mechanisms is needed, which are based on variable reactive elements [18], [54]. However, as demonstrated in [55], a proper selection of the resonator length can help to reduce the variation of the inverter values for keeping a constant fractional bandwidth. To do that, thicker substrates must be chosen for implementing longer coaxial SIW resonators as it has been done in [56].

Figure 12 shows a photograph and measurements of the 2-pole tunable filter designed to work in the S-band, which has been implemented in a 3.81 mm-thick laminate of Rogers TMM10i substrate (ε_r = 9.8 and tanδ = 0.002) and whose size is just 16.5 × 33 mm^2. Measurements show a continuous tuning from 2.4 GHz (IL = 3.4 dB) to 2.6 GHz (IL = 1.9 dB) with FBW_{1dB} (from 3.9% at 2.6 GHz to 3.3% at 2.4 GHz) and almost constant RLs. Good rejection has been easily obtained by using the same tuning mechanism previously described. Moreover, the Q_u is estimated to be 145 @ 2.6 GHz and 110 @ 2.4 GHz, being more than 100 in the whole tuning range.

Besides, the use of tunable couplings in reconfigurable coaxial SIW filters allow designers to achieve extremely high levels of response reconfiguration. Such approach has been reported in [57], [58]. Using as building block the varactor tunable resonator previously described, mixed inter-resonators and external couplings are now employed to implement bandwidth tuning. On the one hand, the primary dominant coupling path is between the magnetic fields created by the two closely placed inner posts of adjacent resonators, with fixed amount of coupling. On the other hand, a weaker secondary path has been generated through the electric fields created by lumped varactors that connect directly the capacitive patches of adjacent resonators and with the feeding lines too (see Figure 13a-b). Here, low Q tuning elements can be used to tune the BW without significantly affecting the resonator Q_u. Then, to increase the tuning range several varactor diodes in back-to-back configuration have been assembled on the resonator capacitive patch [33], [59]. The huge loading effect enables an additional filter miniaturization, which operates in the UHF- and L-band showing measured resonator Q_u in the range 100-200 with off-the-shelf varactor diodes.

