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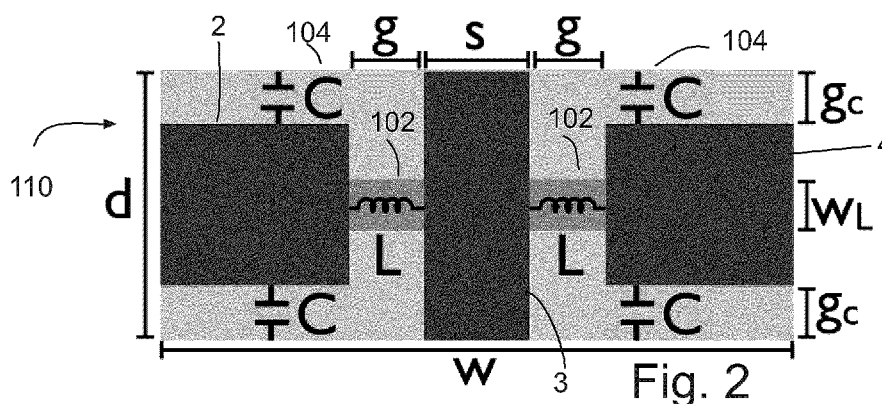
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(54) Title: APPLICATIONS OF METAMATERIAL ELECTROMAGNETIC BANDGAP STRUCTURES



(57) Abstract: An electromagnetic bandgap structure is formed by loading a conductor backed coplanar waveguide with inductors and capacitors selected to cause a frequency-dependent coupling between a parallel-plate waveguide mode and a coplanar waveguide mode to form an electromagnetic bandgap. The structure may be formed by printing (i.e. removal of metallization) on one side of a conventional double-sided printed circuit board.

APPLICATIONS OF METAMATERIAL ELECTROMAGNETIC BANDGAP STRUCTURES

TECHNICAL FIELD

[0001] Electromagnetic Bandgap Structures.

BACKGROUND

[0002] Planar periodic structures for the control of electromagnetic waves at microwave frequencies have grown in interest and use over the last decade, fueled by interest in the fantastic properties of electromagnetic metamaterials – engineered composite structures which exhibit behaviors not found in natural materials. Such structures have been used to create novel microwave circuit components such as couplers and filters, lenses, antennas, bandgap structures, and frequency-selective surfaces. However, the fabrication of many of these planar structures involve interconnects such as vias between layers, which can add to the cost and/ or the complexity of the manufacturing of the device. To overcome this, uniplanar (without interconnects between layers) structures have been proposed, however, these are generally poorly understood and so lack a rigorous model by which to design them, limiting their usefulness. Additionally, these structures are generally electrically large, on the order of one-quarter wavelength per period, which can result in a physically large structure once multiple periods are taken into account. Many of these devices can be generalized as artificial surfaces with which electromagnetic waves are controlled. Such surfaces have applications to antenna ground planes, lenses, or microwave circuits.

[0003] Parallel-plate modes arise in planar environments which are effectively shielded by metallic conductors.

[0004] Examples of such environments include high-frequency multi-layered printed-circuit boards (PCBs), where multiple ground, power, and signal planes may be interconnected with vias, or in closely spaced substrate-integrated-waveguide (SIW) components interacting due to leakage into the surrounding parallel-plate medium, or in the introduction of shielding planes to suppress backward radiation in aperture-coupled patch

antennas. In these situations, the unwanted excitation of parallel-plate modes can cause interference and false signalling, as well as reduction of antenna radiation efficiencies, degrading the overall performance of these systems.

[0005] Suppression of noise in the form of parallel-plate modes has been studied extensively in the literature. Decoupling capacitors have been used to give noise signals a low-impedance path through which to propagate away from sensitive circuit elements; however these generally require fairly large dimensions and are limited to relatively low-frequency applications. Newer methods include simple gaps or defects in the ground plane, commonly known as defected-ground structures (DGSs). However, the application of these techniques is often not guided by any general theory or rigorous analysis.

[0006] One popular technology employs cascaded periodic structures known as electromagnetic bandgap (EBG) structures, which have been used to prevent the propagation of parallel-plate modes in certain frequency bands and directions by employing periodically arranged resonators. In such structures, it is often required that the physical size of the EBG be on the order of one-quarter to one-half of a guided wavelength, which can be a considerable and expensive amount of space in high-density circuits once multiple EBG periods are taken into account. These structures may also employ a large number of vias that complicate their design. The Uniplanar Compact EBG structure (UC-EBG) is uniplanar and has a period of typically one-quarter of a wavelength. It is however not well characterized analytically. A further miniaturizable, uniplanar implementation of an EBG structure is highly desired.

[0007] The recent rise in the use of satellite-assisted positioning systems, such as the global positioning system, has further increased the desire for positioning accuracy. While positioning signals transmitted by these satellites are of incredible precision, there are numerous factors which cause inaccuracies when the signals are received on the earth's surface. Multipath interference is one of these factors, and is the phenomenon in which a signal is transmitted to a receiver along more than one physical path. Ideally, only one signal path exists - from the transmitting satellite, in a direct line-of-sight path to the receiver. However, due to the presence of other objects (typically large ground structures such as buildings), those signals can be reflected, multiple times, and some of these reflections can

arrive at the receiver after a significant time delay from the direct line-of-sight signal, causing an inaccuracy in the estimation of position. Signals reflected an odd number of times possess the reverse polarization (left-handed circularly polarized, or LHCP) from the original signal (which is right-handed circularly-polarized, or RHCP), such that the majority of multipath signals possess LHCP. Subsequently, the effects of multipath signals can be mitigated by suppressing this LHCP component.

[0008] However, these reflected signals have been found to travel to the receiver near the horizon at an angle tangential to, or lower than, a plane parallel to the ground, as shown in Fig. 30. Waves travelling towards the antenna in this direction are coupled into surface waves at the edge of the ground plane. If left unsuppressed, these surface waves carry the signals to the receiving antenna element. Subsequently, the LHCP radiation accepted by the antenna on or near the horizon can be reduced by suppressing surface waves on the antenna's ground plane.

[0009] Surface Wave Suppression

[0010] At GNSS frequencies, only one surface wave mode is supported by a conductive ground plane – the linearly polarized TM_0 surface wave (SW) mode. Several technologies for the suppression of SWs in antenna systems currently exist:

[0011] **Choke rings** are created by forming a corrugated ground plane around an antenna. These ground planes contain a series of grooves, which when excited by a SW with a wavelength corresponding to approximately four times the depth of the grooves, resonates and causes rejection of the SW. However, choke rings are large, bulky, heavy, and require expensive precision machining/fabrication.

[0012] **Resistive ground planes** are used to attenuate the SW mode before it reaches the antenna element. These ground planes are typically coated in a resistive film which has a high ohmic loss, and therefore can generally be thin and light as compared with the choke rings. However, these types of ground planes typically are not as well-performing.

[0013] **Electromagnetic bandgap structures** (EBGs) have lately received a large amount of attention in the academic community due to their ability to reject SWs similar to the choke ring structures, but in a planar form - such that the resulting ground plane can be lightweight, compact, and inexpensive. EBGs are composed of a periodic array of "unit

cells", of which several varieties have been proposed. Some of the recent challenges in the design of the EBGs are their miniaturization (each unit cell is generally on the order of one-quarter to one-half a wavelength) and several unit cells are typically required for acceptable performance, as well as their fabrication without the use of vias (resulting in a *uniplanar* design), since the fabrication of vias can be costly and/or difficult. EBGs may also be referred to as "photonic bandgap structures" (PBGs), "defected ground structures" (DGSs), "high impedance surfaces" (HISs), or "artificial magnetic conductors" (AMCs), depending upon their specific application.

[0014] **Patch antennas**

[0015] Many approaches have been taken to yield dual- or multi-frequency operation of microstrip patch antennas, which may be expensive and/or difficult to implement. Early efforts introduced "stacked" patches, in which patches of different sizes are layered vertically with each underlying layer serving as the effective ground plane of the above patch, which may be directly or parasitically excited. A more simple arrangement involved parasitically exciting patches on the same layer allowing for a single-layer design; however, the parasitic coupling was found to be much less effective in this orientation. Exciting various cavity modes on asymmetric patches has been used; however, this technique inherently requires that the excited modes have different field profiles, polarizations, and possibly different feeding mechanisms, which may not always be desired. A popular contemporary method of exciting various modes in a fully planar structure employs slots etched into the patch or ground plane, but such approaches tend to be empirical and are, therefore, ill-equipped for systematic design. Some designs may employ loading with non-planar components such as vias, but these add to manufacturing complexity. Other antennas, particularly those employing frequency-dependent dispersive properties, achieve multi-band operation through the excitation of a number of different resonance mechanisms; however, these behaviours tend to come at the expense of gain and polarization purity. Moreover, the radiation patterns of these antennas do not typically resemble those of the fundamental patch mode for all radiating frequencies.

[0016] More recently, metamaterial (MTM) structures have been integrated into the design to produce multiple resonances. MTMs are artificial structures possessing properties that may transcend those typically found in nature.

[0017] A class of these materials known as transmission-line (TL) MTMs are particularly useful in engineering dispersion properties in TL environments such as microstrip or parallel-plate waveguide (PPW), created by appropriately loading a TL with discrete inductors and capacitors at deeply subwavelength intervals. Moreover, the dispersive properties of these structures can typically be accurately modelled with an equivalent circuit employing TL theory.

[0018] However, the MTMs used in many of these works pose fabrication difficulties, such as the use of large numbers of vias.

SUMMARY

[0019] There is provided an electromagnetic structure comprising a conductor-backed coplanar waveguide structure including a central line, two side lines and a conductor backing, the conductor-backed coplanar waveguide structure being loaded with capacitors on the side lines and inductors connecting the side lines to the central line.

[0020] In various embodiments, there may be included any one or more of the following features: There may be a shielding plane, the conductor-backed coplanar waveguide structure and the shielding plane together forming a shielded conductor-backed coplanar waveguide structure. There may be capacitors on the central line. The inductors and capacitors may be formed of a uniplanar conductive layer. The inductors and capacitors may be formed by printing. The central line and side lines may be formed of the uniplanar conductive layer by printing.

[0021] There is also provided an enclosure structure comprising plural electromagnetic structures of any of the embodiments above arranged in parallel to surround a central area. In various embodiments, there may be included any one or more of the following features: The enclosure structure may be arranged around a via to suppress parallel-plate noise. The enclosure structure may be arranged around a ground plane for an

antenna to suppress surface waves in the ground plane. The enclosure structure may be arranged around a circular patch antenna to enable dual- or multi-band operation.

[0022] There is further provided a patch antenna comprising a patch and electromagnetic structures of any of the embodiments above, the electromagnetic structures arranged at edge portions of the patch.

[0023] In various embodiments, there may be included any one or more of the following features: the electromagnetic structures have different values of design parameters from each other in order to effect multiple operating frequencies. The patch may generally define a rectangle and the electromagnetic structures may be arranged at one, two, three or four sides of the rectangle. In the case of two electromagnetic structures arranged at two sides of the rectangle the sides at which the electromagnetic structures are arranged may be adjacent or opposing. The patch may be generally circular and the electromagnetic structures may be arranged around a perimeter of the patch.

[0024] There is also provided a dual band power divider having an input line and a pair of output lines, in which an electromagnetic structure as described above is formed at the junction of the input line and the pair of output lines.

[0025] In various embodiments, there may be included any one or more of the following features: the structure may have characteristic impedance equal to that of a quarter wavelength transformer designed to enable impedance matching at the input. The structure may be formed as a varactor.

[0026] There is also provided a method of using a structure as described above as a sensor comprising placing a material in proximity to or in contact with the structure and comparing an electrical characteristic of the structure when the material is in proximity to or in contact with the structure to a reference.

[0027] These and other aspects of the device and method are set out in the claims.

