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Abstract: This work presents a new unit cell electromagnetic bandgap (EBG) design based on HoneyComb geometry (HCPBG). The new HCPBG takes a uniplanar geometry (UCPBG—uniplanar compact PBG) as a reference and follows similar design methods for defining geometric parameters. The new structure's advantages consist of reduced occupied printed circuit board area and flexible rejection band properties. In addition, rotation and slight geometry modification in the HCPBG cell allow changing the profile of the attenuation frequency range. This paper also presents a reconfigurable unit cell HCPBG filter strategy, for which the resonance center frequency is shifted by changing the gap capacitance with the assistance of varactor diodes. The HCPBG filter and reconfiguration behavior is demonstrated through electromagnetic (EM) simulations over the FR1 band of the 5G communication network. Intelligent communication systems can use the reconfiguration feature to select the optimal operating frequency for maximum attenuation of unwanted or interfering signals, such as harmonics or intermodulation products.

Keywords: electromagnetic bandgap; photonic bandgap; electromagnetic compatibility; interference; filtering; HCPBG (HoneyComb PBG)

1. Introduction

Photonic bandgaps (PBGs) are periodic structures that introduce material changes in the waveguide or printed circuit board (PCB), such as holes, patterns, or dielectric rods. They are also known as photonic crystals and are based on Electromagnetic Band Gap (EBG) properties. The geometric modification in the medium produces forbidden frequency bands for propagating waves. In other words, it works as a wave filter with high impedance at desired frequency bands. As a result, the electromagnetic waves in a PBG material are hindered due to the periodic discontinuity, which is equivalent to a photonic crystal in the light domain. They were first reported for structures at optical wavelengths in 1987 [1,2]. Since the first PBG structure publication, designers created many shapes, geometries, and materials for various applications. As an example, we can cite: microwave filters, electromagnetic compatibility (EMC) improvement [3–5], antenna beam steering [6], compact antenna arrays [7], antenna gain, efficiency, and bandwidth improvement [8–10], and beam tilting of 5G antenna arrays [11].

PBG structures have the characteristic of phase control of plane waves enabling suppression of surface waves and higher-order harmonics. It is also an exciting tool for those looking beyond 5G, where the advent of electromagnetic components can shape how they interact with the propagation environment. It is a case for reconfigurable intelligent surface (RIS), a two-dimensional surface of engineered material whose properties are reconfigurable rather than static [8,9,12]. For example, the RIS device can control scattering,



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absorption, reflection, and diffraction properties by software according to how the environment changes over time. In principle, the RIS can form a beam and synthesize the scattering behavior of an arbitrarily shaped surface of the same size. For example, it can create a superposition of multiple beams or act as a diffuse scatterer [13]. Regarding traditional communication systems, the critical difference between a RIS and the common notion of an antenna array in 5G is that a RIS is neither part of the transmitter nor the receiver. However, it is a controllable part of the wireless propagation environment. The system-level role of a RIS is to influence the propagation of the wireless signals sent by other devices without generating its signals [14].

On the scope of the current investigation, we focus on interference filtering, which is a critical problem to be addressed as the number of wireless devices increases every year. It is of great concern that long-term interfering issues could arise due to the increasing number of heterogeneous devices. Aiming on that, it is a necessity to implement such solutions as a filtering option. For example, cable-stayed power line towers monitored by wireless sensors operating at unlicensed spectrum deliver data of utmost importance to the substations. The data help predict the risks of collapse events, and wireless sensor communication may be harmed by adjacent heterogeneous systems interference. In this way, one can use PBG as a tool to contend against interference coming from a crowded environment in several ways. Firstly, the current downscaling of devices makes structure sizes a critical issue. As technologies occupy higher spectrum frequencies, PBG is a viable tool to be adopted since the stopband frequency is proportional to half of the guided wavelength [15]. Secondly, planar PBGs like UCPBG do not need vias or unique materials suitable for standard PCB designs with no extra cost. The UCPBG works as a stopband filter due to the structure of metal etched in slots on the ground plane connected by narrow lines to form a distributed LC network [16,17].