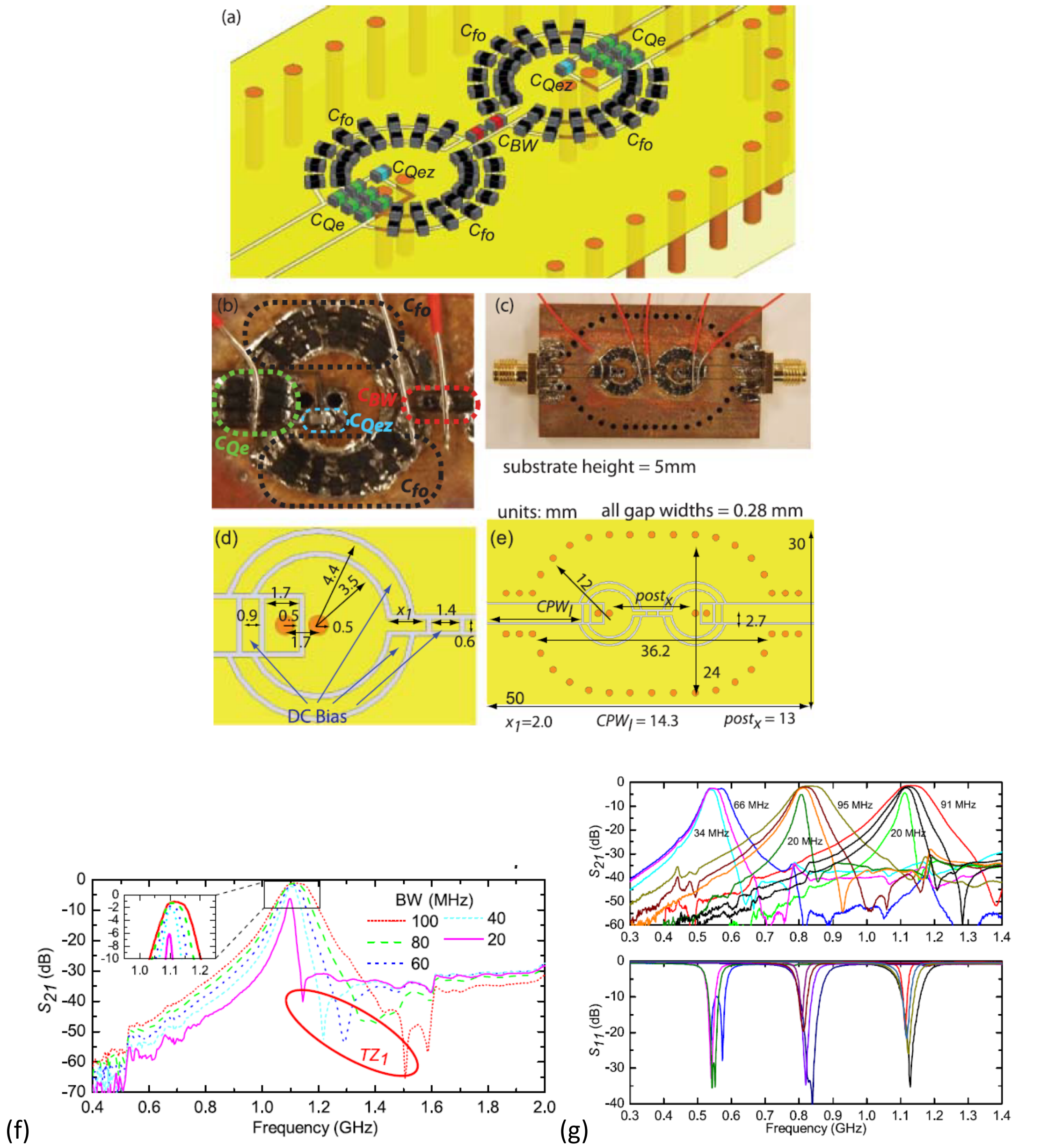


Figure 13: (a) Designed coaxial SIW filter with two ring gaps and back-to-back varactors, (b) close-up and (c) full-view of the fabricated filter, (d)-(e) dimensions of the filter, (f) measured S_{21} showing decreasing BW as the varactor capacitance increases, and (g) measured S21 and S11 for both tunable center frequency and BW [58].

Figure 13(f) and (g) shows the remarkable results that such filter implementation with tunable center frequency and BW can achieve. In particular, the FBW can be controlled as follows: 6.18%–12% at 0.55 GHz, 2.38%–11.3% at 0.84 GHz, and 1.77%–8.05% at 1.1 GHz. The extracted Q_u for the resonators is approximately 80 at 0.5 GHz and 200 at 1.1

GHz. What’s more, this varactor tunable filter is able to show a constant BW anywhere from 34 to 66 MHz.

Nowadays, tunable dual-mode coaxial SIW resonators have emerged as a very attractive building block of reconfigurable filters as they enable the realization of a wide variety of filtering responses while maintaining an extremely compact size and planar integration. One specifically desired feature in such resonators is the ability to control the resonant frequency of each mode independently and generate one TZ in the frequency response.

Indeed, in [60], two vertically-moving tuners enabling independent tuning for each mode of a dual-mode SIW resonator have allowed the design of a high-Q, widely tunable second-order BPF. The layout and a photograph of the tunable filter prototype are shown in Figure 14. Apart for the wide tuning range, this doublet-configuration permits the flexible control of the TZ on the lower or the upper side of the passband.