BRIEF DESCRIPTION OF THE FIGURES

[0028] Embodiments will now be described with reference to the figures, in which like reference characters denote like elements, by way of example, and in which:

- [0029] Fig. 1 is a cross-sectional view of conductor-backed coplanar waveguide (CBCPW);
- [0030] Fig. 2 is a schematic diagram showing a top view of a CBCPW structure loaded with inductors and capacitors;
- [0031] Fig. 3 is a circuit diagram showing an equivalent circuit model of the structure shown schematically in Fig. 2;
- [0032] Fig. 4 is a cross-sectional view of a shielded conductor-backed coplanar waveguide (S-CBCPW);
- [0033] Fig. 5 is a top view of a CBCPW or a cutaway top view of the middle layer of an S-CBCPW;
- [0034] Fig. 6 is a circuit diagram showing an equivalent circuit model of the structure shown schematically in Fig. 2 as applied to a shielded- (S-) CBCPW structure;
- [0035] Fig. 7 is cross section of a CBCPW showing a parallel-plate waveguide (PPW) type mode;
- [0036] Fig. 8 is cross section of a CBCPW showing a coplanar-waveguide (CPW) type mode;
- [0037] Fig. 9 is cross section of a CBCPW showing a coupled-slot-line (CSL) type mode;
- [0038] Fig. 10 is an isomorphic view of a CBCPW showing a parallel-plate waveguide (PPW) type mode;
- [0039] Fig. 11 is an isomorphic view of a CBCPW showing a coplanar-waveguide (CPW) type mode;
- [0040] Fig. 12 is an isomorphic view of a CBCPW showing a coupled-slot-line (CSL) type mode;
- [0041] Fig. 13A is a dispersion diagram for the equivalent circuit of Fig. 6 showing coupled even modes;
- [0042] Fig. 13B is a dispersion diagram for the equivalent circuit of Fig. 6 showing isolated even modes;
- [0043] Fig. 14 is a diagram showing the top surface of a printed unit cell;

[0044] Fig. 15 is a graph showing dispersion data for the equivalent-circuit model of Fig. 6 (solid curves) and simulated using HFSS™ (large dots);

[0045] Fig. 16 is a graph showing simulated PPW mode and measured scattering parameters;

[0046] Fig. 17 is a picture showing a fabricated PCB, with microstrip (MS), taper, and PPW regions of dimensions indicated;

[0047] Fig. 18 is a top view of an example setup for two-layer via-induced PPW-noise-suppressing radial EBG:

[0048] Fig. 19 is a cross-sectional side view of the example setup of Fig. 18;

[0049] Fig. 20 is a graph showing simulated scattering parameters of a single radial section of the 2D EBG of Fig. 18 with periodic transverse boundary conditions to simulate the full circular arrangement;

[0050] Fig. 21 is a graph showing a simulated magnitude of the scattering parameter S_{11} for the structure of Fig. 18 with absorbing boundaries and via excitations, with and without the EBG;

[0051] Fig. 22 is a graph showing a simulated magnitude of the scattering parameter S_{21} for the structure of Fig. 18 with absorbing boundaries and via excitations, with and without the EBG;

[0052] Fig. 23 is a graphic showing complex surface-current-density magnitudes on the EBG layer of Fig. 18 at 9 GHz where the EBG is expected to strongly suppress transmission of waves;

[0053] Fig. 24 is a graphic showing complex surface-current-density magnitudes on the EBG layer of Fig. 18 at 5 GHz where the EBG is expected to allow the transmission of waves;

[0054] Fig. 25 is a graph showing a simulated magnitude of the scattering parameter S_{11} for the structure of Fig. 18 with open boundaries and via excitations, with and without the EBG;

[0055] Fig. 26 is a graph showing a simulated magnitude of the scattering parameter S_{21} for the structure of Fig. 18 with open boundaries and via excitations, with and without the EBG;

- [0056] Fig. 27 is a picture of a Fabricated EBG on the middle layer, printed on a 0.254 mm Rogers RO-3010 substrate;
- [0057] Fig. 28 is a graph showing measured and simulated (assuming average $\epsilon_u = 9.7$ and a 50-um air gap between layers) values of the scattering parameter magnitude S_{11} , for the structure of Fig. 18 with and without the EBG;
- [0058] Fig. 29 is a graph showing measured and simulated (assuming average $\epsilon_u = 9.7$ and a 50-um air gap between layers) values of the scattering parameter magnitude S_{21} , for the structure of Fig. 18 with and without the EBG;
- [0059] Fig. 30 is a schematic diagram showing multipath interference affecting the received signal from a satellite;
- [0060] Fig. 31 is an equivalent circuit model of an EBG unit cell having capacitors on the center line;
- [0061] Fig. 32 is a graph showing dispersions of the equivalent-circuit model of Fig. 31 and numerically computed dispersions of a corresponding unit cell;
- [0062] Fig. 33 is a diagram showing a physical layout, as viewed from the top, of the unit cell for which the dispersion was numerically calculated and shown in Fig. 32.
- [0063] Fig. 34 is a diagram showing a physical layout of the capacitor on the center line (CPW strip line) of the unit cell of Fig. 33;
- [0064] Fig. 35 is a diagram showing a physical layout of the capacitor on a side line (CPW ground) of the unit cell of Fig. 33;
- [0065] Fig. 36 is a diagram showing a cascade of three unit cells, corresponding to that shown in Fig. 33 but trapezoidal, as viewed from the top;
- [0066] Fig. 37 is a diagram showing a layout of a fully printed GNSS antenna ground plane, with circularly arranged EBG comprising trapezoidal cells as shown in Fig. 36, as viewed from the top;
- [0067] Fig. 38 is a diagram showing a top view of a dual-band antenna comprising a microstrip-fed patch with MTM-EBG sections placed on the radiating edges;
- [0068] Fig. 39 is a top-view physical layout of a MTM-EBG unit cell with lumped components;

- [0069] Fig. 40 is a dispersion diagram of a MTM-EBG unit cell (on the front radiating edge of the antenna shown in Fig. 38);
- [0070] Fig. 41 is a top view of the physical layout of Fig. 39 with the components implemented as printed components;
- [0071] Fig. 42 is a graph showing S-parameters for transmission of the PPW mode through one and three printed cells of the MTM-EBG;
- [0072] Figs. 43A-43D are graphs showing simulated and measured values of the scattering parameter S_{11} for a conventional lower-frequency patch (Fig. 43A), a conventional higher-frequency patch (Fig. 43B), a MTM-EBG patch (lower-frequency resonance) (Fig. 43C), and a MTM-EBG patch (higher-frequency resonance) (Fig. 43D);
- [0073] Figs. 44A-44D are graphs showing simulated gains of the conventional versus MTM-EBG patch antennas, for a lower-frequency E-plane (Fig. 44A), a lower-frequency H-plane (Fig. 44B), a higher-frequency E-plane, (Fig. 44C) and a higher-frequency H-plane (Fig. 44D);
- [0074] Figs. 45A-45H are graphs showing simulated versus measured radiation patterns of fabricated antennas, in particular lower-frequency conventional patch E-plane (Fig. 45A), lower-frequency conventional patch H-plane (Fig. 45B), higher-frequency conventional patch E-plane (Fig. 45C), higher-frequency conventional patch H-plane (Fig. 45D), lower-frequency MTM-EBG patch E-plane (Fig. 45E), lower-frequency MTM-EBG patch H-plane (Fig. 45F), higher-frequency MTM-EBG patch E-plane (Fig. 45G) and higher-frequency MTM-EBG patch H-plane (Fig. 45H);
- [0075] Fig. 46 is a diagram showing a top view of a dual-band antenna comprising a microstrip-fed patch with MTM-EBG sections placed on the radiating edges, broader than the antenna shown in Fig. 38;
- [0076] Fig. 47 is a diagram showing a top view of a rectangular patch antenna employing different uniplanar MTM-EBGs perpendicular to both patch axes; and
- [0077] Fig. 48 is a diagram showing a top view of a circular patch antenna employing a uniplanar MTM-EBG on its radiating edge.
- [0078] Fig. 49: Dual-band power divider with fully printed embedded MTM-EBG.
- [0079] Fig. 50: Simulated response of embodiment of Fig. 49

- [0080] Fig. 51: Variable embedded MTM-EBG having a tunable capacitor in the center line and no capacitors in the side lines, embedded in power divider
- [0081] Fig. 52: Response to capacitance variation
- [0082] Fig. 53A: layout of the proposed MTM-EBG-based sensor with a test sample in contact with the structures,
- [0083] Fig. 53 B: S11 response of the sensor, showing a resonant dip without the sample (red curve) and with the sample present (blue curve)
- [0084] Fig. 54A: layout of a hypothetical two-dimensional sensor array, with the highlighted elements mapped uniquely by color to the resonant frequencies in Fig.54B.
- [0085] Fig 55A: suggested layout of planar resonators using MTM-EBG structures, with feed lines from the top and bottom, with rectangular cavity resonator to replace a SIW cavity
- [0086] Fig. 55B: MTM-EBG based ring resonator.
- [0087] Fig. 56A: Original MTM-EBG structure with a solid conductor ground plane on the back side of the dielectric,
- [0088] Fig. 56B: augmented MTM-EBG structure with patterning on both sides of the dielectric.
- [0089] Fig. 57 A: transmission (yellow) and reflection (red) of a proposed structure. Bandwidths and frequencies are bear suitable for the 2.4 and 5.8 GHz ISM bands.
- [0090] Fig. 57B: Beamforming occurring as a result of the proposed structure, with incident and outgoing angles being different and indicated with the solid black arrows.
- [0091] Fig.58: Layout of triple-band stub, left, with fabrication on the right. Close-up of MTM-EBG sections inset.
- [0092] Fig. 59: Simulated and measured frequency response of triple-band stub.
- [0093] Fig. 60 Example diagrams of the cross-section of the truncated conductor-backed CPW, with fields corresponding to a MS-like mode.
- [0094] Fig. 61. Dispersion curves of the MS-like mode for the 50 Ω cell (red, solid) and 35 Ω cell (blue, dashed).

[0095] Fig. 62: Layout of the proposed couplers: a) simulated, inset: relevant dimensions of the MTM-EBG unit cell, and b) fabricated coupler.

[0096] Fig. 63: Scattering parameters (magnitude, top, and phase, bottom) of the simulated (dotted) and measured (solid) scattering parameters. The two vertical black lines correspond to the operating frequencies of GPS L2 (1.228 GHz) and GPS L1 (1.575 GHz).

DETAILED DESCRIPTION

[0097] There is provided a uniplanar (i.e., realizable using a single double-sided metallized dielectric substrate without vias) wave-manipulation structure based on a grid of coplanar waveguide (CPW) transmission lines (TLs) and lumped reactive loading elements (inductors and capacitors) that can be engineered to produce passbands and stopbands as needed and whose implementation is significantly simplified and has the potential for substantial miniaturization as compared to existing technologies. The exotic metamaterial properties of this structure are obtained through a mechanism known as left-handed transmission, which has been shown to occur under certain conditions of periodic loading -- a technique by which electronic components such as inductors and capacitors are inserted along the structure at regular intervals. This periodic loading can be achieved using completely printed methods, for which meandered and/or strip inductors are placed in shunt in the CPW line, and series capacitors are realized by etching transverse gaps into the CPW grounds and/or the CPW strip line. The loading could also be realized with discrete lumped elements, for which tile meandered or strip inductors are replaced with a discrete inductor component, and discrete capacitor components are placed inside the transverse capacitive gaps.

[0098] Our research has determined a method by which the dispersive properties of the structure (i.e., the properties of the passbands and stopbands) can be very accurately modelled. This has not been achieved previously for a multi-mode uniplanar metamaterial structure. Specifically, the theory arrived at by our research details that the structure supports four dominant modes, namely:

[0099] • A parallel-plate waveguide (PPW) type mode, which is composed of primarily vertical electric fields in the dielectric region,

[00100] • A coplanar waveguide (CPW) type mode, which is composed of primarily horizontal electric fields, which are directed towards the center of the structure,

[00101] • A coupled slotline (CSL) type mode, which is composed of primarily horizontal electric fields, which are directed towards one of the edges of the structure,

[00102] • A surface-wave (SW) type mode, which is composed of primarily vertical electric fields above the dielectric region.

[00103] Our theory predicts the manner of coupling of these modes, which results in the dispersive properties of the structure. This understanding allows for the engineering of these structures, which in turn results in the following functionalities:

[00104] • Miniaturization. To achieve operation at the same frequency, the period of the loading (and hence the structure) could be reduced if the values of the loading components were increased. Similarly, the proposed structure can be used to create miniaturized devices over those already existing through the use of its lumped loading- importantly, due to the fact that the proposed model of this structure allows for the precise determination of the effect of this loading.

[00105] • Uniplanar control of waves. This structure enables coupling between the modes travelling in or above the dielectric (i.e. the PPW or SW modes) with those on the surface (i.e. the CPW or CSL modes), such that the first set of modes may be controlled simply by altering the patterned uniplanar surface.

[00106] • Extension to a two-dimensional structure. While the (one-dimensional) proposed structure has many uses on its own, our research has uncovered two layouts of this structure for which it can be used in two dimensions. Firstly, our research has shown that the one-dimensional behaviour of this structure can be extended to two dimensions by "distorting" the rectangular structure into a trapezoidal form. These trapezoidal sections are then arranged side-by-side around a common point to form a closed circle. This device can then be used to control the propagation of waves propagating in cylindrically symmetric environments, for which there are many applications, such as noise suppression emanating from a small source or the suppression of surface waves generated by an antenna. Secondly, initial studies have indicated that a two-dimensional intersection of four of these one-dimensional sections can form a structure analogous to a common structure already well

known as the uniplanar compact electromagnetic bandgap structure (UC-EBG). This connection is quite novel and useful, as the UC-EBG has been demonstrated to have practical value in many applications, but does not currently have an accurate analytical model with which its behaviour can be described and/or predicted. Describing the UC-EBG in such a manner would allow for its miniaturization, which currently cannot be easily achieved, and its integration into a larger variety of systems, such as those described in the following section. In fact, wherever the one-dimensional structure is used, the UC-EBG can be used to effect the same behaviour in two dimensions for propagation in any arbitrary (two-dimensional) manner. A detailed understanding of its operation based on our newly proposed rigorous analytical methods should allow for benefits beyond what has already been demonstrated.