In this paper, we present the HCPBG unit cell, a novel geometry for EBG structures that introduces new features when compared to that of the traditional UCPBG [9]. For instance, it further decreases the consumption area compared to that of the UCPBG equivalent due to the proposed hexagonal geometry. Secondly, the hexagon can keep the same perimeter as the side of a square, where the microstrip line crosses the PBG structure while occupying 65% of the area. Besides, the HCPBG unit cell can be configured with different traces and orientations, adding asymmetry to the design and producing different filtering characteristics, as simulations and measurements demonstrate. As a result of this study, we can highlight the following HCPBG advantages:

- Introduction of a new EBG geometrical structure for RF filtering applications;
- A more effective area usage for PCBs due to the hexagon geometry;
- Multiple rejection frequency bands and bandwidths are achieved depending on the orientation or number of traces;
- Addition or suppression of traces allow shifting the first resonance frequency;
- Suppression of high-frequency rejection band by changing the orientation;
- Wide bandwidths are obtained by changing the number of traces or orientation;
- Active reconfiguration of the first resonance frequency.

As a second novel feature, we also present a reconfiguration method to modify the center frequency of the rejection band by controlling the gap capacitance through a varactor diode. The reverse voltage of the varactor diode changes its capacitance, allowing a control system to shift filtering center frequency and bandwidth range.

Although the technique of using active discrete components to reconfigure different types of PBG structures was studied before, the reconfiguration of the UCPBG type structures, as presented in this work, is not yet reported in the literature.

Different applications of reconfigurable EBG structures were investigated over the years, such as antenna array beam-steering at 6 GHz using metal tape [6], change of patch antenna polarization for navigation systems with the use of varactor diodes [18], modification of antenna frequency and radiation pattern for WiFi/WiMAX using PIN

diode [19], cavity resonator at 10 GHz range with mechanical switching reconfiguration [20], and reconfigurable 3D MEMS filter for optical applications [21].

The HCPBG is a planar structure, and no unique material, vias, or fabrication process is needed to produce such structures. This is an advantage over tunable bandpass filters based on microelectromechanical systems (MEMS) [22]. On the other hand, planar reconfigurable PBG filters, as described in [23–25], have the advantage of large bandwidth and dynamic range. However, to achieve a frequency range of a few GHz, they show an increased area consumption and circuit complexity compared to that of the proposed filter.

This paper organizes as follows: Section 2 refers to the presented HCPBG model. Section 3 presents the simulation results comparing the classic UCPBG to the HCPBG and multiple geometries through transmission loss of a microstrip line over one single cell; it also presents the effect of different orientations. Section 4 shows the measurement setup and transmission loss results for the fabricated samples. Section 5 describes the design of the reconfigurable HCPBG filter and presents the simulation and measurement results. Finally, the discussion is closed in the Section 6.

2. Design of HCPBG

The method used to evaluate the UCPBG and HCPBG unit cells consists of a PCB having a microstrip line that crosses the entire top layer and one single cell at the bottom layer underneath the microstrip line. Figure 1 illustrates a diagram used in this work for simulation and measurements. By supposing an interfering signal impinging the stripline, the objective here is to calculate a resonance frequency for the structure to filter out the interference.



Figure 1. Reference diagram for simulations and measurements.

Figure 2 shows (a) the design reference UCPBG and (b) the new proposed HCPBG unit cells on microstrip substrates. The dark area is the metallic ground layer, while the white area represents the removal of the metal (air gaps). The geometry design parameters for both UCPBG and HCPBG are the gap size g_1 , g_2 , the total structure size a, and the trace dimensions where t_l is the length and t_w is the width. The gap sizes define the capacitance, while the trace length and width define the inductance. For example, reducing the gap size g_1 , g_2 increases the capacitance and, henceforth, resonance moves to lower frequencies. On the other hand, a longer length t_l will produce a higher inductance, consequently reducing the rejection band frequency. Similarly, the width t_w affects the inductance and, henceforth, the resonance frequency.



Figure 2. PBG structure designs—(**a**) Uniplanar Geometry PBG (UCPBG) and (**b**) HoneyComb PBG (HCPBG).

