Coplanar High Impedance Wire on ferrite substrate : Application to isolators

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This article concerns a structure of high-impedance wires (HIW) based on ferrite substrate coplanar waveguide (CPW). Slotted planar stubs are etched on one or two ground planes of the CPW line. Several prototype measurements have been performed showing excellent performance to implement narrow-band microwave isolators. These HIW components can work at low bias-field giving an easier implementation in microwave devices.

***Index Terms*—High-impedance wire (HIW), defected ground structure (DGS), non-reciprocal component, isolator, coupled slotlines, ferrite substrate, metamaterial, metaline.**

I. Introduction

SEVERAL years ago, some authors [1] have developed metallic electromagnetic structures known as artificial magnetic conductor, or high impedance surface (HIS). The concept of high-impedance wire (HIW) proposed in [2] can be considered as 1-D metamaterial similar to high-impedance "meta-surfaces". HIS and HIW are usually fabricated with resonant structures periodically repeated in order to transform a conventional conductive ground plane into a high-impedance domain avoiding the propagation at resonance frequencies. They are generally modeled as small resonant circuits connected in series.

HISs forbid TM waves, and thus can be used as frequency selective surfaces, but can support TE waves in the form of leaky waves [3] which radiate or "leak" energy continuously when they travel through the surface, as in [26]–[28]. Similarly, in guided mode, several 1-D version of high impedance structures (HIW) were implemented in microstrip or coplanar waveguide (CPW) technology for various applications mainly associated with antenna technology [4]–[7] or low-pass filters [8], [9]. Unlike periodic structures named electromagnetic band gap (EBG) structures which operate at Bragg resonance frequency, high impedance structures, also known as defected ground plane structures operate below Bragg resonance frequency. For those structures, most of the applications presented in literature have a reciprocal behavior since the resonant elements are placed on dielectric substrates. But for applications such as circulators and isolators, magnetic materials and external magnetic bias field are needed to generate the non-reciprocal behavior. Several isolators based on ferromagnetic resonance, Faraday rotation and/or field displacement were developed in waveguide or in planar technology [10]–[20].

However, the miniaturization and the integration of non-reciprocal components still remain a major challenge for the microwave telecommunication industry. In planar technology, M. E. Hines [18] has designed an isolator using a microstrip line with a magnetized ferrite substrate. The non-reciprocal propagation is created by the field displacement phenomenon involving a non-uniform field distribution on each side of the metal strip. On one side is placed an absorber inducing high losses for only one propagation direction. This isolator works in a wide frequency band with low insertion losses.

To go to more compact and integrated components, coplanar structure seems to be more convenient. A narrow-band isolator has been performed using a non-symmetrical CPW on polarized ferrite substrate [10]. The operating frequency is about 10.6 GHz, the insertion losses are close to 1 dB and the isolation is superior to 20 dB. But a strong DC field is applied by a permanent magnet making it difficult to adjust the operating frequency.

In order to increase the operating frequency, other structures, combining HIW and ferrite substrate, have been developed in microstrip technology by J. Wu [19] and R. Guo [20]. Isolations of about 11.5-19 dB are reached in Ku band but the insertion losses (greater than 3.5 dB) are too high for an industrial exploitation, and a high magnetic bias field is required ($0.5 - 4$ kOe). To our knowledge, there is no previous work on structures combining magnetic materials and HIW in coplanar technology for microwave non reciprocal applications, that is the novelty of our design.

In this paper, we propose a novel non-reciprocal microwave component, using the concept of HIW. The structure is performed from a CPW printed on ferrite substrate (yttrium iron garnet). Under low magnetic bias field of about 70 Oe, prototypes show isolations better than 55 dB with a bandwidth at -20 dB of 80 MHz and insertion losses lower than 1 dB at the operating frequency of 13.6 GHz. The performance of the proposed structure is at the state of the art of commercial narrow-band isolators.