[00107] Applications:

[00108] *Electromagnetic bandgap structures*: The coupling of the left-handed loaded CPW mode with the right-handed PPW mode(s) allows the formation of bandgaps in these modes. This allows for the development of EBGs for operation on PPW modes, with bandgaps that can be controllable by adjusting many of the structures parameters, and without the use of vias. This results in designs suitable for use in both low-cost applications where vias are undesirable, and in high-frequency applications where large lumped loading components are impractical and microvias cannot be reliably fabricated. The interaction with surface-wave (SW) modes allows for the development of surface-wave suppressing metasurfaces, such as those used in antenna applications. Both of these mechanisms also allow for the suppression of the CPW and CSL modes, which allow for the creation of mode-dependent bandgap structures. Furthermore, with the addition of variable-valued loading components, these bandgaps could be made tunable, for reconfigurable systems applications or sensors.

[00109] Tuneable filters may be created which can be designed very simply using the transmission-line analysis derived for our EBG structures. This analysis can be considered an alternative to existing filter theory.

[00110] Planar, high-quality-factor resonators may be created in rectangular or circular form. These may be similar to substrate-integrated-waveguide (SIW) resonators in achieving high quality factors, but without the necessity of vias.

[00111] *Sensing*: The coupling of the CPW and PPW modes results in the creation of a highly dispersive mode, which in turn can cause very narrowband resonances. These resonances can be detected by observing the return loss of either the PPW or CPW modes, allowing for a very compact, inexpensive, and lightweight sensor or sensor array. The miniaturization afforded by the loading components allow the structure to be miniaturized, allowing a sensor array composed of these structures to possess a very high spatial resolution compared to conventional sensors. Many of the design parameters can be adjusted, to allow for a wide range of operating bands from radio to millimeter-wave frequencies. The prominent location of the loading components on the top face of the structure allow for the sensing of a materials's electric and/ or magnetic properties. By replacing the traditional lumped components with temperature- or field-sensitive components (e.g. ferroelectric or ferromagnetic materials), high –spatial-resolution temperature or field sensors may be realized.

[00112] Miniaturized sensor arrays may be developed in which the individual "pixel" elements' size are determined by the resonator dimensions. Since our unit cells have been demonstrated to be created in form factors much smaller than one wave length, the arrays could achieve a very high spatial resolution compared with existing technologies. Many of the design parameters can be adjusted, to allow for a wide range of operating bands from radio to millimeter-wave frequencies.

[00113] The structure is sensitive to the values of the loading components (capacitors and inductors). The prominent location of these loading components on the top face of the structure implies that, if these loading components are replaced with a material-under-test (MUT), it may be possible to sense a MUT's electric and magnetic properties by examining shifts in the properties of the structure. This could allow for the development of multi-use sensors, which currently require two separate and unrelated devices.

[00114] Multi-band sensors may be developed due the capacity of the structure to support multiple propagating modes. These devices could be used for sensing properties at

various discrete bands, or for enabling broadband sensing over a single, although enhanced, bandwidth.

[00115] *Antennas:* Multiple applications of this technology to antennas exist. Our research has revealed that by integrating these structures into widely prevalent patch antennas as EBGs, dual-band behaviour could be achieved in a fully printed manner while still achieving moderate gain and low cross-polarization. The application of a larger number of cells, each slightly de-tuned, could result in wide-band behaviour. Another type of antenna that could utilize these structures is a leaky-wave antenna. Due to the fact that the dispersive properties of the structure can be easily predicted, the region in which leaky-waves are produced can be easily identified, and moreover, the angle of radiation can be tuned to a desired value with the structure's various properties. Again, the operating bandwidth of these antennas could be specified in a large range of frequencies, and could be made tunable with the inclusion of variable-valued loading components. The use of this technology may enable miniaturization of antennas and a lower cost, lighter weight solution over current products.

[00116] Utilizing the multi-band nature of the unit cells, antennas which operate at multiple discrete bands could be created — or, by compacting the resonances close together, a single broad bandwidth antenna may be realized.

[00117] Antenna radiation beam-forming substrates may be enabled by use of the structure's ability to interact with surface-wave modes, allowing for antennas with highly desirable or reconfigurable radiation patterns.

[00118] Leaky-wave antennas may be created owing to the accuracy of the developed transmission-line model in predicting the dispersions of the structure's guided modes.

[00119] Superstrates may be enabled which perform beam-shaping or allow for the creation of miniaturized Fabry-Perot cavity antennas.

[00120] *Absorbers:* Since the absorbing properties of the structure are related to its leaky-wave behaviour, the absorption frequency could also be easily specified. Furthermore, the angle of absorption and angle range can be set by the structure's dispersive properties. Initial research indicated that if a flat dispersive profile is used, then the absorption should take place over a wide range of incident angles. The inclusion of variable-valued loading

components could not only change the frequency of operation, but also potentially change the angle of absorption as well, leading to partially reflecting surfaces which only absorb from specific and adjustable ranges of incident angles. Traditional absorbers are often created with three-dimensional shapes such as pyramids for wideband operation. Our technology may enable wideband operation while maintaining a planar profile, allowing for large space savings.

[00121] *Waveguides*: The bandgap properties of the structure may also be used to contain power inside a specific region while guiding it along another axis, creating the boundary conditions necessary for a waveguide. An example utilizing the PPW mode may be a far simpler and cheaper alternative to the substrate-integrated waveguide (SIW) which, instead of possessing large numbers of vias embedded into the substrate, simply contains the proposed fully printed structures on its easily-accessible top face. The development of such novel SIW structures also naturally prompts the development of novel SIW-based waveguide components such as filters and couplers. These structures would have the advantage of being lower cost (owing to the absence of vias), reduced width (since the EBGs could potentially relax the half-wavelength condition), and lower loss (due to the presence of an additional TEM mode) than traditional SIW.

[00122] The structure's bandgap may also be employed to decouple adjacent waveguides in high-density, high-frequency environments such as PCBs, allowing them to possess a much higher degree of isolation.

[00123] Using the structure in devices such as couplers, T-junction power splitters, and fin- or diaphragm-loaded waveguides may result in new or enhanced properties, and/or miniaturization.

[00124] *Frequency selective surfaces*: While the UC-EBG has been studied for amplitude and phase control of transmission through a surface for any arbitrary polarization, the one-dimensional structure may find use in polarization-specific applications, such as polarization filtering screens or transmitarrays.

[00125] Partially reflecting surfaces with tuneable properties could be enabled by removing the structure's solid conductor backing and/or shield layers.

[00126] Polarization-sensitive surfaces may be developed since the structure has a dominant axis along the direction of the host transmission-line. This could be used to implement an antenna superstate which acts as a polarization filter.

[00127] Lenses, especially the novel negative-refractive-index (NRI) lens, could be created by employing the various properties of the structure's supported modes, which can be designed to exhibit carefully tuned propagation characteristics. Volumetric lenses can be created by stacking layers of the 2D structures.

[00128] Artificial prisms which make use of the structure's dispersive properties can be developed which spatially separate frequency-multiplexed signals, enabling ultra-high speed (de-)multiplexers for modern communications systems.

[00129] Cloaking structures could be enabled with the cylindrical form of the structure, by again stacking multiple layers, or cloaking surfaces (mantle cloaks) may be created with the uniplanar structure.

[00130] *Printed microwave circuit components:* Passive microwave devices such as couplers and filters often rely on their electrical length to determine various properties. The dispersive nature of the coupled CPW and PPW modes allows for the design of any practical, desired electrical length of tile structure. This can be utilized in the creation of fully printed and/ or miniaturized circuit components. Variable-valued components would allow for devices such as tunable filters, which additionally could have different properties for each supported mode.

[00131] Since the structures may be implemented in a fully-printed fashion (made entirely of, e.g., copper traces on a dielectric substrate), it would be possible to fabricate the previously mentioned devices on flexible dielectric substrates, allowing the creation of conformal and/or wearable components.

[00132] 3-D printing may be employed to create the structures, as opposed to standard PCB processes. This could allow for greatly increased loading component values, while still employing a cost-efficient and simple manufacturing process. Throughout this application including the claims, the term "printing" is used to refer both to additive techniques, such as additive 3-D printing methods, or simply printing a metal surface on a de-metallized substrate, and to subtractive techniques, such as lithography in which metal is removed to

create an impression. The term should not be taken as limited to any particular variant of printing or lithography.

[00133] Potential products:

[00134] 1) A printable multi-band or wide-band patch antenna (near term)

[00135] 2) A textured metallization layer for the suppression of noise in parallel-plate environments (near term)

[00136] 3) A printable surface-wave suppressing ground plane for precision GPS antennas (near term)

[00137] 4) A substrate-integrated waveguide without the use of vias (near term)

[00138] 5) A sensor array for temperature or dielectric constant measurements (long term)

[00139] 6) A circularly polarized leaky-wave antenna (long term)

[00140] 7) A wide-angle, metamaterial absorber (long term)

[00141] 8) Miniaturized, printable microwave couplers and filters (long term)

[00142] Other planar, periodic (often metamaterial-based) structures exist (such as the Sievenpiper mushroom structure, and the UC-EBG) which serve similar functions. However, this structure is uniplanar, can be miniaturized, and has shown to be accurately modelled to enable rapid and simple design. The transition of other such structures to commercial applications has been complicated and protracted by the lack of a simple design methodology and by drawbacks such as complex fabrication and the use of electrically large structures -- all of these may be avoided with the use of our structure, enabling it to be much more commercially viable. Our structure has the potential to represent one of the first successful mobilizations of metamaterial technology in industry.

[00143] The suppression of parallel-plate modes is difficult to control over electrically short lengths. The use of electromagnetic bandgap (EBG) technology allows for a large suppression with controllable bandwidth, but traditional EBGs are often too electrically large to be used in practice. Moreover, in some forms they require several metallization layers and/or interconnecting vias. We introduce a metamaterial-based EBG that is miniaturized, uniplanar, and fully printable for the suppression of signals carried by the parallel-plate mode. We also present a corresponding multiconductor transmission-line analysis for

accurate modelling of the EBG's dispersive properties, which arise from the coupling of contradirected forward and backward modes. The theory is supported by full-wave finite-element-method simulations and verified by measurements of a fabricated EBG. To demonstrate the practical value of the metamaterial-based EBG, we propose an alternate implementation that extends the one-dimensional structure to a two-dimensional, radial EBG suitable for the suppression of high-frequency parallel-plate noise coupled between adjacent via interconnects. The simulation and measurement results for this device were found to be in agreement with each other and with the predicted bandgap.

[00144] One method for the miniaturization of EBGs is the application of transmission-line metamaterial (TL-MTM) techniques. TL-MTMs have been used extensively for the miniaturization of a number of microwave-circuit components, and their capacity for precise control of passband and bandgap properties has been well-documented. TL-MTMs operate on the principle that the introduction of reactive loading components (e.g., capacitors and inductors or additional resonators) in series or shunt into a regular TL allows one to engineer the phase shift per unit length, with the potential to mimic the behavior of much longer unloaded TLs. Since TL theory is well understood, it proves useful in obtaining analytical expressions for the propagation characteristics of the entire TL-MTM system. Multiconductor EBGs can also be modelled to a high degree of accuracy as TL-MTMs through a multiconductor transmission line (MTL) analysis.

[00145] MTMs afford the ability to control dispersion properties such as the phase velocity and attenuation constant for each supported mode, subject to the physical constraints imposed by causality. TL-MTMs, a sub-class of a large variety of MTMs, can be realized by periodically inserting shunt inductors and series capacitors into a host TL structure.

[00146] Another advantage of TL-MTMs is that they can support a *backward-wave* propagation characteristic (also referred to as 'left-handed' or 'negative refractive index'), in which the desired mode incurs phase *advance* as it propagates. In this case, the phase and group velocities possess opposite signs, where the group velocity represents power flow in an isotropic medium. When this backward mode interacts with a traditional forward mode, contradirectional coupling causes the formation of a bandgap. Importantly, this behavior can be realized when the size of the EBG unit cell is extremely sub-wavelength.

[00147] An embodiment of a novel one-dimensional uniplanar EBG based on transmission-line metamaterials is introduced for parallel-plate noise suppression, and is analyzed using multiconductor transmission-line theory. It is shown that the bandgap arises as the result of the contradirectional coupling of the parallel-plate waveguide mode and a backward coplanar waveguide mode. The concept is illustrated using a fully printed implementation suited to high-frequency applications which exhibits an 83% fractional bandgap.

[00148] The “host” transmission-line used for the EBG structure is a one-dimensional conductor-backed coplanar waveguide (CBCPW), a cross-section of which is shown in Fig. 1. The structure, indicated generally by reference numeral 100, has four independent conductors, a backing conductor 1, center (strip) line 3 and side (typically grounded, and thus referred to as grounds) lines 2 and 4. The structure therefore supports three TEM modes. The electric-field lines corresponding to these modes are depicted in Figs. 7, 8 and 9 and in Figs. 10, 11 and 12: the first, shown in Fig. 7 and Fig. 10, is a parallel-plate waveguide (PPW)-type mode, the second is a coplanar waveguide (CPW) mode, shown in Fig. 8 and Fig. 11, and the third is a coupled slot line (CSL) mode, shown in Fig. 9 and Fig. 12. By loading the CPW and CSL modes with shunt inductors 102 and series capacitors 104, as shown in Fig. 2, these modes are forced to support only a negative-refractive-index (NRI) mode (also referred to as backward, or left-handed). The PPW and CPW modes can be made to couple strongly, and their contradirectional nature causes a substantial bandgap to form. Fig. 2 shows a unit cell 110 except that the capacitors 104 are shown twice (at the top and bottom); multiple unit cells can be arranged in series (and in parallel with the addition of more conductors).