The lumped capacitors and inductors introduced by gaps and traces form a parallel LC network as described in [2]. The determination of a stopband frequency for a single PBG structure is roughly determined by [5]

$$f_o = \frac{1}{2\pi\sqrt{LC}},\tag{1}$$

where *L* is the inductance of the trace and *C* is the capacitance of the gaps connecting the PBG structure to the ground. The method for calculating *C* follows the model of two metal sheets coplanar capacitance on a PCB [26], which can be calculated as

$$C = \begin{cases} \frac{l\varepsilon_{r}}{377\pi c} \ln \left[-\frac{2\left(1+\sqrt[4]{1-\frac{g_{1}^{2}}{(2w+g_{1})^{2}}}\right)}{-1+\sqrt[4]{1-\frac{g_{1}^{2}}{(2w+g_{1})^{2}}}} \right], 0 < \frac{g_{1}}{g_{1}+2w} \leqslant \frac{1}{\sqrt{2}}, \\ \frac{l\varepsilon_{r}}{120c \ln \left[-\frac{2\left(1+\sqrt{\frac{g_{1}}{2w+g_{1}}}\right)}{-1+\sqrt{\frac{g_{1}}{2w+g_{1}}}} \right]}, \frac{1}{\sqrt{2}} < \frac{g_{1}}{g_{1}+2w} \leqslant 1, \end{cases}$$
(2)

where *w* is the size of the metal plates of the considered capacitor model, $\varepsilon_r = 4.3$ (FR4) is the substrate dielectric constant, *c* = 299,792,458 m/s is the vacuum speed of light, and $l = a - t_w - g_1 - g_2$ is the gap length. It is also convenient to make $g_2 = t_w$ to minimize the effect on the inductance *L*.

One can calculate L as [27]

$$L = \rho(t_w) \times t_l, \tag{3}$$

where

$$\rho(t_w) = \begin{cases} 60 \frac{\ln\left[\frac{8n}{t_w} + \frac{t_w}{4h}\right]}{c}, & \text{if } \frac{t_w}{h} \le 1\\ \frac{180h\pi}{c(h(2.0895 + \ln\left[1.444 + \frac{t_w}{h}\right]) + 1.5t_w)}, & \text{otherwise'} \end{cases}$$
(4)

and h is the substrate thickness.

The calculated values of f_0 , C, and L define the dimensions considered in the EM simulations. It is also a starting point where the designer must proceed a fine-tune to achieve a desired resonance center frequency. Also, a second method one can use is scaling based on a predesigned structure. The rejection band moves to lower frequencies as the structure is scaled up and vice-versa. Furthermore, the attenuation further improves if the PBG structure forms a lattice. As the designer works at higher frequency ranges, UCPBG achieves smaller geometric sizes, improving its applicability. In this case, metal slots are etched in the ground plane connected by narrow lines to form a distributed LC network [17]. This method fits well for standard PCB designs because it is planar, countering the need for vias or unique materials, saving costs. Concerning the hexagonal geometry, the reduction

in the PCB-occupied area is the most attractive design feature. It can keep the same gap perimeter on the side of a square while occupying 65% of the area.

The current work investigates six variations of HCPBG geometries. The desired f_0 is achieved by simply changing the number of traces, introducing different gap lengths, rotation of PBG structure relative to the microstrip line, and geometric scaling. Table 1 describes and illustrates the variations proposed for simulations and measurements. Each line of the table constitutes the structure case, the number of traces, rotation, and the PCB bottom layer PBG structure relative to the microstrip line represented as dashed contours. The cases of Table 1 (a) and (b) explores the asymmetry in the HCPBG structure by suppressing some of the traces, which changes the LC elements from the basic geometry presented in Figure 2b.

The capacitance of the gap and the trace inductance closer to the point where the transmission line crosses the PBG have a more substantial influence on the resonance frequency. Adding asymmetry changes the values of the LC and the influence over the resonances.

The cases (a) and (b) are the three traces and show asymmetries for both gaps and traces close to the microstrip line. For cases (c) and (d), we can notice that traces and gaps are symmetrically apart from the microstrip line, and in case (e), there are two traces parallel to the microstrip line due to the rotation of 30°, for which it is expected to have similar results to the UCPBG. Finally, the case (f) is the fundamental geometry presented in Figure 2b.

Label	Traces	Rotation	Structure
(a)	3	0°	
(b)	3	30°	
(c)	4	0°	
(d)	4	90°	
(e)	4	30°	
(f)	6	0°	

Table 1. Six variations of HCPBG geometries considering rotations and suppressed traces.