II. Structure and Design considerations

The structure under consideration is depicted in figure (1). The host transmission line is a 4 mm-length CPW, with geometrical parameters designed to achieve a characteristic impedance of 50 Ω and low radiation losses: signal line $S = 212$ μm, slot width $W = 212$ μm, thickness of YIG

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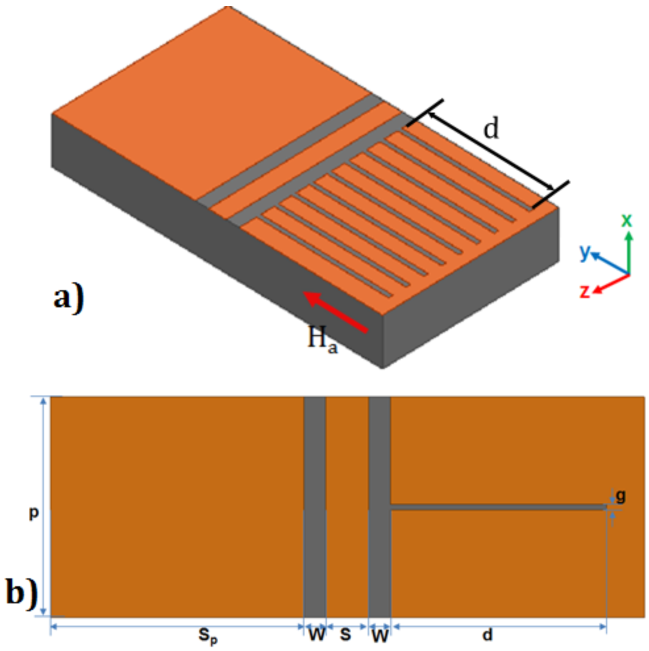

Fig. 1. Design of the structure: a) non-symmetrical configuration of the component, b) geometrical parameters of the initial unit cell: $S = 212$ μm, $W = 212$ μm, $W_p = 5.1$ mm, slot width $g = 40$ μm, slot length $d = 5$ mm, period $p = 400$ μm.

substrate $h = 1000$ μm, thickness of copper layer, $e = 3$ μm and width of ground plane $S_p = 5.1$ mm. Two configurations are possible for the proposed structure: in the symmetrical configuration the slots are engraved on both ground planes while in non-symmetrical configuration they are etched on one ground (figure 1.a). According to the transmission line model of HIW presented in [2], the impedance provided by the single series stub of the non-symmetrical configuration is higher than that provided by the two stubs connected in parallel of the symmetrical one. In this paper we are interested in the non-symmetrical version since we are looking for the higher value of the impedance of the stubs in order to produce stronger resonances.

A. Design of the HIW structure

HIWs are defined by three design parameters [2]: the period of the unit cell p, the length of the stub d, and the width of the slot g. A theoretical analysis of this kind of structure and the design procedure have been widely developed in [24] and [25]. Following this procedure HIWs were designed for applications in microwave in previous works [26]–[28]. Due to the complexity of the structures, which involves strong coupling between the slots, as anisotropic nature of the substrate, quasi-static equations are not enough accurate, so electromagnetic simulations using finite elements softwares are usually needed to perform the accurate values of the design parameters.
The first step is to design the unit cell of the resonant circuit shown in (figure 1.b). The equations giving the approximate values of the geometrical parameters of the unit cell can be found in [24] and [25]: the width of the slot, g, was set to 40 μm much smaller than λ_g, so that the slot can be treated as a series stub, and the length of slot, $d = 5000$ μm, was tailored to achieve the first resonance frequency f_r at 5 GHz corresponding to $\lambda_g/4$ on a YIG substrate with a dielectric constant $\epsilon_r = 15.3$. According to the equation (1) [2] which approximately gives the dimensional resonant frequencies, the second resonance ($m = 1$) will occur in Ku-band, around 15 GHz:

$$f_r \approx (2m+1)\frac{c}{4d\sqrt{\varepsilon_{eff}\mu_{eff}}} \quad (1)$$

with $m = 0, 1, 2, 3, \ldots$, c the speed of light in vacuum, ϵ_{eff} and μ_{eff} respectively the effective permittivity and permeability on the slotline stubs.
To remain in the theoretical case of HIW and a metamaterial viewpoint, the length of the unit cell, known as the period p of the HIW, must be smaller than length of slot d. On the other hand, p is usually chosen as large as possible to avoid strong coupling between the slots and simplify the analysis of the structure. However, this constraint leads to large structures, and we wish to make a device as compact as possible. Anyway, coupling between slotlines on an axially magnetized ferrite as been treated in [29], leading to theoretically describe the coupling in a HIW. So, in our case the period p was set to 400 μm to ensure compactness of the structure. The global resonance frequency of the unit cell is of type LC, with the inductive part brought by the resonating stub, and the capacitive one by the parasitic capacitance between the coupled stubs. To adjust the operating frequency of the isolator one has thus the choice to modulate the effect of the parasitic capacitance, which increase as p shortens, by lengthen the size of the isolated stub. To fix the number of stubs, following the procedure in [7], we add unit cells to the structure until the performance criteria are reached. With the given values of p, d and g, 9 stubs and static magnetic field $H_a = 70$ Oe, by simulation, two resonance frequencies are obtained at 13.7 GHz for the forward propagation and 15.2 GHz for the backward propagation (Fig. 4).

B. Magnetic field distribution on the defected ground plane

As mentioned in [19] the analytic close form distribution equations for the defected ground plane structure can be complicated. To analyze the magnetic field distribution, the Ansoft HFSS full wave simulator has been used. For the original ground plane, the current flows mainly on the edge along z axis and the magnetic field distribution is dominated by the components H_x and H_y. Whereas for the DGS the presence of slots forces the current flowing around the new edges as schematically presented in figure 2.a. This leads to rotating magnetic field with components H_x and H_z [19], [20], [29]. When an in-plane DC magnetic bias is applied along y-axis the magnetic field around the coupled slots is characterized by Right-Handed Circular Polarization (RHCP) for forward input (figure 2.b) and a Left-Handed Circular Polarization (LHCP) for backward input (figure 2.c). Then the propagation constant in the stubs takes two different values, depending on the direction of propagation. According to [29], even and odd modes in the parallel slots are coupled, and RHCP and LHCP propagation constants can be express as a

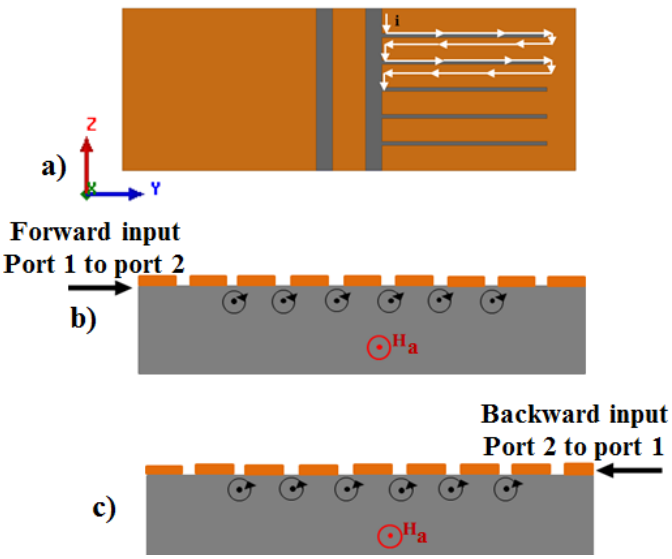

Fig. 2. Configuration of magnetic field on the defected ground plane: a) direction of current, b) RHCP for forward transmission, b) LHCP for backward transmission.

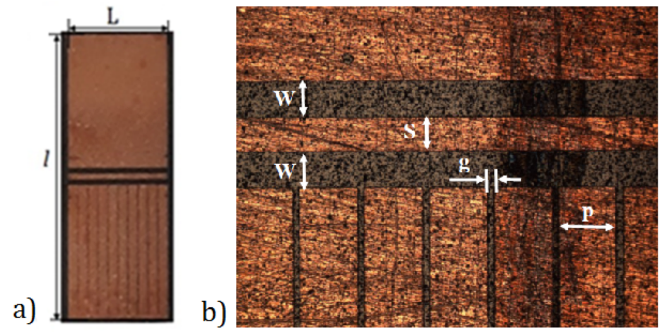

Fig. 3. Prototype of coplanar HIW isolator: a) photograph of prototype with 9 slots, l = 11 mm and L = 4 mm, b) geometrical parameters measured with optical microscope: S= 212 μm, W=220 μm, slot width g= 40 μm, slot length d= 5 mm and period p = 400 μm.