To enable this tuning, two flat tuners with conductive surfaces are positioned on top of the resonator creating two different controllable gaps (i.e. g_1, g_2) with the primary role of modifying the embedded capacitances (i.e. C_1 and C_2 shown in Figure 14c) that belong to the dual-mode resonator. The filter has been designed on a 5.08mm-thick Rogers TMM3 substrate, whereas the tuners are cut from a 0.75mm-thick Rogers 4003 substrate, and attached to a M3L linear actuator in order to create the vertical motion. The measured results are very good with a wide tuning range (2 GHz-3.2 GHz), and high Q_u better than 200 between 1.7 GHz and 3 GHz.

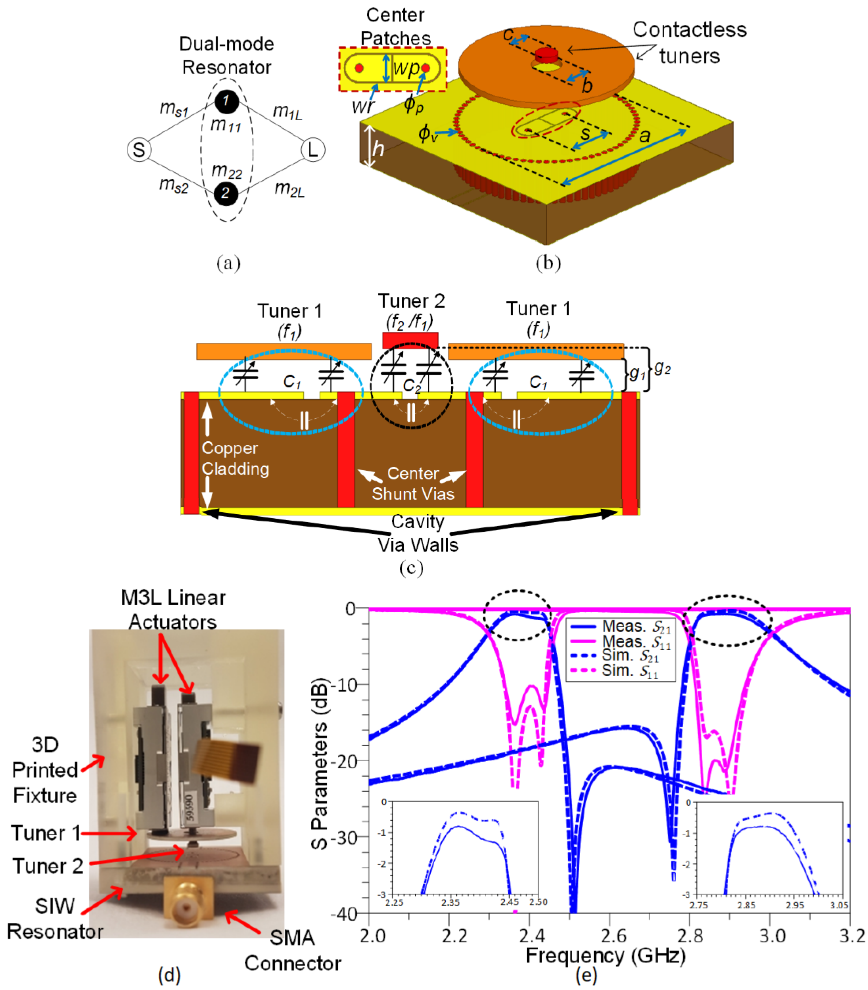


Figure 14: (a) General coupling diagram with a dual-mode resonator. (b) Proposed dual-mode SIW resonator. (c) Cross-section of proposed resonator. (d) Picture of fabricated filter with fixture and M3L linear actuators. (e) Measurement and simulation comparison for two tuning states [60].

However, such approach has the disadvantage to need a more complex tuning structure, resulting in a bulky filter implementation. At the expense of poorer *Q* performance (i.e. in the range 50-150), this promising tunable doublet topology can be also deployed by using SMD varactor diodes [61], or manually-adjustable thin-film trimmer capacitors [62].