[00149] In another embodiment the host TL selected is the shielded conductor-backed coplanar waveguide (S-CBCPW), which is shown in cross section in Fig. 4. Fig. 5 shows a top view of the center layer of the S-CBCPW 106 or top layer of the CPCPW. Since both the series and shunt loading components can be inserted into the three coplanar-waveguide (CPW) conductors on the same plane (conductors 2, 3, and 4 in Fig. 4), this host TL enables a fully uniplanar design without the need for vias. Furthermore, the presence of the conductor backing (conductor 1) and shield (conductor 5) allow for the interaction of the

CPW mode with parallel-plate waveguide (PPW) modes, supported between these conductors and those of the CPW. The two parallel-plate modes, corresponding to fields above or below the CPW conductors in Fig. 4, will be referred to as the *upper* and *lower* PPW modes, respectively. When the upper PPW region is air-filled ($\epsilon_u = 1$) and its height h_u is sufficiently large (typically $h_u \gtrsim \lambda/10$), it has been found that the upper PPW mode can be an effective low-frequency model for a loosely bound TM_0 surface-wave (SW) mode.

[00150] As with the CPCPW embodiment, this MTL system not only supports the PPW and CPW modes, but also a coupled slot-line (CSL) mode. Figs. 7-9 and 10-12, which depict the electric-field lines corresponding to these modes for the CPCPW, also apply to the modes for the S-CPCPW, with the addition of the upper PPW mode (not shown) for the S-CPCPW. These TEM modes may be classified as either *even* or *odd*, based on their electric field distributions. The even modes are described by an electric field tangent to the symmetry plane indicated by the white dashed lines in Figs. 10 through 12 (equivalent to a perfect-magnetic conduction (PMC) boundary), while the odd nodes can be considered to be those which support an electric field *normal* to the symmetry plane (equivalent to a perfect-electric-conducting (PEC) boundary). Even modes have the potential to couple with other even modes (and likewise for odd modes) while even and odd modes do not couple. According to these definitions, the PPW, CPW, and SW modes are even modes, whereas the CSL mode is odd.

[00151] MTL theory can be used to model this system. The host TL properties are determined by extracting the per-unit-length capacitance and inductance from finite-element-method (FEM) simulations (assuming PMC boundaries on the transverse edges of the S-CBCPW TL, as indicated by the dashed lines in Fig. 4), from which propagation constants and characteristic impedances are derived. The TL-MTM unit cell is created by periodically loading the waveguide structure with series capacitors and shunt inductors, as shown in the MTL equivalent circuit in Fig. 6 for the S-CBCPW version or in Fig. 3 for the CBCPW version.

[00152] The dispersive properties of the TL-MTM can be analyzed by assuming an infinite cascade of unit cells. Firstly, the unit-cell equivalent circuit in Fig. 3 or Fig. 6 is

generalized to an ABCD transmission network, in which the input field quantities (currents and voltages) of the n^{th} unit cell are related to those at the output as follows:

$$\begin{bmatrix} \vec{V}_n \\ \vec{I}_n \end{bmatrix} = \begin{bmatrix} [A] & [B] \\ [C] & [D] \end{bmatrix} \begin{bmatrix} \vec{V}_{n+1} \\ \vec{I}_{n+1} \end{bmatrix} = \begin{bmatrix} [A] & [B] \\ [C] & [D] \end{bmatrix} \begin{bmatrix} \vec{V}_n \\ \vec{I}_n \end{bmatrix} e^{-\gamma d} \quad (1)$$

[00153] Bloch's theorem has been invoked to relate the input and output circuit quantities between the ports as indicated, where d is the physical length of the unit cell imparted by the host TL and $[A]$ is the sub-matrix of the unit cell's transmission matrix describing the transmission of voltage across the ports. The propagation constants $\gamma = \alpha + j\beta$ represent the complex coupled Bloch-mode solutions supported by this system. Using (1) and the commutative property of the sub-matrix components of the symmetric, reciprocal transmission network, the Bloch modes can be simply expressed as the solution to the characteristic equation,

$$\det([A] - [I] \cosh(\gamma d)) = 0 \quad (2)$$

from which the frequency dispersion of each of the modes (γ as a function of ω) may be obtained. An example one-dimensional dispersion diagram, based on the equivalent circuit in Fig. 6, is shown in Figs. 13A and 13B, which respectively present the dispersions of the coupled and corresponding isolated modes. These diagrams will be used to explain several notable features. Since the odd (CSL) mode does not couple with the other modes, its dispersion curves have been omitted for clarity.

[00154] Referring to Figs. 13A and 13B, the dashed-dotted black line represents the vacuum light line, while the dotted line represents propagation in the substrate dielectric, or equivalently, the dispersion of the isolated PPW mode. The solid grey lines correspond to $\beta d = \text{Im}(\lambda d)$, while the dashed grey lines correspond to $\alpha d = \text{Re}(\gamma d)$. Generally, the sign of the slope at any point of the βd curves on this diagram indicates the direction of the mode's group velocity (hence, direction of power flow) relative to its phase velocity. Since here we excite the coupled system using a PPW mode, we establish the reference that power flows into the system in the forward direction. Therefore, for the purposes of the present discussion, a positive slope corresponds to a power flow in the positive direction, and likewise a negative slope indicates power flow in the negative direction.

[00155] It is worth noting that the attenuation constants shown in Fig. 13A and 13B do not correspond to resistive losses, which were not considered in the analytical derivation of these modes, but rather represent reactive attenuation due to two distinct mechanisms. The first mechanism arises from a mode being *cut-off*, which is to say that the propagation constant exhibits $\beta \ll \alpha$. This describes an evanescent mode, in which power is reflected due to an inability of the system to support propagation. The second mechanism arises from contradirectional forward-backward coupling. In this case, attenuation results from the coupling of power from one mode travelling in one direction into another mode travelling in the opposite direction, such that there is no net transmission of power for the infinite periodic structure. Since propagation occurs simultaneously with attenuation, the coupled mode is a complex mode exhibiting a complex propagation constant with $\alpha \approx \beta$. The presence of either form of attenuation may be exploited to suppress modes of interest.

[00156] To begin, it is worthwhile considering the dispersions of the isolated (even) modes, which are determined by removing the unnecessary conductors from the MTL equivalent-circuit model. The TL properties, however, are computed assuming that these conductors exist, but serve only as parasitic elements. The origin of these modes is evident from the unit cell in Fig. 2 (illustrated for the CPCPW but also applicable to the S-CPCPW; the conductor backing, present in both versions, and shield, present in the S-CPCPW version only, are both omitted) and its MTL equivalent-circuit model (for the S-CPCPW version) in Fig. 6. In particular, it can be seen that the shunt capacitors and series inductors serve to load the CPW mode, such that it exhibits a backward characteristic. The forward characteristic of the PPW modes are largely unaffected by the loading.

[00157] Coupling between these isolated modes has two criteria: 1) mode-matching of the transverse fields, and 2) phase-matching in the longitudinal direction. The CPW and PPW modes exhibit a large degree of field overlap, and this satisfies the first criterion. The second criterion is satisfied where the two isolated-mode dispersion curves intersect on the dispersion diagram.

[00158] Recognizing these features in the coupled system of Figs. 13A and 13B, several regions may be identified. At low frequencies, only the PPW modes propagate. As frequency increases towards 2.4 GHz, the lower PPW mode becomes increasingly dispersive

as it couples more strongly with the backward CPW mode. At 2.4 GHz, the CPW mode starts to propagate; however, since it is strongly coupled with the lower PPW mode, the contradirectional power flow between the forward lower PPW mode and the reactively loaded backward CPW mode results in a complex mode bandgap from 2.4 to 5.0 GHz. It should be noted that the portion of the attenuation (αd) curve above 5.0 GHz which exists between 60° and 90° corresponds to the cutoff CPW mode, and is *not* part of a complex-mode system with the propagating lower PPW mode above 6 GHz. From 5.0 GHz to 6.0 GHz, the lower PPW and CPW modes start to decouple, after which propagation of the lower PPW mode begins to be restored (identified by the slope of the asymptote at higher frequencies). The CPW mode's propagation is not restored until well above 10 GHz. Lastly, there is weak coupling between the forward lower PPW mode and forward upper PPW mode near 6.5 GHz. Although in this document we focus exclusively on the even modes, it is worth noting that the odd CSL mode has a bandgap between 6.5 and 7.0 GHz and is cut off below 3.2 GHz. It is unaffected by the coupled system, since being odd it does not directly couple with any other supported mode.

[00159] The theory presented above was validated using a proof-of-concept design, which possesses a one-dimensional layout for ease of design, fabrication, and characterization. The EBG is designed for fabrication on a single substrate, and as an initial goal of the design, suppression of the (lower) parallel-plate mode is sought between 2.4 and 6.0 GHz. A unit cell is designed with use of the equivalent-circuit model to be electrically small and to employ low-valued LC loading, which enables its realization in fully printed fashion. Theoretically, this topology can be modelled using the conditions previously given for the modelling of the SW mode – that is, the upper dielectric may be assigned a relative permittivity of 1, and the shield height (h_u) may be made sufficiently large (100 mm is used in this case).

[00160] The layout of the LC-loaded CPW layer of the designed unit cell 110 is shown in Fig. 14; the dimensions of the design are the unit cell period $d = 5$ mm, width $w = 10$ mm, the CPW strip line width $s = 0.2$ mm, the CPW gap width $g = 0.5$ mm, loading capacitor length $g_c = 0.8$ mm, and loading inductor width $w_L = 0.2$ mm. A minimum feature size of 0.2 mm is used for the interdigitated capacitor's finger and gap widths, for ease of

fabrication using standard etching processes. The printed loading components are designed (estimated using empirical formulas and then mildly tuned in simulation) to provide loading values of $C = 0.8$ pF and $L = 0.8$ nH. The properties of the dielectric ($h_l = 1.524$ mm, $\epsilon_l = 3.66$, $\tan \delta = 0.004$, and 1-oz. copper cladding on both sides with a bulk conductivity of 5.8×10^7 S/m) are determined by the preselected substrate (Rogers™ RO4350™).

[00161] The EBG's dispersive properties were confirmed by performing an eigenmode simulation in HFSS. This simulation setup involves embedding the unit cell in a vacuum box with a perfectly matched layer, master/slave, and PMC boundaries applied to the top surface, longitudinal faces, and transverse faces, respectively. These support the necessary fields to simulate an infinite array of unit cells in the transverse direction.

[00162] The results of this simulation for the even-mode solutions are shown in Fig. 15, along with the curves obtained from the MTL equivalent-circuit model. The solid curves dispersion data for the equivalent-circuit model of Fig. 6 and the large dots show data simulated using HFSS™. Generally, they demonstrate excellent agreement, but there is a moderate divergence between these data towards larger βd values. This is attributed to the fact that the printed loading components cannot strictly be regarded as lumped, as assumed by the equivalent-circuit model; indeed, they are frequency-dependent, and this attribute is most evident when they are responsible for generating large phase shifts per unit cell.

[00163] The weak interaction between modes near 6.5 GHz validates the previous statement that the vacuum-filled upper PPW mode with a relatively large height (as was modelled as described above) can be a good approximation for the loosely-bound $T M_0$ SW mode, since an open boundary condition, rather than a shield conductor, was used above the EBG layer in this simulation. This is a behaviour which may be exploited in the formation of SW bandgaps, such as in the design of SW-suppressing ground planes for antenna applications.

[00164] The parallel-plate-mode suppression ability of the EBG structure was examined by simulating the scattering parameters of the PPW mode. Using HFSS, nine EBG unit cells were cascaded and the (lower) PPW mode was excited using waveports. PMC boundary conditions were again used on the transverse faces, and a radiation boundary was used on the remaining faces above the unit cells. A 25 mm section of unloaded PPW was

used to interface the waveports to the EBG. The results of this simulation are shown by the solid and dashed curves in Fig. 16. The solid curve shows simulated data for the S_{11} scattering parameter, the dashed curve shows simulated data for the S_{21} scattering parameter, the dotted curve shows measured data for the S_{11} scattering parameter, and the dash-dotted curve shows measured data for the S_{21} scattering parameter. The 10-dB insertion-loss points indicate a bandgap region from approximately 2.6 to 6.4 GHz, which is very close to the design criteria, and also very close to the dispersion data given by the HFSS eigenmode simulation. These results validate the accuracy and utility of the equivalent-circuit approach in predicting and designing the EBG bandgap properties.

[00165] To confirm the simulation results, a PCB containing the designed, fully printed EBG was fabricated. Using a 60-mil (i.e. 1.524-mm-thick) Rogers™ RO4350™ substrate, a 5×9 grid of unit cells was connected to a PPW, in order to sufficiently approximate the simulation setup. This PPW was then linearly tapered to a 50Ω microstrip (MS) line for ease of measurement. The fabricated structure is shown in Fig. 17, along with the appropriate dimensions. The total length of the EBG is 45 mm, and the total width is 50 mm. The length of the PPW region on either side of the EBG is 10 mm, and the microstrip sections are 20 mm long and 3.3 mm wide. The linear tapers connecting the PPW and MS were 50 mm long. SMA connectors were used to interface an Agilent™ Technologies N5244A vector network analyzer (VNA) with the PCB to perform the measurements.