TABLE I
MAGNETIC PROPERTIES OF YIG SUBSTRATES.

Reference	μ_0Ms (mT) ±5%	ΔH (A/m) ±5%	ϵ_r ±5%	tanδ	T_c (°C) ±5%	H_C (A/m) ±5%
YIG 101	182	1590	15.4	$< 2.10^{-4}$	280	< 100

function of β_{even} and β_{odd} and a coupling coefficient C caused by the gyrotropy of the ferrite as shown in equation (2) [29].

$$\beta_{RHCP}, \beta_{LHCP} = \left(\frac{\beta_{even} + \beta_{odd}}{2}\right) \pm \sqrt{\Delta\beta^2 + |C^2|} \quad (2)$$

with $\Delta\beta = \frac{\beta_{even} - \beta_{odd}}{2}$ and C depending on an integral of H fields over the cross-section of the substrate [29].
This behavior of coupled slots on an axially magnetized ferrite slab leads to non reciprocal resonance frequencies of the HIW, as shown in figure 4 and theoretically detailed in [30], and quantitatively explains the difference between the two resonance frequencies. Using the coupled-mode theory, the authors develop in the reference [30] the theory on the coupling in symmetrical structures when the ferrite is magnetized in the Faraday model similar to the isolator structure under consideration.
Finally, as demonstrated in previous works, the metallization thickness effect produces a slight non-reciprocity in CPWs with ferrites magnetized in the plane of the slab because of the difference of the field distribution between the backward and forward waves [23]. The contribution of this second source added to the effects of circular polarizations thus leads to a non-reciprocal structure.

III. EXPERIMENTAL RESULT AND DISCUSSIONS

A. Manufacture of the structure

The fabricated prototypes are shown in figure 3. The 1000 μm-thick substrate is obtained by thinning a bulk YIG (Temex Ceramics) whose properties are given in table I. The surface of the substrate was then grinded and polished in order to reduce his roughness and thus improve the homogeneity of the copper layer that will be deposited by RF sputtering. The thickness of the copper ground planes and signal line is around 3 μm. The coplanar transmission line and the stub network on the ground plane are made by conventional photolithography processes. Several prototypes were then characterized under low-bias DC magnetic field with a vector network analyzer (A37397 Anritsu) and a coplanar probes system. The magnetic bias field (70 Oe) can be produced either by an electromagnet or a permanent magnet, and is applied along the stubs during the S-parameters measurement. According to the applied field configuration (transverse polarization), the demagnetization field is negligible and for low bias-field (over 60 Oe) the YIG can be considered in the saturated state, since the YIG is magnetized in the plane, we can say that it is near saturation state in agreement with the measurement made in [31]. That constitutes a great advantage compared to other structures requiring high bias field.

B. Results and discussions

For the applied magnetic bias field of about H_a= 70 Oe, the measured and simulated S-parameters shown in figure 4 confirm the non-reciprocal propagation on the structure. For the forward propagation, the isolation is better than 55 dB at 13.6 GHz, with the insertion losses lower than 1 dB (around 0.8 dB) and the reflection parameters are lower than -20 dB, which proves that the missing energy is not reflected back to the port 1. However, the narrow isolation peak leads to small bandwidth of 80 MHz at -20 dB. In the reverse direction, the isolation peak appears around 15.2 GHz but does not reach 10 dB. The return and insertion losses are respectively about 8 and 5 dB with, notwithstanding, a wider bandwidth. Good agreement between simulation and measurement is observed for the resonance frequencies, with a slight shift of 0.1 GHz. However, the performance related to amplitudes is better in measurement than in simulation. This difference in amplitudes can be explained by the magnetization conditions