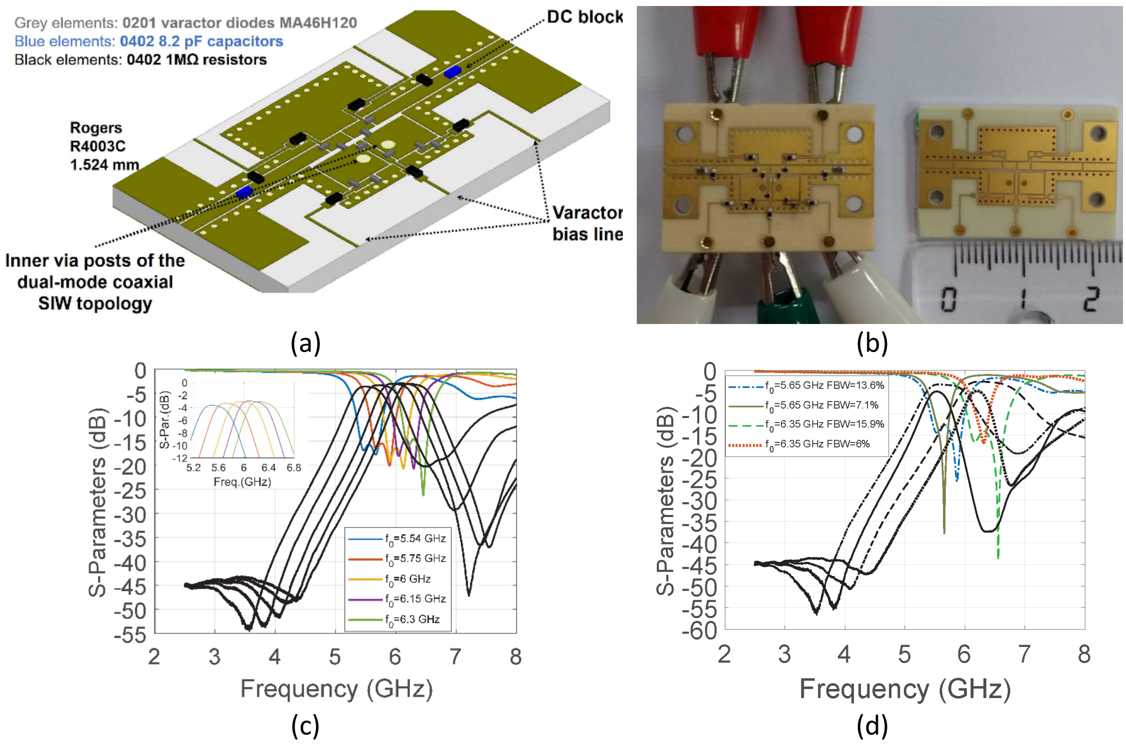


Figure 15: (a) Tunable doublet layout with SMD components, (b) a photograph of some filter prototypes, measured frequency responses (c) having constant BW3dB = 600 MHz. and (d) measured frequency responses when both BW and f_0 are controlled simultaneously [61].

Indeed, in [61], flip-chip varactor diodes are utilized as tuning elements for the frequency- and bandwidth-tuning of a C-band coaxial SIW doublet, which is also able to show two TZs at both sides of the passband thanks to a direct source-load bypass. As Figure 15 shows, the filter keeps a full planar integration since both SMD footprint pads and biasing lines have been embedded in the very same substrate, which means that such tunable device requires just a low-cost single-layer PCB manufacturing and a standard SMT assembly.

Figure 21:

The measured results shown in Figure 15(c)-(d) exhibit an improved filter response flexibility that allows the filtering responses to have either constant absolute BW or adjustable BW over a 13% tuning range at the C-band. Specifically, the IL and RL are always better than 5 dB and 15 dB (i.e. Q_u between 60 and 100 which are appealing values considering the working frequency band), respectively, over the whole TR. Finally, the filter is tuned from 5.65 GHz to 6.35 GHz with a FBW range of 6%-15.9% at 6.35 GHz and 7.1%-13.6% at 5.65 GHz.