3. Simulation of HCPBG

The simulation setup consists of a reference plane PCB (without PBG on the ground layer), a UCPBG, and six HCPBG combinations shown in Table 1. Six out of seven simulations on HCPBG structures have dimensions defined in Table 2 case (a). The seventh simulation has dimensions defined in Table 2 case (b), where g_2 is the only different parameter. We calculated the PCB's frequency profile using CST Studio[®] [28], an electromagnetic field simulation software.

In the CST Studio simulation, we set the time domain solver, which employs finite integration technique (FIT) [29]. The time domain solver has the ability to handle large and

complex structures, and it allows for memory efficient computation. We set the accuracy to -40 dB and hexahedral mesh type. For the mesh properties, we defined 12 cells per wavelength near/far from model and 35 cells of fraction minimum cell near model.

The PCB is an FR-4 two-layer of 90 mm length, 60 mm width, and 1.6 mm thickness. The microstrip line is 50 Ω , 3 mm width and 90 mm length on the top layer. The ground plane is placed on the bottom side of the PCB, where the HCPBG and UCPBG structures are located, as shown in Figure 1. The simulations consider the dimensions of Table 2 case (a) where L = 3.26 nH, C = 1.38 pF and $f_0 = 2.37$ GHz when using Equations (2) and (3). The nomenclature to name the HCPBG structures is defined as HCPBG-xx-yy, where "xx" represents the number of traces and "yy" is the orientation relative to the microstrip line.

 Table 2. Dimensions used for UCPBG and HCPBG designs.

Case	а	<i>g</i> 1	<i>g</i> ₂	t_w	t_l
(a)	28.5 mm	1.5 mm	1.5 mm	7.5 mm	1.5 mm
	$0.23\lambda_{2.37~\mathrm{GHz}}$	$0.012\lambda_{2.37~\mathrm{GHz}}$	$0.012\lambda_{2.37~\mathrm{GHz}}$	$0.06\lambda_{2.37~\mathrm{GHz}}$	$0.012\lambda_{2.37~\mathrm{GHz}}$
(b)	28.5 mm	1.5 mm	0.5 mm	7.5 mm	1.5 mm
	$0.23\lambda_{2.37~ m GHz}$	$0.012\lambda_{2.37~\mathrm{GHz}}$	$0.04\lambda_{2.37~\mathrm{GHz}}$	$0.06\lambda_{2.37~\mathrm{GHz}}$	$0.012\lambda_{2.37~ m GHz}$

The most significant characteristic of HCPBG unit cell is that multiple resonance profiles emerge due to rotation and trace suppression. Figure 3 shows the transmission loss profile for UCPBG and HCPBG Table 1 case (e). Both have similar results up to 3.3 GHz, with the first resonance at 2.36 GHz and similar bandwidth (BW); BW = 136 MHz at -10 dB. This case presents a second strong resonance at 4.13 GHz with BW = 100 MHz and the third one at 5.38 GHz, but with an attenuation lower than the -10 dB. This third resonance coincides with the second resonance of the UCPBG structure but with lower attenuation.



Figure 3. Simulation-transmission loss for UCPBG and HCPBG-4t-o3.

Figures 4 highlights the effect of HCPBG rotation for Table 1 (c) and (d) cases. HCPBG and UCPBG present a similar filtering profile observed in the previous simulation. Notice that the main differences are due to the rotation of HCPBG-4t-o1 and HCPBG-4t-o2 (Table 1 case (c) and (d), respectively). The HCPBG-4t-o1 has two strong attenuated bands, one at 2.17 GHz (BW = 167 MHz) and the second at 3.61 GHz, with a wide bandwidth (BW = 982 MHz). In HCPBG-4t-o2, the first attenuation band has an extensive bandwidth at 2.25 GHz (BW = 1.4 GHz), and the second resonance locates at 4.14 GHz (BW = 260 MHz).



Figure 4. Simulation—transmission loss for HCPBG with 4 traces—orientations o1 and o2.

We also simulated HCPBG with three traces and six traces, keeping the same dimensions to verify the attenuation band profile's behavior. Figure 5 shows the transmission loss for the HCPBG with 3 traces and two different orientations, as indicated in Table 1 by cases (a) and (b). Both have similar filtering profiles up to 4.5 GHz, with a first attenuation around 2 GHz, and with BW of 645 MHz for HCPBG-3t-01 and BW of 380 MHz HCPBG-3t-02. The second attenuation band is close to 3 GHz for both. The main difference is that HCPBG-3t-01 produces a third strong wideband attenuation at 5.75 MHz (BW = 1670 MHz). The LC network seen by the microstrip line depends on the HCPBG orientation, and it seems to have more impact when the HCPBG traces are more aligned with the microstrip line.