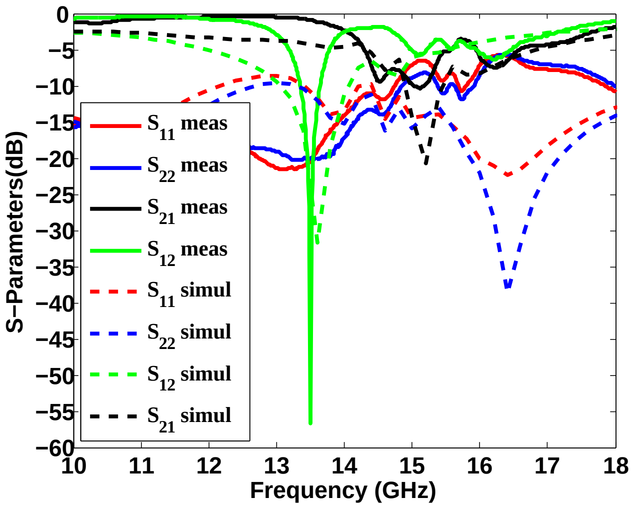

Fig. 4. S-parameters of the structure with 9 stubs under a magnetic bias field H_a = 70 Oe. Solid curve: measurement, dashed curve: simulation.

of the ferrite. Indeed, the simulations consider that the internal magnetic field is constant and the material saturated in all the material, whereas in measurement the internal DC field is inhomogeneous [32]. For better accuracy, the static magnetization state of the slab could be first simulated and integrated in the dynamic simulation.

This structure can be used as simple narrow-band isolator showing high performance. A comparison among different isolators given in table II confirms the performance of our design. The power budget shows strong losses at resonance frequencies. As operating frequencies are far from the gyroresonance of the ferrite, the main origin of losses can not be magnetic. A prototype fabricated on alumina showed the same behavior as the non magnetized HIW-CPW on ferrite, as shown in figure 5, which indicates that losses are coming from the metallic structure. Losses could be due to leaky waves propagating on the DGS. Under bias field, HFSS simulations show that the structure is radiating, as it can be seen on figure 6. In this simulation the HIW-CPW is in the YZ plane, with signal flowing along z-axis. The radiation pattern of the structure has strong similarities with that of a leaky wave antenna. Further measurements in an anechoic room should confirm these simulation results.

IV. Conclusion

In summary, a non-reciprocal microwave device with a coplanar defected ground transmission line based on a ferrite substrate has been designed, fabricated, and measured under DC magnetic field. With 70 Oe magnetic bias field, prototypes show good performance to implement a narrow-band tunable isolator: isolation better than 55 dB with a bandwidth at -20 dB of 80 MHz and insertion loss of about 0.8 dB at 13.6 GHz. Since a low bias magnetic field is required, the use of a small magnet or a small coil is therefore possible. Accordingly, the operating frequency can be controlled by switching the direction of the bias-field, allowing the concept of switched