[00166] The measured data are plotted in Fig. 16 along with the simulated data, and it is clear that they exhibit very good general agreement, despite the finite width of the EBG section and the large taper and microstrip sections, which were not included in the simulation model. Indeed, in both data the bandgap behavior of the EBG is clearly visible between 2.6 GHz and 6.4 GHz, as indicated by the dashed vertical lines. It should be noted that the resonant behavior below the bandgap region is owed to Fabry-Perot resonances of the highly dispersive coupled PPW-CPW mode. Discrepancies in the upper passband may be attributed to the frequency response of the microstrip and taper sections in the fabricated device.

[00167] It is worth noting that the electrical size of the unit cell over the designed bandgap region ranges from approximately $l_g/12$ to $l_g/5$, where l_g is the wavelength in the dielectric. This demonstrates the strong degrees of miniaturization possible with the TL-

MTM approach, which can be further appreciated by noting that the strength of the loading components (and hence miniaturization) is only limited by the minimum feature size of the manufacturing process.

[00168] TWO-DIMENSIONAL RADIAL EBG

[00169] The validation of the bandgap properties of the 1D EBG prompts us to examine whether it may be straightforwardly extended to parallel-plate mode suppression in multilayer, 2D applications. For example, parallel-plate noise is detrimental to signal integrity in high-speed PCBs, which contain multiple ground and/or power layers. This noise can be created by the routing of signal paths between layers with the use of vias, and the resulting noise propagates away radially through parallel-plate modes. In order to suppress this noise, a 2D solution is required. An EBG is provided here which can be constructed in radial form to decrease coupling between two parallel vias, as shown in Figs. 18 (top view) and 19 (side view). Plural electromagnetic structures as described above are arranged arranged in parallel to surround a central area. This EBG, generally indicated by reference numeral 112, is composed of trapezoidal sections 114, which are slightly distorted sections of cascaded one-dimensional unit cells that have been arranged side-by-side in order to form a complete circle. Note that in this embodiment the widths of the inductors 102 have been dramatically increased, and the lengths greatly shortened, in comparison to embodiments discussed above.

[00170] This setup is similar to that used in R. Abhari and G. V. Eleftheriades, “Metallo-dielectric electromagnetic bandgap structures for suppression and isolation of the parallel-plate noise in high-speed circuits,” *IEEE Trans. Microw. Theory Tech.*, vol. 51, no. 6, pp. 1629–1639, 2003, which compared the transmission between two vias in a bi-layer medium with and without a Sievenpiper mushroom EBG, in order to determine its effects. The Sievenpiper structure operates extremely well as an EBG, but its construction is complicated by the requirement for a via for each unit cell. The proposed radial EBG enables the suppression of signals coupled into the parallel-plate mode through 2D cylindrical waves, while maintaining its simplistic 1D, uniplanar design approach. To demonstrate the versatility of the design procedure and exploit its fully printed nature, this EBG is designed

to present a bandgap around X-band, where discrete (surface-mount) inductors and capacitors cannot be used due to their typically low self-resonance frequencies.

[00171] This embodiment of an EBG is designed to suppress the upper parallel-plate mode supported by a $h_u = 0.254$ -mm (10-mil) RO3010™ ($\epsilon_u = 10.2$, $\tan \delta = 0.0035$) substrate. The EBG is realized on the bottom metallization layer and the unit cell has the following properties, in keeping with the symbols previously introduced: $d = 2.5$ mm, $w = 1.6$ mm (average), $s = 0.1$ mm, $g = 0.1$ mm, $w_L = 0.7$ mm, $g_c = 0.7$ mm, with the interdigitations each 0.1 mm wide and spaced 0.1 mm apart. The EBG is interfaced with a $h_l = 1.524$ -mm (60-mil) FR-4 ($\epsilon_l = 4.2$, $\tan \delta = 0.0016$) layer placed below the RO3010™ layer, which serves as a low-cost shielded substrate.

[00172] The EBG comprises three unit cells in the radial direction, and in its full radial form, employs 36 unit cells around the azimuth. The vias, which are used for both excitation and detection, and which were designed to be connected to 50- Ω , teflon-filled SMA connectors, are separated by 20 mm. The distance between the center of the excitation via and the inner radius of the EBG is 7.5 mm. It should be noted that the theory disclosed above allows these dimensions to be considerably reduced if necessary through various design choices such as using a smaller number of unit cells (at the expense of suppression ability), or using an etching process that could produce reduced feature sizes, which would allow for increased loading component values and hence a smaller period and/or increased bandgap width. These techniques would prove advantageous where space is limited, e.g., in systems with densely packed vias.

[00173] B. Simulation - Absorbing Boundaries

[00174] The case of an effectively unbounded PPW medium was investigated first in order to establish the realistic suppression ability of the EBG unobscured by multiple reflections that would be introduced by finiteness of the simulation domain. This was accomplished by placing absorbing boundaries around the edges of the finite PPW medium in simulation. The transmission response of this EBG, as measured through the upper layer RO3010™ dielectric, was simulated over a single radial section (as indicated in Fig. 18) with PMC transverse boundary conditions. The resulting scattering parameters are shown in Fig. 20, which reveal a suppression of around 20 dB over the 7.5 mm extent of the EBG.

[00175] The resulting scattering parameters of the complete EBG are shown in Figs. 21 and 22. Suppression across the X-band by up to 50 dB is observed when the EBG is present. Resonant interactions caused by coupling between the vias and the 2D EBG structure are observed at some frequencies (e.g., at 7.0 and 11.3 GHz). These resonant frequency points were investigated and found to depend on a number of factors, such as the radius of the exciting vias, the inner radius of the EBG, and the number of EBG unit cells employed radially – as such, they could likely be mitigated through a number of design choices which vary some or all of these parameters. Nevertheless, even with these resonances, the suppression maintains significant improvement over the case without the EBG at all frequencies.

[00176] Figures 23 and 24 examine the simulated fields respectively at 9 GHz (the frequency exhibiting maximum parallel-plate-mode suppression) and 5 GHz (outside the EBG bandgap). They detail the complex current-density magnitudes (plotted on an identical, logarithmic scale) on the metallization layer at the boundary between the RO3010™ and FR-4 dielectrics, which contains the EBG (the same layer shown in Fig. 9a). The excitation via is on the left-hand side surrounded by the EBG. At 9 GHz, the EBG effects a drop in the field level by approximately two orders of magnitude (40 dB), confirming the suppression suggested by the scattering parameters. The null between the EBG and the excitation via is evidence of the standing wave created by the signal being reflected by the EBG, and is noticeably absent at 5 GHz, where the EBG essentially completely transmits the (upper) PPW mode. From the corresponding cross-sectional complex magnitudes of the electric fields (calculated but not shown here) it can be seen that the field decay primarily takes place inside the EBG region as expected. There is some field leakage into the FR-4 layer, but it appears to be confined within the EBG region and is relatively small in magnitude (approximately 10 dB lower than the maximum fields in the RO3010™ dielectric). At 5 GHz, the fields are still constrained by the EBG (the currents must still pass through the thin CPW strips), but there is much less suppression as indicated by the field strengths over the outer-most unit cells. There is also slightly less leakage into the lower dielectric, indicating that the PPW mode in the upper dielectric is better guided by the EBG at this frequency.

[00177] Simulation - Open Boundaries

[00178] To enable comparison to a fabricated prototype, which would possess finite dimensions, the EBG was also simulated with open boundaries, with both layers of size 60 mm x 80 mm and embedded in vacuum. Figures 25 and 26 show the resulting simulated scattering parameters, for which up to approximately 40 dB of suppression and a corresponding improvement in return loss is observed over the frequency range of 7.5-11.5 GHz, corresponding to up to roughly 67 dB per guided wavelength of suppression. This is slightly lower, but comparable with reported results for Sievenpiper EBGs, which have had suppression of up to roughly 75 dB per guided wavelength for a similar two-layer setup, or roughly 95 dB per guided wavelength for a UC-EBG. Other suppression mechanisms such as high-dielectric-constant rodged photonic crystals have reported up to roughly 92 dB per guided wavelength, and circular high-impedance surfaces have been reported to obtain up to roughly 100 dB per guided wavelength. However, it should be recalled that our proposed unit cell is uniplanar, and therefore much easier to fabricate, as well as having a bandgap that may be accurately designed using MTL theory, both of which provide clear advantages over these other devices. The apparent noise in the resulting data is due to the fact that the open boundaries of the finite-sized PPW (which is electrically large at X-band) create reflections that establish a large number of 2D resonances (cavity modes).

[00179] D. Experiment

[00180] The PPW with EBG was fabricated by LPKF Laser & Electronics AG™ using a high-resolution laser-based PCB prototyping system, and is depicted in Fig. 27. The vias were realized with the use of flush-mount SMA connectors attached to the top face of the upper conductor, for which the center pin was clipped and soldered to the back side of the middle conductor.

[00181] The two layers (FR-4 and RO3010™) were compressed together using two clamps. The pressure was distributed with the use of a hard plastic spacer with a rectangular aperture (approximately 25 mm thick) and a layer of firm styrofoam (approximately 14 mm thick). A layer of masking tape was used to hold the two dielectrics together and prevent them from sliding laterally.

[00182] The measured results are indicated in Fig. 28 and 29. These data exhibit a frequency up-shift relative to the simulated case, which could be attributed to a slightly

lower ϵ_u resulting from substrate tolerances and possibly from a small air gap between layers (since the layers were not bonded by any means, but rather manually compressed together during measurement). It was found that if the simulation was re-run assuming an average dielectric constant of $\epsilon_u = 9.7$ and an average air gap of $50\mu\text{m}$ between layers, then the simulated and measured data sets matched reasonably well, as shown in the figures.

[00183] The proposed EBG could prove useful in mitigating the effects of PPW-mode excitation in several applications noted at the beginning of this work, including coupling reduction between adjacent substrate integrated waveguide (SIW) circuits, reduction of parasitic PPW coupling in conductor-backed aperture-coupled patch antennas, and even the design of miniaturized and/or multiband patch antennas as disclosed below. SW applications of the EBG are suggested by the observed coupling between the TM_0 surface-wave mode and the even modes of the S-CBCPW structure. This coupling may be exploited in many applications, including the mitigation of multipath interference in global navigation satellite system (GNSS) antennas, affording additional degrees of freedom in steering surface waves on surface-wave antennas and launchers, and generally in miniaturizing surface-wave components. Incidentally, the odd CSL mode could be used to couple with TE surface-wave modes, creating odd-mode bandgaps. Furthermore, the coupling between the even modes (odd modes) and the TM (TE) surface-wave modes results in dispersion features inside the light line, which could be used in combination towards the design of miniaturized dual-polarized or circularly polarized leaky-wave antennas.

[00184] Thus, a uniplanar EBG and MTL equivalent-circuit model have been proposed for the suppression of parallel-plate modes. This EBG is based on the TL-MTM and operates on the principle of contradirectional coupling between the one of the forward PPW modes and the backward CPW mode of a S-CBCPW structure, providing a large, controllable stopband. This allows for a uniplanar, printable design without any vias or discrete (surface-mount) elements, which makes it both low-cost and suitable for high-frequency applications. The dispersion of the supported modes was verified by full-wave simulation, and the transmission properties of the PPW mode were confirmed by simulation and in experiment. All of these results demonstrated very good agreement, validating the accuracy of the MTL equivalent-circuit model. The suppression of radially propagating

PPW-mode noise in multilayer PCBs between vias was suggested as a practical application and also validated in simulation and experiment. The 1D EBG was radially arranged around a via, and this was found to suppress the radially propagating PPW mode in two dimensions by approximately 50 dB at X-band over a length of just 7.5 mm.

[00185] A PRINTED DUAL-BAND ELECTROMAGNETIC BANDGAP
STRUCTURE GROUND PLANE FOR GNSS ANTENNAS

[00186] There is provided a novel ground plane design for GNSS antennas. The ground plane is a printed circuit board which contains electromagnetic bandgap structures which suppress the effects of multi path interference at two independent frequencies. This is accomplished by the suppression of surface waves which travel on such conductive ground planes, the end result of which is decreased left-handed circularly polarized receiver sensitivity above and near the horizon.

[00187] The novel structure introduced here is an EBG which operates at two distinct frequencies and is completely uniplanar without discrete (i.e., surface-mount) loading components, such that it can be fully printed using standard printed circuit board (PCB) manufacturing techniques. Furthermore, the design of this EBG is based on the transmission-line (TL) metamaterial, which can enable strong miniaturization of the EBG unit cells.