Figure 5. Simulation—transmission loss for HCPBG with 3 traces—orientations o1 and o2.

Figure 6 shows the transmission loss simulation results for the 6-trace HCPBG, Table 1 case (f), with outer gaps of 1.5 and 0.5 mm. A smaller gap results in a greater capacitance, leading to lower rejection band frequencies. The HCPBG-6t with a gap of 1.5 mm has a significant bandwidth response if we use a criterion of -5.8 dB (BW = 3.16 GHz). This structure could be helpful in the suppression of broadband noise.



Figure 6. Simulation—transmission loss for HCPBG with 6 traces—gap distances 1.5 and 0.5 mm.

4. Measurements of HCPBG

We also verified the analytical guidelines and simulation results performing measurements in an actual PCB. An LPKF machine milled the 2-layer FR-4 samples of UCPBG and HCPBG structures. Figure 7 shows the bottom side of two fabricated samples, UCPBG (to the left) and the HCPBG-4t-o3 (to the right). The HCPBG-3t-o1, HCPBG-3t-o2, HCPBG-6t (g = 1.5 mm), and the reference board (ground only) were also fabricated. The sample dimensions and substrates are the same used for the simulations.



Figure 7. Fabricated UCPBG (left) and HCPBG with 4 traces, orientation 3 (right)-PCB bottom view.

The measurement setup, shown in Figure 8, is composed of the Keysight Field Fox RF Analyzer N9914B, 2 RF cables, N-SMA adapters, and the samples. The analyzer is set to network analyzer mode and calibrated from 30 kHz to 6.5 GHz.

The measured logarithmic magnitude of S21 parameter for the UCPBG and HCPBG-4t-o3 can be seen in Figure 9. When comparing the results in Figure 3, we can observe that the simulation curves' profiles are very close to the measurement results. However, the resonance frequencies showed a slight discrepancy, with the UCPBG's case being the more evident one. This discrepancy is probably due to fabrication process variations, such as gap width, depth, and substrate dielectric constant variation. Nonetheless, the simulation results showed good accuracy that can be observed in all measurements. The UCPBG's measured first resonance frequency is at 2.18 GHz (BW = 160 MHz), and the second one at 5.4 GHz. For the HCPBG-4t-o3, the resonance frequencies are 2.4 MHz (BW = 160 MHz), 4.26 GHz (BW = 130 MHz), and 5.56 GHz. The bandwidth and maximum attenuation are



also very similar in both measurement and simulation. The reference line represents the measured S21 for the transmission line with a solid ground plane (no PBG).

Figure 8. Measurement Setup for Transmission Loss.



Figure 9. Measurement—transmission loss for HCPBG with 4 traces, orientation 3 vs. UCPBG.

The graphics of Figure 10 show the measurement results for the three trace geometries: HCPBG-3t-o1 and HCPBG-3t-o2. The simulation results are in good accordance with the measurements concerning curve profile, rejection band center frequency, and bandwidth. Similarly to the simulation, the HCPBG-3t-o1 first attenuation band is at 2.04 GHz (BW = 650 MHz), and for HCPBG-3t-o2, it is at 2.08 GHz (BW = 390 MHz). The second resonance appears at 2.92 GHz (BW = 260 MHz) and 3.05 GHz (BW = 292 MHz) for ori-



entations 1 and 2, respectively. Also, as predicted by simulation, a third wide resonance occurs at 6.01 GHz (BW = 1.4 GHz) for the case of HCPBG-3t-o1.

Figure 10. Measurement—transmission loss for HCPBG with 3 traces, orientations 1 and 2.

Finally, Figure 11 shows the transmission loss measurement results for the six trace HCPBG with a gap of 1.5 mm. Although it shows a significant bandwidth response and a similar profile, when compared to that of the simulated model, the bandwidth is reduced when we use -5.8 dB criteria. To have similar bandwidth, the criteria would need to be around -5 dB. The more substantial attenuation is at 4.55 GHz (-18.8 dB).