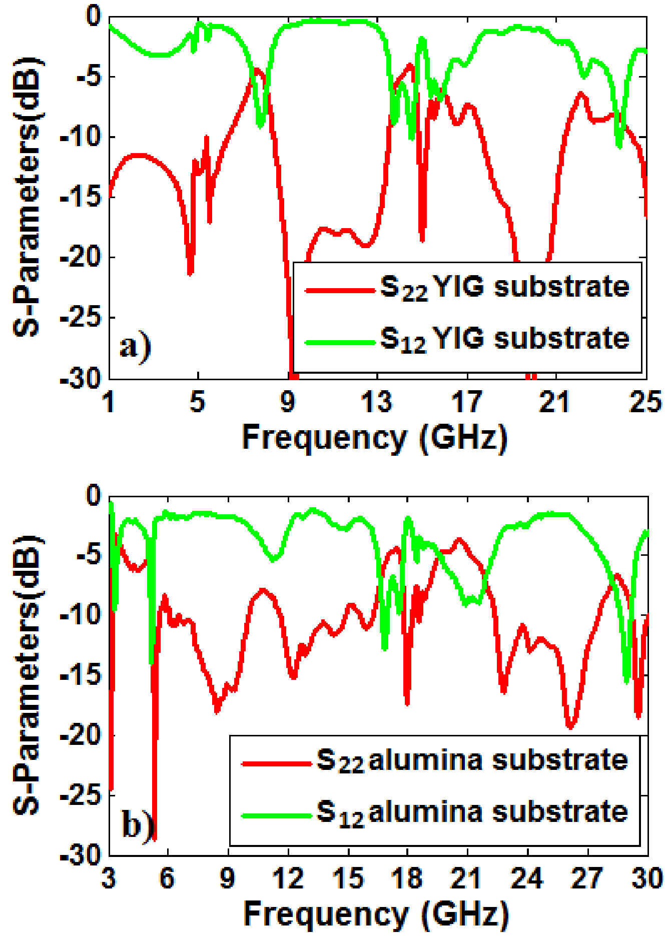

Fig. 5. Measured S-parameters of the HIW-CPW: a) on demagnetized YIG, b) on alumina substrate.

dual passive filters. At the operating frequency the missing energy seems to be radiated as in leaky waves antennas, giving the possibility to put absorbents outside the structure. Other experimental works are under consideration to widen the bandwidth of the isolator and make it tunable, as well as to quantify the radiated energy, which may allow the use of the structure as an dual-band antenna.

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TABLE II
COMPARISON OF THE PERFORMANCE OF OUR STRUCTURE AND OTHERS DESIGNS FOUND IN THE RECENT LITERATURE.

Reference work, year	Operating frequency (GHz)	Isolation (dB)	Insertion loss (dB)	Applied field (Oe)	Size (mm^2)	Ferrite	Technology
This work	13.6	57	0.8	70	11x4	YIG	HIW-CPW
[20], 2018	17.57	11.6	5.78	3500	-	NiZn	HIW-Microstrip
[16], 2014	9.81-10.24	30-55	1.1-2.2	2630	22.6×8	YIG	SIW
[15], 2013	13.5-16	17-30	2-5	4000	≈39.2×12	YIG	SIW
[19], 2012	13.5	19.3	3.5	4000	≈22×9	YIG	HIW-Microstrip
[10], 2011	10.6	17	1	2260	10×4	YIG	CPW

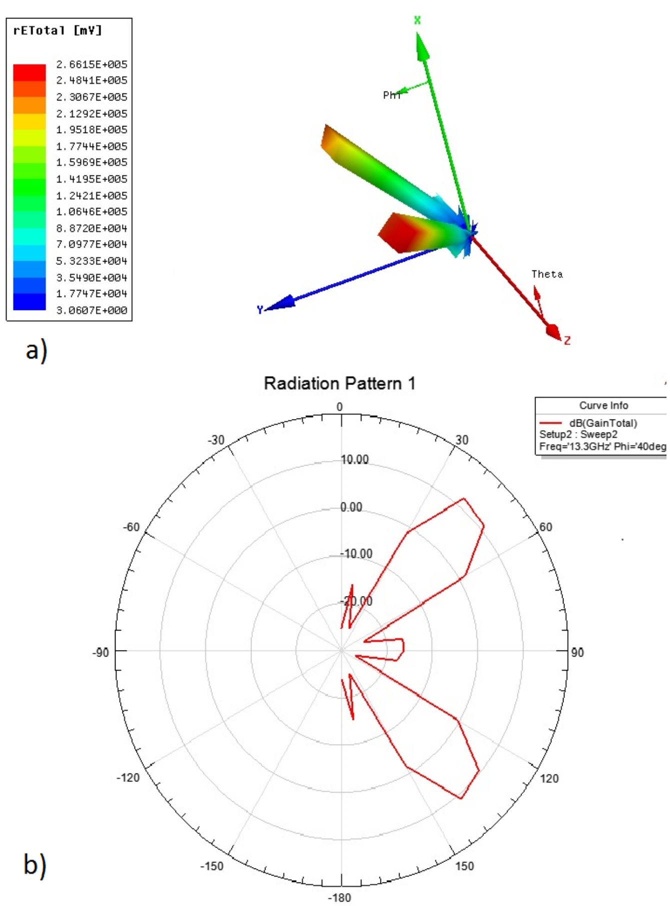

Fig. 6. Radiation pattern of the polarized HIW-CPW at the resonance frequency. a) 3D polar plot of total E field. b) Total gain in dB of the radiation pattern in the plane $\phi = 40^o$.

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