[00188] The developed EBG is based on the popular PCB waveguide, conductor-backed CPW (CBCPW), as shown in cross section in Fig. 1, or in top view in Fig. 5. Referring to Fig. 5, propagation is in the x-direction. As discussed above, it consists of four independent conductors, and therefore supports three quasi-TEM modes, as shown in Figs. 7 through 9, and with a three-dimensional view of the waveguide in Figs. 10 through 12. These modes are the parallel-plate waveguide (PPW) mode, the coplanar waveguide (CPW) mode, and the coupled slotline (CSL) mode. This structure also supports a SW mode on the three upper conductors. Fig. 5 details the properties of the upper conductor layer. The unit cell length is d , the CPW strip width is s , and the CPW gap width is g . The dielectric height (not labelled) will be referred to as h .

[00189] The bandgap properties of the EBG are obtained by *loading* this waveguide (referred to as the "host"), with capacitors and inductors.

[00190] Bandgaps are formed by creating the conditions under which modes exchange power. Metamaterial modes have the capacity to carry power in the opposite direction of their propagation, such that all of the power entering a metamaterial region can be guided in any desired direction, including the direction from which it arrived (reflection). The proposed EBG design exploits modal coupling to reflect surface waves, as well as couple them into the dielectric, in order to prevent them from reaching the antenna element. By coupling the PPW mode with the loaded CPW mode, a new dual-band mode is formed which in turn couples to the SW mode to form SW bandgaps. Importantly, both of these modes (PPW and CPW) are required to exist in order to create the dual-band functionality of this EBG, and therefore each of the four conductors is necessary for this dual-band design.

[00191] In the embodiment described here, a layout is used that has the features of the layout shown in Fig 2, but instead of a continuous center conductor of the upper conductor layer, center capacitors 116 are introduced to that conductor in parallel with the capacitors of the side conductors of the upper conductor layer. The capacitors are inserted into slots where the conductor has been removed from the CPW strip line (center conductor) and grounds (side conductors), whereas the inductors are inserted into the CPW gaps. These loading components can be realized in a manner compatible with a PCB process. An equivalent-circuit model of this loaded structure (for propagation along the x -direction) is shown in Fig. 31. The value L is the equivalent shunt loading inductance, the value C_s is the equivalent series loading capacitance in the CPW strip line, and the value C_g is the equivalent series loading capacitance in each of the two CPW grounds.

[00192] This equivalent-circuit model can be analyzed using a periodic analysis to yield the EBG's dispersive properties (phase shift across the length d as a function of frequency), from which the bandgaps in the SW mode can be readily observed. Fig. 32 shows such a numerically computed dispersion diagram, for which the SW mode on a solid conductor is represented by the solid black line. The dots represent points solved in the numerical simulator for the EBG structure, for which it can be seen that two "gaps" exist in which the black line does not overlap the dots. The gaps are the SW bandgaps created by this structure, and are a critical feature of the EBGs behaviour. The other regions that were found in the numerical simulator (indicated by the black dots) represent the coupling of the CPW

and PPW modes, which as previous stated are necessary for the formation of the SW bandgaps.

[00193] A prototype ground plane was designed for operation at GPS L1 and L2, using a RO3010™ dielectric (relative permittivity $\epsilon_r = 10.2$) with a thickness of 50 mils (1.27 mm). Referencing Fig. 5, the constituent unit cells were designed to have a length $d = 29\text{mm}$, width $w = 20\text{mm}$, CPW strip widths $= 2.5\text{mm}$, and CPW gap widths $g = 2\text{mm}$. The loading components L , C_s , and C_g are realized by strip inductors and interdigitated capacitors, both of which can be created in a standard PCB etching process, as shown in Fig. 33. The bandgap frequencies can be adjusted at this stage by tuning C_s and C_g . The dimensions of features of the interdigitated capacitors are labelled in Figs. 34 and 35. The CPW strip line capacitor has $N_{fs} = 7$ fingers, length $l_{sc} = 4.7\text{mm}$, finger spacing $g_{fs} = 0.2\text{mm}$, and finger width $W_{fs} = 0.2\text{mm}$. The CPW ground capacitors have properties $N_{fg} = 9$ fingers, length $l_{gc} = 0.6\text{mm}$, finger spacing $g_{fg} = 0.2\text{mm}$, and finger width $w_{fg} = 0.2\text{mm}$. The strip inductors have a width of $w_l = 1.0\text{mm}$.

[00194] Once this design is complete, three of these unit cells are cascaded, and distorted into a trapezoidal arrangement 118 by removing sections of the dielectric around the outer edges of the CPW grounds, as shown in Fig. 36. The trapezoid has an arc length of 12° , and an average width $w_a = 20\text{mm}$.

[00195] These trapezoidal sections can then be cascaded side-by-side into a closed circle 120, as shown in Fig. 37 (The four holes have been drilled through this dielectric to host coaxial feed lines for the antenna element (not shown)). The arc length of the trapezoidal sections determine how many unit cells can be arranged azimuthally to form a circle, and should therefore be an even factor of 360° (e.g., 12 degrees). The width of the first unit cell will then determine how large the diameter of the resulting circle will be - in this design, the inner diameter $c = 100\text{mm}$ and the outer diameter $e = 274\text{mm}$. The inside of the ground plane is a solid conductive layer. This final ground plane can be manufactured in its entirety using a standard PCB process on a single layer dielectric. The back side of the dielectric is simply a solid conductor.

[00196] This design was simulated along with a stacked patch antenna at these frequencies. It was compared with the original antenna element without an extended ground

plane, and an extended solid conductive ("bare") ground plane with the same radius as the EBG ground plane. The simulation results indicated that the LHCP is significantly suppressed on and above the horizon at both L1 and L2. The axial ratio (AR) is low and the multipath ratio (MPR- the ratio of RHCP at a point to the LHCP on the opposite side of the horizon) is increased near the horizon, as desired.

[00197] Dual-band Microstrip Patch Antenna Using Integrated Uniplanar Metamaterial-Based EBGs

[00198] There is provided a novel dual-band microstrip patch antenna that employs a metamaterial-based EBG (MTM-EBG) integrated into its radiating edges to create two distinct operating frequencies. The resulting antenna is compact, uniplanar, completely printable, and via-free. Dispersion engineering of the MTM-EBG unit cell through a rigorous MTL analysis allows easy design for two or more arbitrary frequencies. Additionally, a novel approach is taken to employ the same MTM-EBG to impedance-match the antenna to an inset microstrip feed at both operating frequencies. A dual-band MTM-EBG antenna designed to radiate at 2.4 GHz and 5.0 GHz is simulated and tested, and experimental results demonstrate radiation performance comparable to the corresponding conventional patch antennas in excellent agreement with simulations, while also affording some degree of miniaturization at lower frequencies.

[00199] Many approaches have been taken to yield dual- or multi-frequency operation of microstrip patch antennas, which may be expensive and/or difficult to implement. Early efforts introduced "stacked" patches, in which patches of different sizes are layered vertically with each underlying layer serving as the effective ground plane of the above patch, which may be directly or parasitically excited. A more simple arrangement involved parasitically exciting patches on the same layer allowing for a single-layer design; however, the parasitic coupling was found to be much less effective in this orientation. Exciting various cavity modes on asymmetric patches has been used; however, this technique inherently requires that the excited modes have different field profiles, polarizations, and possibly different feeding mechanisms, which may not always be desired. A popular current method of exciting various modes in a fully planar structure employs slots etched into the patch or ground plane, but such approaches tend to be empirical and are, therefore, ill-

equipped for systematic design. Some designs may employ loading with non-planar components such as vias, but these add to manufacturing complexity. Other antennas, particularly those employing frequency-dependent dispersive properties, achieve multi-band operation through the excitation of a number of different resonance mechanisms; however, these behaviours tend to come at the expense of gain and polarization purity. Moreover, the radiation patterns of these antennas do not typically resemble those of the fundamental patch mode for all radiating frequencies.

[00200] More recently, metamaterial (MTM) structures have been integrated into the design to produce multiple resonances. MTMs are artificial structures possessing properties that may transcend those typically found in nature.

[00201] A class of these materials known as transmission-line (TL) MTMs are particularly useful in engineering dispersion properties in TL environments such as microstrip or parallel-plate waveguide (PPW), created by appropriately loading a TL with discrete inductors and capacitors at deeply subwavelength intervals. Moreover, the dispersive properties of these structures can typically be accurately modelled with an equivalent circuit employing TL theory.

[00202] While the MTMs used in many of these works pose fabrication difficulties, such as the use of large numbers of vias, embodiments of TL MTMs disclosed in this document avoid these complications. For example, the 1D uniplanar electromagnetic bandgap (EBG) structure illustrated in Fig. 2, and with equivalent circuits CBCPW and S-CBCPW versions shown in Figs. 3 and 6 respectively, introduced and characterized for the suppression of PPW modes is ideal for PCB integration. This MTM-inspired EBG (or MTM-EBG for short) consists of electrically small unit cells and employs the contra-directional coupling between a PPW-like mode and a left-handed co-planar waveguide (CPW) mode to create a large bandgap that can be accurately described with multiconductor TL (MTL) theory. Therefore, the dispersive and bandgap properties of this material can be controlled and altered in a predictable manner, making it a suitable candidate for a wide range of microstrip and patch-antenna applications.

[00203] Here, the MTM-EBG is integrated directly into the metallization layer of a microstrip patch antenna 122, as shown in Fig. 38. It is demonstrated that the MTM-EBG

may be dispersion-engineered to present either bandgap or passband characteristics, which effectively modifies the electrical length of the patch as a function of frequency. This enables the patch's fundamental cavity mode to be excited at multiple different frequencies, such that all resonances possess the same polarization and radiation profiles. We present a MTM-EBG that enables dual-band operation of the patch by creating two different resonant patch lengths: one at a higher frequency, where the EBG operates in its bandgap so the fields are confined to the patch region, and another at a lower frequency where the EBG allows PPW propagation, and thus the patch operates as if it were electrically much longer. In addition, the antenna achieves miniaturization at the lower frequency due to the highly dispersive nature of the EBG's PPW mode. By implementing printable inductive strips and capacitive gaps into the MTM-EBG instead of using discrete (e.g., surface-mount) components, the full dual-band antenna can be printed in the patch metallization layer over a uniform ground plane, without the need for vias. Finally, owing to its accurate equivalent-circuit MTL model, the MTM-EBG can be carefully integrated to serve a second novel purpose: to provide a high degree of impedance matching at both of the antenna's resonances. Specifically, by adding a section of the MTM-EBG to the feed side of the antenna, the electrical length of the inset also changes with frequency. This approach ensures that both operating frequencies are well matched and produce gains comparable to those of a conventional patch antenna.

[00204] The proposed patch antenna makes use of the standard cavity model by replacing the perfect-magnetic-conductor (PMC) boundary conditions on the radiating edges with a frequency-dependent boundary condition realized by the MTM-EBG. Specifically, the MTM-EBG allows the TM_{10} mode to be supported both at what will be referred to as the lower frequency (located in its dispersive, low-frequency propagating band) and higher frequency (located in its higher-frequency bandgap), with the lower frequency being determined by a combination of the length of the cavity and the dispersive nature of the MTM-EBG, and the higher frequency being determined by the size of the cavity without the EBG. This method allows for the antenna to operate at two bands, which both maintain a simple excitation using a microstrip feed and the familiar radiative properties of the TM_{10} mode.

[00205] In order to effect the frequency-dependent behaviour described previously, a specific type of EBG is required. Firstly, it must be capable of interacting with the PPW mode, which shares the transverse field profile of the TM_{10} patch mode.

[00206] Secondly, its properties must be well known, since its response must be finely tuned in order to achieve the correct operating frequencies. Thirdly, the EBG should be uniplanar and via-less, as a simple antenna fabrication method is desired.

[00207] An EBG as disclosed above is such an EBG, and an MTL analysis was developed and shown to accurately describe its dispersion features. The host TL of the MTM-EBG is a conductor-backed co-planar waveguide (CBCPW), which supports the interaction of a PPW-like mode and a CPW mode, of which the latter is made to be left-handed (i.e., it supports backward-wave propagation) using left-handed MTM loading (series capacitors and shunt inductors, as shown in Fig. 39). The contra-directional coupling of these two modes results in a bandgap consisting of both complex-mode and evanescent frequency regions. Geometric features and loading values of the unit cell determine the position and size of the bandgap. A typical dispersion diagram is shown in Fig. 40, in which only the coupled system of even-polarized modes (PPW, CPW, and TM_0 surface-wave) is presented for clarity. Although other odd-polarized modes exist (e.g. coupled slotline), these shall be suppressed through the selection of a symmetrically placed feed. The bandgap region (shaded) describes complex and evanescent modes resulting from backward coupling, where propagation is strongly reactively attenuated.

[00208] The higher-frequency patch resonance is created by employing the EBG's PPW bandgap (from approximately 3.0 GHz to 5.5 GHz in Fig. 40), for which the PPW mode is prevented from propagating and the fields are strongly and reactively attenuated inside the EBG. For the lower-frequency resonance, the EBG's highly dispersive PPW mode is employed (below 3.0 GHz in Fig. 40). At these frequencies, the PPW mode still propagates, but incurs large phase shifts with distance travelled. Since the phase constant as a function of frequency is known, the resonant frequency is determined by noting that the dispersion angle of the EBG plus the phase incurred through the unloaded cavity must equal 180 degrees.