A further development of the dual-mode topology includes loading a SIW cavity by four identical metallic posts, giving rise to two degenerate modes. By loading the ends of the inner posts with variable-reactance elements, these modes can be exploited to design quasi-elliptic-type filtering transfer functions with multiple levels of spectral adaptivity, which comprises center-frequency control, bandwidth tuning, and intrinsic RF switching-off [62]. To improve the overall performance, high-Q manually-adjustable thin-film trimmer capacitors have been used. As a result, the four-pole/four-TZ filter shown in Figure 16 exhibits center-frequency reconfigurability in the range 1.7-2.5 GHz and BW tunability between 58 MHz and 156 MHz, achieving that Q_u is in the range 120-150. The intrinsic switching-off mode presents a measured maximum in-band attenuation of 38 dB and minimum in-band attenuation of 21.5 dB within the range of 1.91-2.1 GHz.

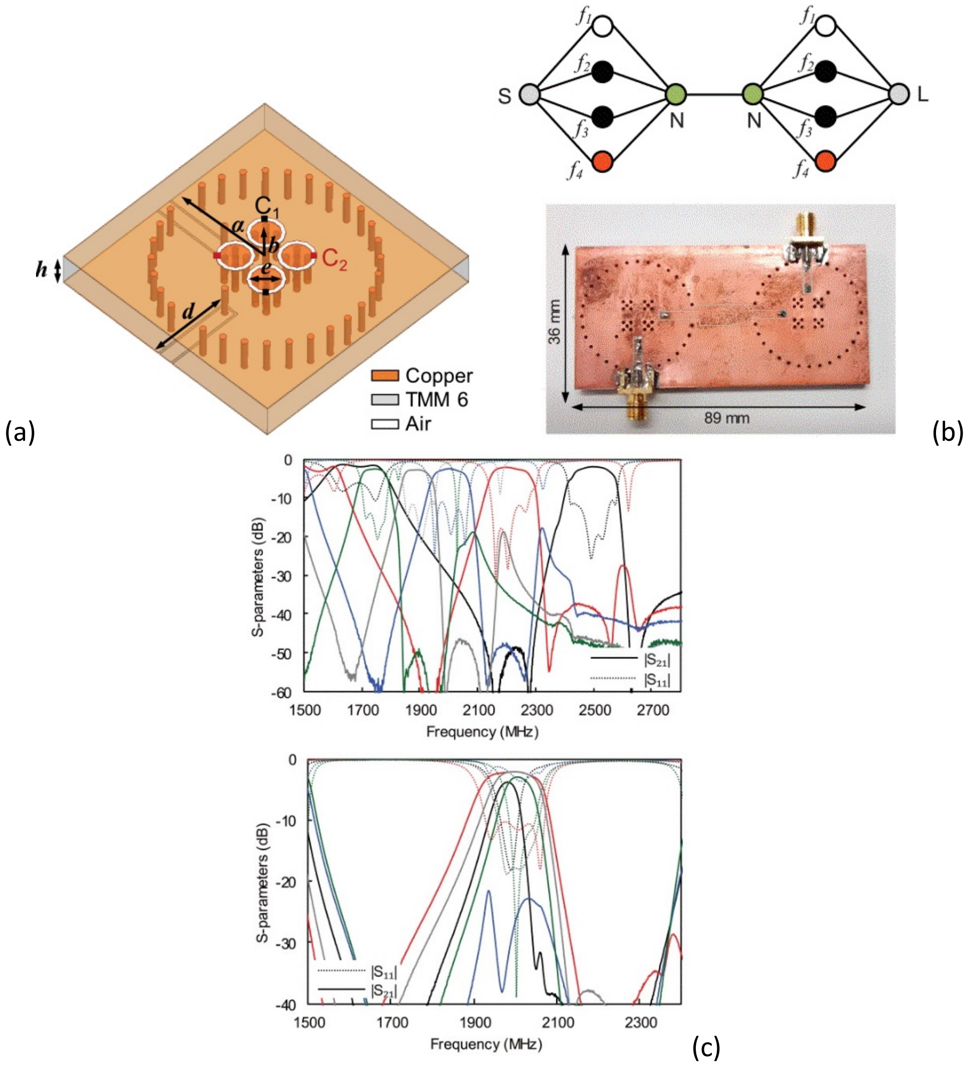
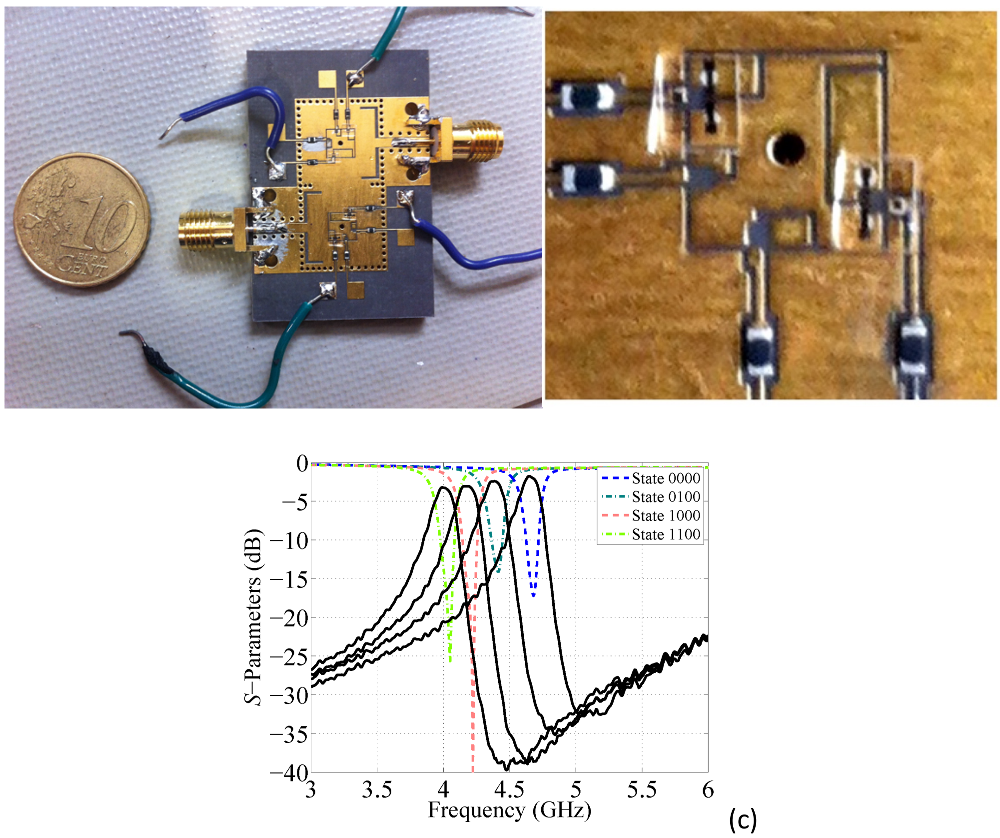


Figure 16: (a) Three-dimensional simulation model of the dual-mode- SIW-cavity- based BPF single-section. (b) Electric-field plots of the first four modes in the air gap between the metal posts and the upper wall of the cavity [62].

Finally, it is evident that a simple digital tuning based on switchable planar capacitances can be also applied to the coaxial SIW topology. Even if it was proved the improvement on the overall *Q*-factor enabled by this approach, just few examples of digitally tunable coaxial SIW filters have been proposed so far [63], [64], [65], showing tuning range in the range 15%-20% with Q_u easily higher than 100 at C-band working frequencies.