The simulation results show an error of less than 2% in the resonance frequencies below 4 GHz, while for values above 4 GHz, the error is less than 4%. The discrepancy between the measurements and simulation results for higher frequencies could be due to a mismatch between the dielectric characteristics of the PCB and the simulation model, over frequency. The variations in the fabrication process of the board would also impact the structure geometry; for example, changing the gap capacitance and displacing the resonance frequency.



Figure 11. Measurement—transmission loss for HCPBG with 6 traces—gap distances 1.5 mm.

5. Design of Reconfigurable HCPBG

First, it is essential to remind that the central concept of HCPBG concerns the model of two metal plates separated by gaps of etched copper. Also, the traces of each hexagon face and gaps define the features of an equivalent LC network. Then, it is possible to tune the resonance center frequency of the filter by actively controlling the parameters *L* and *C*. The simplest method to achieve frequency tuning is to switch on/off one or more traces that result in inductance changes. For this purpose, it is possible to use FET transistors or micro-electro-mechanical systems (MEMS) switches. The caveat of this approach is the additional capacitance of the FET transistor or MEMS while in an off state, making the implementation more complex.

On the other hand, it is also possible to change the LC network using a variable capacitor at the gaps close to the transmission line. The technique allows actively reconfigure the first resonance frequency adding parallel capacitance to the overall intrinsic gap capacitance. In this case, a varactor diode connected to a bias tee circuit with a port connected to a DC voltage supply allows controlling the capacitance. The varactor diode method is less complex than active inductor circuits or RF switches and provides fine-tuning of the first resonance frequency. Hence, the cost of implementing switches or the complexity of polarizing FET transistors for each trace of the HCPBG would be higher than using two varactor diodes to change the capacitance.

The selected HCPBG model for the reconfiguration study is the 6-trace type, and it can be observed in Table 1 case (f). The main dimensions used to create the reconfigurable HCPBG structure can be seen in Table 3.

Table 3. Reconfigurable HCPBG dimensions.

а	<i>g</i> 1	<i>g</i> ₂	t_l	t_w
28.5 mm	0.5 mm	0.5 mm	7.5 mm	1.5 mm

The model shown in Figure 12 was simulated in the CST studio software. It consists of a 50 Ω microstrip line on a 2-layer FR4 PCB, with a dielectric constant of 4.3, length of 90 mm, a width of 60 mm, and thickness of 1.6 mm. The ground plane is placed on the bottom side of the PCB, where the HCPBG structure is also etched. All pads, traces, and vias are designed to emulate the fabricated board.

On the bottom layer, we have the varactor diode $(D_{varicap})$ and the DC block capacitor $(C_{dcblock})$. The inductor (L_{choke}) to isolate the DC from the AC part of the circuit is placed on the top layer. A metal via connects the DC power supply traces from the top layer to the varactor on the bottom layer. At the end of the DC power supply trace, a capacitor (C_{supply}) is connected to the ground to represent the power supply's capacitive coupling to the common ground.

A simplified series RLC varactor model is used in the simulation. It is composed of a capacitor (C_T) in series with the parasitic inductance (L_S) and resistance (R_S). According to the datasheet of SMV1247 from Skyworks, L_S is 0.7 nH for the SC-79 package, and R_S is dependent on the applied reverse voltage (V_R), having a value that ranges from 2.5 to 9 Ω . The applied reverse voltage controls the varactor's capacitance (C_T). The rationale behind the choice of the SM1247 is that its capacitance range is in the same order as the gap capacitance of the HCPBG, giving a good dynamic range for frequency shifting. Table 4 shows selected C_T versus V_R used for the simulations.



Figure 12. Reconfigurable HCPBG—detailed simulation design view.

Table 4. Reverse voltage (V_R) versus capacitance (C_T) of SMV1247 varactor diode.

V_R	7.5 V	4.0 V	3.5 V	3 V	2.5 V	2.0 V	0.0 V
C_T	0.64 pF	0.77 pF	0.83 pF	0.95 pF	1.22 pF	1.88 pF	8.86 pF

The representation of the varactor's circuit model (C_T , R_S , L_S) and bias circuit electrical connections can be seen in Figure 13. The varactor's anode is connected to a common ground plane (PCB GND) and the cathode to the DC block capacitor (C_{dc_blk}), which is also connected to the HCPBG's inner GND. This configuration allows DC bias isolation from the RF ground, for the C_T works as a second DC block element. Completing the bias tee, the inductor (L_choke) isolates the RF signal from the DC power supply line. The DC power supply coupling capacitance (C_supply) connects the DC line to the PCB GND.