[00209] A unit cell was designed to realize the dispersion profile shown in Fig. 40, in which it can be seen that there is excellent agreement between the MTL theory and full-wave eigenmode simulations using Ansys HFSS. The substrate chosen for the unit cell was Rogers RO3003™ ($\epsilon_r = 3.00$, $\tan \delta = 0.0010$, 1.524 mm thick and clad with 17 μm copper on both sides) as it exhibits low-loss properties and is a common inexpensive substrate material used in microstrip and patch-antenna applications. The effective loading values of the unit cell are $L = 1.00$ nH and $C = 1.08$ pF as defined in Fig. 39, with dimensions listed in Table I. Note that L and C are defined differently in Fig. 39 in relation to this discussion of a patch antenna as compared to Fig. 2. Whereas the dispersion diagram assumes an infinite array of MTM-EBG unit cells, the MTM-EBG employed in the patch antenna will necessarily be limited to a small, finite number of unit cells. Fig. 42 compares the return and insertion losses through both a single cell and a cascade of three cells and confirms that even a single cell is capable of producing a bandgap, although it may be widened by using a larger number of cells. In fact, the bandgap edge frequencies are observed to approach the analytical prediction as the number of cells is further increased. To maximize compactness, the example antenna disclosed here employs a one-cell-long MTM-EBG, though in other embodiments more cells could be used. The bandgap region is more well-defined for the three-cell case but strong passband and bandgap regions are maintained for a single cell. Note that the band edges approach those predicted in the dispersion diagram (Fig. 40, shaded region in both figures) as the number of cells increases.

[00210] Lumped components are expensive, they complicate the fabrication procedure, and their maximum operating frequencies are constrained by their self-resonance frequencies (typically several GHz). To mitigate these drawbacks, the MTM-EBG cells can be designed for minimal reactive loading to enable the use of fully printed lumped elements, which allows for a unit cell that is completely printable as shown in Fig. 41, with dimensions provided in Table II. Propagation is in the x-direction. The inductance of the strip inductors can be estimated using empirical formulas found in J. K. A. Everard and K. K. M. Cheng, "High performance direct coupled bandpass filters on coplanar waveguide," IEEE Trans. Microw. Theory Tech., vol. 41, no. 9, pp. 1568–1573, 1993, and the design of interdigitated capacitors can be guided by G. D. Alley, "Interdigital capacitors and their application to

lumped-element microwave integrated circuits,” IEEE Trans. Microw. Theory Tech., vol. 18, no. 12, pp. 1028–1033, 1970. Thereafter parametric tuning is used to achieve the desired values. Additionally, the resonant frequency of the resulting antenna provides enough information to determine the effective values of the inductive and capacitive loading.

[00211] With an understanding of the properties of the MTM-EBG, the operation of a dual-band antenna can be considered. By applying a MTM-EBG to its radiating edges, a patch antenna is designed to operate at frequencies of 2.4 GHz (the “lower frequency”) and 5.0 GHz (the “higher frequency”). As a basis for comparison, corresponding conventional higher-frequency and lower-frequency patch antennas were designed with identical parameters l_p , w_p , and i_p (Fig. 38) chosen to achieve resonance at the above frequencies. The inset length i_p is important for impedance matching, as determined by the design equations in C. A. Balanis, *Antenna Theory: Analysis and Design*. Hoboken, New Jersey, USA: John Wiley & Sons, Inc., 2005. Note that, while the patch width was chosen to be the optimal width of the high-frequency antenna, the wider low-frequency patch width could have been chosen as well. Whereas this is an equally valid approach, it results in a high-frequency antenna that is much wider than it is long, and as a result, higher-order transverse modes may be excited. This leads to high levels of cross-polarization. Other patch widths may also be used. Fig. 46 shows an example antenna with a wider patch width than shown in Fig. 38, optimized for the lower frequency. Either width works sufficiently well, and other widths could also be used.

[00212]

	Conv. Low-Freq.	Conv. High-Freq.	MTM-EBG
l_p	35.15	16.67	16.80
w_p	43.32	21.21	21.00
i_p	10.40	4.80	4.50
w_i	2.00	2.00	1.62
w_f	3.76	3.76	3.76

[00213] TABLE I: Antenna design values in mm for the conventional patch antennas and the MTM-EBG antenna. Refer to Fig. 38 for parameter descriptions.

[00214]

	MTM-EBG, Front side	MTM-EBG, Feed side
l_c	5.90	5.50
w_c	7.00	7.00
s	0.20	0.20
g_L	0.80	0.80
g_C	1.70	1.60

[00215] TABLE II: Design values for the MTM-EBG antenna's two MTM-EBG regions in mm. Refer to Fig. 41 for parameter descriptions.

[00216] In general, inset length decreases when matching antennas at higher frequencies, and this observation suggests a novel application for the MTM-EBG: to integrate it into the feed side of the antenna for a high degree of impedance matching at both operating frequencies. The difference in inset length can be expressed as an electrical length, and then a MTM-EBG cell can be designed from its dispersion diagram to have this electrical length at the lower frequency, while still reflecting at the higher frequency. The effect of this is that the inset length is frequency-dependent, and is optimized for each resonance.

[00217] The remaining electrical length needed for a resonance at the lower frequency can be easily calculated, and another EBG cell with appropriate dispersion characteristics is affixed on the front end of the patch. The MTM-EBG antenna should now resemble each individual patch at its respective operating frequency, and after simulation only minimal tuning (generally in patch length or MTM-EBG unit-cell length) should be needed to ensure the operating bands are at the desired frequencies.

[00218] The result of this process, as described below, is a well-matched antenna that operates in a standard patch TM_{10} mode for both frequencies. Additionally, due to the highly dispersive nature of the MTM-EBG, the patch is moderately miniaturized at the lower

frequency, with respect to the conventional low-frequency patch. The uniplanar MTM-EBG is shown to be ideal for this application, and the result is a practical, fully printed, and analytically designable dual-band microstrip patch antenna.

[00219] Following the design procedure given above, the dual-band MTM-EBG antenna as well as conventional lower- and higher-frequency patch antennas were simulated and fabricated on a Rogers RO3003™ substrate. Dimensions of the antennas are given in Tables I and II.

[00220] Looking first at return loss shown in Figs. 43A-43D, it is evident that the MTM-EBG antenna is well matched, exhibiting a return loss better than 10 dB in all cases. The operating frequency is shifted up for the higher-frequency resonance by about 3%, though a 1% shift was also seen in the corresponding conventional antenna. This may be attributed to fabrication tolerances in the construction of the antenna features, which are particularly sensitive at higher frequencies. A final note on the return loss is that the low-frequency resonance of the MTM-EBG antenna is slightly more narrowband than that of the corresponding conventional patch; this is to be expected since, due to the dispersive properties of the MTM in its low-frequency transmitting region, a small change in frequency results in a correspondingly large change in electrical length.

[00221] Simulation provides further validation of the design concept by producing plots of the complex magnitude of the electric fields in the patch at the operating frequencies of the antennas. The simulated fields were plotted through the center of the dielectric, and it was found that the MTM-EBG regions behave precisely as desired, since for the low-frequency resonance they are clearly transmitting, yet for the high-frequency case the fields are confined to the patch region. In addition, each resonance strongly excites the fundamental TM_{10} patch mode which implies that the radiation patterns of corresponding frequencies should resemble each other fairly closely.

[00222] The gains of the simulated antennas are compared in Fig. 44A-D. Viewing the MTM-EBG antenna in direct comparison with conventional patch antennas at the same frequency, it is apparent that the radiation patterns are very similar in terms of both co- and cross-polarizations. While the gain of the MTM-EBG patch antenna is almost identical to that of the conventional patch at the higher frequency, it is approximately 2.5 dB lower than

conventional patch at the lower frequency. While the MTM-EBG patch antenna still remains an effective radiator at this frequency, the degradation in performance is likely due to small scattering losses in the fine features of the MTM-EBG and to the decrease in electrical size of the antenna.

[00223] The directivities of both the conventional and MTM-EBG patch antennas were measured in an anechoic chamber for comparison to the simulation results. Antenna measurements initially showed high cross-polarization and a combination of simulations and rigorous measurements suggested that this was due to radiation from unbalanced currents along the coaxial feedline. A solution to this for the lower-frequency resonances was to attach ferrite beads to the feedline; this directly contributed to a reduction in cross-polarization. The ferrite beads were not rated for 5 GHz but when attached very close to the antenna still contributed to the reduction of cross-polarization, if to a lesser extent than at the lower frequency. Therefore they were included in all measurements of the antennas.

[00224] Overall, simulation and measurement of antenna directivities exhibit excellent agreement, as shown in Fig. 45A-H, and many of the finer details such as small lobes and nulls are present in both cases. There are only two notable disagreements; the first is a null in the back-fire direction of every antenna which is attributed to blockage due to the metallic mounting apparatus. The other major disagreement is seen in the back lobes of the H-planes for the higher frequency resonances, where measurement shows significantly higher gain than simulation. Since both the conventional and MTM-EBG antennas exhibit this behaviour, it could be a result of the continued existence of unbalanced currents on the feed line, which maintain a presence due to the less-than-optimal performance of ferrite beads at this frequency. Overall however, the measurements both successfully verify the simulation results and confirm that the MTM-EBG antenna produces patterns that are very similar to those of the corresponding conventional patch antennas; in other words, the presence of an MTM-EBG that is either transmitting or in its bandgap region does not significantly affect the radiation pattern of the antenna.

[00225] In a further embodiment, multi-band antennas could be created by appending multiple unique MTM-EBG rows onto the radiating edge of the patch. Furthermore, if the MTM-EBG cells can be miniaturized further broadband behaviour may be achieved by

pushing the multiple resonances close together. In fact, even moderate improvements in the bandwidths of the dual-band MTM-EBG antenna described in this work would suit it ideally to WLAN applications. Additionally, dual- or multi-band circularly polarized antennas may be produced through inclusion of MTM-EBGs along both axes of the patch and with different phase responses, with the additional advantage of impedance matching at all frequencies. Multiple adjacent structures with individually selected parameters may be used to effect multiple operating frequencies. Fig. 47 shows a rectangular patch antenna employing different uniplanar MTM-EBGs perpendicular to both patch axes, thereby allowing independent phase and amplitude control of orthogonal modes. As shown in Fig. 47, and embodiment of a patch antenna 122 has a first set of MTM-EBGs 124 at a first pair of opposing edges and a second set of MTM-EBGs at a second pair of opposing edges. In the embodiment shown, the capacitors 104 in first set of MTM-EBGs 124 have smaller gaps than capacitors 104 in second set of MTM-EBGs 126, and inductors 102 in the first set of MTM-EBGs 124 have smaller widths than the inductors 102 in second set of MTM-EBGs 126, in order to cause different electromagnetic properties between the sets of MTM-EBGs. Any other relevant parameters could also be changed. It is not necessary for opposing sides to have the same design of EBG as shown in Fig. 47. Different designs could be used on opposing sides, or in an embodiment only one side of a pair of opposing sides may have an EBG. More specifically, rectangular antennas could be made with EBGs arranged at 1, 2, 3 or 4 sides of the rectangle, and in the case of 2 sides the sides could be adjacent or opposing. Dual band circularly polarized antennas such as the embodiment shown in Fig. 47 would be useful in applications such as satellite-assisted positioning. Overall, the versatility of the MTM-EBG cell and accurate design procedure allow for a wide variety of inexpensive, multi-band antennas to be easily designed and realized.

[00226] Fig. 48 shows a circular patch antenna 128 employing a uniplanar MTM-EBG 130 on its radiating edge. This figure extends the concept of MTM-EBG loading of the rectangular patch antenna to a circular patch antenna, which is particularly useful for applications requiring dual-band circular polarization, e.g. GPS antennas, and has the advantage that it is uniplanar.

[00227] The large PPW-mode bandgap presented by the MTM-EBG allows the embodiment of an antenna tested to possess one operating frequency in each of the bandgap and passband regions, with the further advantage of impedance matching at both frequencies. The embodiment of an antenna is uniplanar, compact, printable, well-matched, and easy to design. Simulated and experimental results confirm this theory and exhibit excellent agreement, suggesting that the MTM-EBG antenna performs very comparably to conventional patch antennas at both of its resonances. Further embodiments may provide low-cost, low-profile, multi-band, circularly polarized, and/or multi-/wide-band MTM-EBG-based antennas, that could be tuned to any desired frequencies, and which are easy to manufacture.

[00228] Application #1: Tuneable Devices

[00229] The MTM-EBG is engineered to possess one or more frequency bandgaps (frequency ranges not permitting wave propagation, akin to filtering). Since the frequency ranges of these bandgaps may be controlled by the design of the MTM-EBG, and specifically by varying the lumped loading elements that constitute it, the filtering effect can be made tunable. For example, the capacitive loading elements may be replaced by varactors (essentially variable capacitors). Other methods of tuning include the use of ferroelectric components or liquid crystals that may be tuned using an external bias field. This tunability could realize new functionalities in other microwave components containing MTM-EBGs. An example is provided below.