Such straightforward method to obtain digital frequency control is based on the introduction of embedded tuning capacitances on the loading capacitive patch of a resonator by creating electrically isolated elements. Then, PIN diodes as well as RF MEMS resistive switches are assembled between those integrated capacitors in order to perform the switching operability. Apart from the Q_u improvement, such tuning capacitance integration has several advantages in terms of limited BW variation versus frequency, higher tuning range, as well as a flexible positioning of the frequency states by adjusting separately each tuning capacitance size.

(a)  (b)

(c)

Figure 17: (a) Picture of the discretely tunable filter with four assembled RF MEMS resistive switches, (b) particular of the MEMS assembly on one resonator and (c) preliminary measurements for the four frequency states [64].

All of this is clearly visible looking at the results of the digitally tunable 2-pole C-band bandpass filter implemented in 3.175 mm-thick Rogers 5880 that has been proposed in [64], [66]. Four distinct bit capacitances are created in the capacitive patch, as shown in Figure 17, so that to provide up to 16 equally-spaced narrow-band frequency states requiring again a single-layer manufacturing. To perform the frequency tuning, RF MEMS resistive switches from the XLIM research group of the University of Limoges (France) have been mounted by means of a flip-chip assembly, thus avoiding the need for wire bonding. Such switches show a very good Q-factor above 200 @ 1 GHz and a reduced actuation voltage of 35 V.

Even if preliminary tests have been performed with only two RF MEMS switches assembled on each resonator, the achieved filter performance seems promising taking into account the working frequency band. Particularly, such filter can be discretely tuned from 4 GHz in 1100-state and 4.65 GHz in 0000-state (i.e. tuning range = 15%) providing four equi-spaced responses with constant FBW around 2.5% and a measured Q_u of 105-205. Power handling capabilities have been also tested at 4.7 GHz, corresponding to the off-state MEMS position, being the worst case in terms of self-actuation. The RF MEMS are able to handle up to 33.5 dBm before self-actuation occurs, which represents a remarkable advantange of such devices with respect to other tuning solution.

Table 1 A comparison of the different types of tuning mechanisms applied to the coaxial SIW topology described in this review.

Type	Tuning Element	Frequency	Tuning Range	Q-factor	Complexity	Integrability
Analog [52-54], [56-59]	Varactor Diode	UHF L-band S-band	>30%	50-200	Low	High
Analog [62]	Trimmer Capacitor	UHF L-band S-band	>30%	100-200	Very low	High
Analog [60]	Linear actuator + Tuning disk	UHF L-band S-band	>50%	200-300	High	Poor
Digital [63]	PIN Diode	Up to C-band	<30%	100-150	Medium	Low
Digital [64-66]	RF MEMS	Up to X-band	<30%	100-200	Medium	High

Conclusions

Low cost, high Q and easy integration are well recognized features of SIW technology. As we have reviewed and discussed in this paper, coaxial SIW technology enables to enhance some of the former advantages especially in terms of compactness, design flexibility and easy tunability. Starting from the basic resonator, different fixed filters with simple and advanced responses have been discussed, as well as some of the more recent building blocks for implementing dual-mode filters as well as singlets and doublets. Finally, some examples of tunable coaxial SIW filters have been discussed in detail and the main features of each approach are summarized in Table 1. As we have demonstrated, very relevant area reductions can be achieved by capacitively loading a conventional SIW cavity resonator while keeping good electrical performance and improving the out-of-band rejection. Moreover, a broad range of coupling structures and topologies can be easily implemented while keeping a very low cost and massive manufacturing process in single-layer PCB technology. Lastly, filter tunability can be easily performed due to the direct access to the capacitive part of the resonator, enabling to integrate different tuning elements such as trimmers, screws, varactors, p-p-i-n diodes and RF MEMS.

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