Figure 13. Reconfigurable HCPBG—varactor and bias tee schematic.

The discrete component values used in the simulation can be seen in Table 5.

Table 5. Simulation—discrete component values.

C_T	L_S	R_S	$C_{dcblock}$	L_{choke}	C_{supply}
0.64–8.86 pF	0.7 nH	2.5–9 Ω	1 µF	0.39 μΗ	10 µF

To validate the simulation model, we fabricated the testing sample using an LPKF milling machine. In Figure 14a, is the top view of the fabricated sample, showing the microstrip line, SMA connectors, choke inductors, and DC power supply traces and pads. The bottom side of the sample is shown in Figure 14b, containing the GND plane, HCPBG, varactor diodes, and DC block capacitors.



Figure 14. Reconfigurable HCPBG-6t-fabricated sample. (a) Top view; (b) bottom view.

The filtering characteristic of the reconfigurable HCPBG can be observed by the simulated and measured transmission loss (S21) in Figures 15 and 16, respectively. Tables 6 and 7 show the simulated and measured resonance frequencies and bandwidths for different V_R or C_T values.

Both results show similar curve profiles, and reasonable approximation with respect to resonance center frequency and bandwidth. For instance, considering C_T of 0.64 pF (V_R = 7.5 V), the simulated first resonance frequency f_1 is 2.49 GHz (BW_1 = 331 MHz) and the measured f_1 is 2.31 GHz (BW_1 = 260 MHz).

Starting the analysis with the simulation results, as we apply the lowest C_T value (0.64 pF) two main resonances occur, the first one at 2.49 GHz ($BW_1 = 331$ MHz) and the second one at 4.63 GHz ($BW_1 = 647$ MHz). As expected, when we increase C_T , the resonance frequencies are shifted to lower values. For example, if we take C_T values of 0.83 pF and 0.95 pF, f_1 is displaced by 100 MHz to 2.32 GHz ($BW_1 = 244$ MHz) and 2.22 GHz ($BW_1 = 209$ MHz), respectively. The same behavior is observed for the measurement results; for example, for V_R of 3.5 V (0.83 pF) and 3.0 V (0.95 pF), the f_1 is displaced by 70 MHz, going from 2.15 GHz ($BW_1 = 160$ MHz) to 2.08 GHz ($BW_1 = 160$ MHz).



Figure 15. Simulated—transmission loss for voltage values: 7.5, 4.0, 3.5, 3.0, 2.5, 2.0, 0.0 Volts.



Figure 16. Measured—transmission loss for voltage values: 7.5, 4.0, 3.5, 3.0, 2.75, 2.5, 2.33, 2.0, 0.0 Volts.

V_R (pF)	C_T (pF)	f_1 (GHz)	BW_1 (MHz)	<i>f</i> ₂ (GHz)	BW ₂ (MHz)
7.5	0.64	2.49	331	4.63	647
4.0	0.77	2.36	266	4.56	690
3.5	0.83	2.32	244	4.53	712
3.0	0.95	2.22	209	4.50	733
2.5	1.22	2.02	137	4.42	755
2.0	1.88	1.73	(-7.29 dB)	4.34	762
0.0	8.86	0.98	(-1.35 dB)	4.23	741

Table 6. Simulation—stop band center frequency and bandwidth (BW).

<i>V</i> _{<i>R</i>} (pF)	<i>C</i> _{<i>T</i>} (p F)	<i>f</i> ₁ (GHz)	BW_1 (MHz)	<i>f</i> ₂ (GHz)	BW ₂ (MHz)
7.5	0.64	2.31	260	4.78	650
4.0	0.77	2.21	230	4.58	710
3.5	0.83	2.15	160	4.55	750
3.0	0.95	2.08	160	4.52	810
2.75	-	1.95	130	4.49	780
2.50	1.22	1.82	97	4.45	810
2.33	-	1.72	(-10.19 dB)	4.45	840
2.0	1.88	1.5	(-6.97 dB)	4.39	810
0.0	8.86	0.85	(-1.26 dB)	4.29	840

Table 7. Measurement—stop band center frequency and bandwidth (BW).

The difference between the simulated and measured resonance frequencies is dependent on the varactor's bias voltage or the selected C_T . Considering C_T values of 1.22 pF and above, the first resonance frequency showed an error of 15%. On the other hand, for capacitance values of 0.95 pF and bellow, the difference is less than 8%. According to the datasheet, the varactor SMV1247 can show 7% to 11% variation in the capacitance, depending on the applied voltage. Added to that, we have the fabrication process and substrate dielectric variations. For the second resonance, around 4.5 GHz, for which the varactor does not play a strong influence, the error is only 3%.

Another observed characteristic is the bandwidth and maximum attenuation reduction, as the resonance is shifted to lower frequencies. This reduction is strongly influenced by the varactor's R_S . For example, considering C_T of 0.77 pF, then R_S is 3.9 Ω and simulated S21 equals to -20 dB @ 2.36 GHz. For C_T of 1.22 pF, then $R_S = 6.3 \Omega$ and simulated S21 equals to -12.6 dB @ 2.02 GHz. Due to R_S effect, the -10 dB criteria can not be achieved for 1.88 pF (S21 = -7.29 dB @ 1.73 GHz) and 8.86 pF (S21 = -1.35 dB @ 0.98 GHz), limiting the dynamic range of the reconfigurable filter. The measurement results confirm this characteristic. When we set V_R to 2.0 V, or equivalently, C_T to 1.88 pF, the filter attenuation stays above the -10 dB criteria (S21 = -6.97 dB @ 1.5 GHz). If we remove R_S from the simulation, the bandwidth is also reduced as the frequency shifts, however the entire range of C_T achieves the -10 dB attenuation criteria.

 C_T has a stronger influence over the first resonance frequency. For example, in Table 6, if we observe the resonance frequencies at 0.64 pF and 8.86 pF, f_1 shifts 1.51 GHz, however, the second resonance (f_2) shifts only 400 MHz.

For a -10 dB criteria, the measurements show a filtering capability that ranges from 1.72 GHz (V_R = 2.33 V) to 2.44 GHz (V_R = 7.5 V, upper band of f_1 = 2.31 GHz), that is 720 MHz bandwidth coverage.

6. Conclusions

In this paper, we presented a new planar PBG geometry based on the UCPBG: the HCPBG. It occupies less PCB area on the reference plane allowing a more compact usage of the space area. The simulations showed good accuracy when compared to that of the measurements. We showed through simulation and measurement that HCPBG unit cell structure can produce similar characteristics to the UCPBG, but it also adds additional resonance bands of interest. Furthermore, the rejection band is reconfigurable depending on the orientation and number of traces used, which results in more flexibility on filter designs for interfering signals. We also presented a reconfigurable HCPBG single-cell structure that can be controlled electronically. It allows for changing the rejection band profile by applying DC voltage at the control port. The control circuit, composed of two varactor diodes and a bias tee circuit, allows changing the gap capacitance of the HCPBG structure. The concept was demonstrated by EM simulation, including the varactor model and bias tee components. The measurement results point out that the resonance center frequency changes proportionally to the varactor's capacitance, thereby agreeing with the results observed in simulations. The simulated model showed less than 4% error in the

resonance center frequency. For higher frequencies, the difference is stronger, probably due to variations in the fabrication process and a model mismatch of the PCB dielectric characteristics. In the case of reconfigurable HCPBG, we observed that the varactor's capacitance plays a strong role in the difference between the simulated and measured S parameters. The first simulated resonances showed 15% differences with respect to the

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measured ones, however, the second resonance has a maximum of 3% variation.

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Abbreviations

The following abbreviations are used in this manuscript:

HCPBG	Honey Comb Photonic Band Gap
UCPBG	Uniplanar Compact Photonic Band Gap
PBG	Photonic Band Gap
EM	Electromagnetic
FR1	Frequency Range 1
РСВ	Printed Circuit Board
EBG	Electromagnetic Band Gap
EMC	Electromagnetic Compatibility
RIS	Reconfigurable Intelligent Surface
MEMS	Microelectromechanical Systems
BW	Bandwidth

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