[00230] Often in microwave networks, it is necessary to branch out a single transmission line into two or more branches. However, this introduces an impedance-matching problem, because the input line sees the combination of the impedances looking into the output lines. To enable matching, a matching network known as a quarter-wavelength transformer is used. As the name suggests, this is an intermediate transmission line of a length equal to a quarter wavelength, which possesses a characteristic impedance equal to the geometrical average of the impedance of the input line and the combined impedances of the output lines. When realized using microstrip technology, this implies a trace width that is wider than the input and output lines. Because its operation is tied to electrical length (of one-quarter of a wavelength), this devices only works at a specific

frequency (and odd harmonics of this frequency). However, by embedding a MTM-EBG into the device as shown in Fig. 49 to provide the same quarter-wavelength phase (90 degrees) at two different frequencies separated by a bandgap, we achieve dual-band functionality as well as strong filtering capability between these two frequencies (due to the bandgap). See Fig. 50. The benefit of our approach is that the MTM-EBG occupies no more space than the original microstrip transformer (and can often be miniaturized at the lower of the two operating frequencies) and so we refer to it as an “embedded MTM-EBG”.

[00231] Applications:

[00232] 1. Dual-Band embedded matching networks: The MTM-EBG is used as a substitute to a conventional quarter-wavelength impedance transformer. The cell is designed such that the microstrip mode it supports has identical characteristic impedance to the quarter wavelength transformer. See Fig. 49. The phase of the cell can then be designed at each of the desired frequencies to simultaneously improve the response of the network and reject undesired frequencies. The capacitive and inductive elements can be realized using either discrete lumped elements or fully printed.

[00233] 2. Variable embedded MTM-EBG: An alternate cell design can be embedded into the desired microstrip network with minimal disruption outside of the bandgap. By replacing the printed or discrete lumped capacitor with a tunable element such as a varactor, the bandgap location can be tuned by an external bias network as in Fig. 51. Initial results indicate the bandgap location can be tuned over a wide range of frequencies with small variations in the varactor capacitance (see Fig. 52).

[00234] Tunable filtering is of great use in industry. One feature which is anticipated in 5G specifications is the ability to actively tune to “whitespots”, where no frequency transmission is occurring, or actively tuning to remove interference from alternate frequency sources. Technologies which can be actively tuned exist, but may require complex 3D structures and aren’t well suited to small devices, or are far more complex than the proposed technology. There may be limits on the frequency ratio for dual-band applications. A cascaded MTM-EBG may be used for more broadband passbands.

[00235] Application #2: Sensors

[00236] It is proposed that the MTM-EBG structure may be used as a material or strain sensor. The MTM-EBG can be designed to have a very narrowband (resonant) response, the deviation of which may be sensed with basic processing electronics to determine changes in the electromagnetic properties of the surrounding medium, or strain-induced mechanical changes to the MTM-EBG itself.

[00237]

[00238] Near-Field Material Sensing

[00239] Because the strongest fields supported by the MTM-EBG structure are in the vicinity of the loading components, these areas are most sensitive to changes in the surrounding medium. Specifically, the resonant frequency will shift in response to changes in the permittivity of the material in close proximity to the capacitive gaps, as well as changes in the permeability of the material in close proximity to the inductive strips. By measuring the resonant frequency, the electromagnetic properties of the materials may be determined. An example of the MTM-EBG structure operating in this manner is shown in Fig. 53A, in which the signal is extracted from the bottom left. The simulated resonant frequencies with and without a magnetic test sample is shown in Fig. 53B, where the magnitude of the reflected signal is shown without the sample (red curve) and with the sample (blue curve). The shift in frequency can be correlated with the material properties of the test sample.

[00240] Strain Sensor

[00241] A similar MTM-EBG structure may be printed on a flexible/malleable substrate, which when placed under tension or deformed, would also slightly shift its resonant frequency. In this way, strain inside of a solid may also be sensed by tracking this frequency.

[00242] Temperature Sensor

[00243] Similarly, a temperature sensor may be developed through the inclusion of temperature-dependent loading components, such as Barium-Strontium-Titanate (BST)-based capacitors or substrate materials, for which the associated resonant frequency would vary with temperature.

[00244] Sensor Array

[00245] Arrays of such sensors may be utilized to determine the spatial variation of the previously proposed properties, such as: material permittivity or permeability, strain, or temperature in two dimensions. By slightly detuning each element to a known frequency, deviations from the nominal response may be sensed electronically and mapped to a unique position in the array. For example, the cells highlighted in Fig. 54A may be uniquely mapped via the corresponding colors to the resonant frequencies shown in Fig. 54B. In this manner, a two-dimensional “image” may be generated by the array. Provided that the individual cells are deeply subwavelength, it is possible to use this approach to perform subdiffraction imaging merely by observing the far-field transmission and/or reflection spectra.

[00246] Compact, inexpensive sensors are currently of great interest to many industries. Many modern “smart” devices are equipped with basic sensors, and the “internet of things” promises to integrate sensors for a wide variety of purposes. Sensor arrays may be found in industrial and commercial manufacturing and processing. Multiple sensors may be combined to create a sensor array.

[00247] Application #3: Planar Resonators

[00248] Multiple MTM-EBG structures may be cascaded to create very high quality factor planar resonators, which are traditionally fabricated using substrate integrated waveguide (SIW) technology. It is known that the variant of the MTM-EBG with a capacitor in the center conductor (in addition to the outer two) possesses a very resonant and low-bandwidth propagation mode. This may be utilized to replace traditional rectangular SIW and microstrip ring resonators with MTM-EBG equivalents, as shown in Figs. 55A and 55B, which would have the advantage of being fully uniplanar, and even possess a higher quality factor than the original devices. Fig. 55A shows a rectangular cavity resonator to replace a SIW cavity, and 55B shows a MTM-EBG based ring resonator.

[00249] More compact and easy-to-manufacture planar resonators are of interest to integrated circuit manufacturers, where cost and fabrication process simplicity are key factors that are constantly being driven to improve.

[00250] Application #4: Beamforming and Frequency Selective Surfaces

[00251] According to the theory of images, a solid conductor such as the conductor backing of the MTM-EBG acts as a mirror, to virtually replicate any electromagnetic fields

and objects in front of it as if they are equivalently spaced behind it. Under such a viewpoint, the MTM-EBG (shown in Fig. 56A) may be seen as a system of back-to-back coplanar waveguide transmission lines, each loaded with series capacitors and shunt inductors, on opposite side of a dielectric (as shown in Fig. 56B).

[00252] It has been found that arrays of this structure can support high transmission/reflection ratios, as shown in Fig. 57A, while also possessing the ability to re-direct incident plane waves to a different angle. This behavior is shown in Fig. 57B, in which the incident and transmitted angles are indicated with the black arrows.

[00253] It may also be possible that the structure may be useful for manipulating the phase of the applied signal, such that the MTM-EBG may also be used for canonical frequency-selective-surface applications, such as beamforming, transmission control, polarization control, etc.

[00254] This beam steering behavior is extremely useful for antenna radomes, since a radome could be internally or externally covered with an inhomogeneous array of these modified MTM-EBG structures, each of which is designed to direct a normally incident beam in a desired direction. This allows the already existing radomes to assist with beamsteering, increasing the gain and/or selectivity of the underlying antenna. Such a technology would be useful to any antenna designers in the communications, imaging, and defense industries.

[00255] This structure may be modeled using a rigorous multiconductor transmission-line model similar to the one already validated for the MTM-EBG structure, which shall generate straightforward equations for calculating the redirection angles given by the unit-cell.

[00256] Application #5: Absorbers

[00257] The MTM-EBG's dielectric substrate is typically desired to be low-loss, however, it may be replaced with a lossy material such as carbon-impregnated styrofoam such that it dissipates accepted power. It has been demonstrated that the MTM-EBG accepts incident radiation at a designable frequency, allowing it to function as an absorber. This is shown below in the resonances of Fig. 53B, which shows that there is a drop in reflected power (and hence absorption) at a particular frequency.

[00258] Absorbers are of strong interest in the fields of microwave measurements and defense. Since the MTM-EBG is lightweight and low-cost, it could present a strong candidate to replace existing technologies such as volumetric absorber cones or components with surface-mount chip components.

[00259] Application #6: Multiband Tuning Stubs

[00260] Microstrip stubs are common in low-cost microwave systems since they are simple to design and fabricate and take up a relatively small amount of circuit board space. They can be employed as band-stop or band-pass filters, or more generally used for impedance matching to arbitrary loads. While they tend to be narrowband, they are common due to their simplicity and effectiveness, and as a result wide- or multi-band stubs would have many further applications in PCB-based devices.

[00261] Since dual-band behaviour has been demonstrated by integrating MTM-EBG unit cells onto the radiating edges of a patch antenna, a similar technique can be applied to microstrip stubs. MTM-EBGs can be appended onto the open-circuited end of a stub to enforce frequency dependent operation, such that the stub appears to be shorter where the MTM-EBG is in its bandgap region, and longer when it is transmitting. A second MTM-EBG cell can be appended onto the end of the first to create another distinct electrical length when it is transmitting, and any additional number of unit cells can further be added to produce additional bands.

[00262] Such a design could be used to provide any desired reactance at an arbitrary number of frequencies, but to demonstrate the operating principles of this structure, a triple-band, bandstop filter, shown in Fig. 58, are being investigated. The stub will enforce a bandgap at its quarter wavelength resonances, so at each operating frequency, the stub must electrically appear to be this length. Operating frequencies of 2.4 GHz, 3.6 GHz, and 5 GHz are chosen. The MTM-EBGs can produce the three required lengths through dispersion engineering of their bandgap and phase properties. If the stub alone is a quarter wave length at 5 GHz, MTM-EBG A must have a bandgap at this frequency, and a dispersion angle such that the stub plus MTM-EBG electrical length is a quarter-wave at 3.6 GHz. MTM-EBG B must then have a bandgap at 3.6 GHz, with the combined electrical length of the entire stub a quarter-wave at 2.4 GHz. Following this procedure, the MTM-EBG unit cells were designed

(with fully printable interdigitated capacitors and strip inductors having minimum feature size of 50 μm , although surface-mounted elements could be used instead), and the results are shown in Fig. 59. The three bands are clearly displayed with strong insertion loss of over 20 dB; measurements corroborate the data, with a slight frequency shift caused by minor overmilling of the capacitors. The final design is compact, uniplanar, tri-band, and has higher selectivity than conventional stubs.

[00263] There are many possible applications of the multi-band stub beyond the demonstrated bandstop filter. The simplest of these is switching the design for bandpass operation, or even a combination of both filter types; or, rather than relying on resonant stub lengths, each length could be designed for a specific reactance, which would be useful for multi-band matching to an arbitrary complex load. Operation could also be modified to meet bandwidth specifications; selectivity could be increased by operating in more dispersive regions of the MTM-EBG's passband, or bandwidth increased by using a radial stub design. Many existing technologies could be miniaturized or improved through the use of such MTM-EBG-based designs since microstrip stubs are so widely employed in a variety of device and systems.

[00264] A hybrid coupler is now described with reference to Figs. 60-63.

[00265] The novel MTM-EBG structure presented in this work utilizes a host transmission line (TL) created though inserting two gaps of equal width along the length of a microstrip, a cross-section of which is shown in Fig. 60, such that the microstrip is transformed into a truncated conductor-backed CPW. The reactively loaded structure as described above presents a bandgap to the MS-like mode, which causes the useful dispersion feature that two frequencies in close proximity to the bandgap (one above the upper frequency bound and one below the lower) possess the same phase angle over each unit cell, as indicated with the dashed vertical line in Fig. 61, where the upper and lower frequencies have been chosen to be the GPS L1 and L2 bands. A smaller bandgap, necessary for operation at relatively close frequencies, is enabled by placing the series loading capacitance in the CPW center strip rather than the grounds. These design features allows the MTM-EBG structure to be integrated into a single MS TL, and present a similar, predictable, and designable bandgap to the MS mode.

[00266] Fig. 3 shows an example of a layout of the proposed couplers. Since the coupler depends on a 90 degree phase shift along its arms, it is sufficient to employ one unit cell (although additional cells may be employed to achieve greater operating bandwidths, at the expense of additional circuit area), and utilize the two dispersion frequencies that correspond to 90 degrees. While the upper frequency will always possess a higher phase velocity than that of the MS mode, the loss due to leakage is minimized through employing only one unit cell. Furthermore, since the lower frequency is very similar to that of the non-dispersive MS mode, the physical length of the unit cell should be very close to that of a quarter-wavelength MS TL at the lower frequency. Table I shows an example of coupler geometry.

TABLE I
PROPOSED COUPLER GEOMETRIC PARAMETERS

M	23.20 mm	
W	27.80 mm	
L	23.05 mm	
	50 Ω Branch	35 Ω Branch
l	20.40 mm	19.60 mm
w	1.30 mm	2.50 mm
s	0.30 mm	0.40 mm
g	0.20 mm	0.40 mm

[00267] In a simulation, the branches of the coupler were designed through eigen-mode analysis in HFSS. Using MS TLs of typical lengths and widths as the starting point to design the unit cells on a 1.524 mm-thick RO-3010 substrate ($\epsilon_r \sim 10.2$) with 35 μm copper thickness, the geometric values of the completed coupler are shown in Table I and correspond to the layout in Fig. 62. The resulting simulated dispersions of the two branches – in which loading capacitance values of 1.1 pF for the 50 Ω MS and 1.3 pF for the 35 Ω MS were used – are shown in Fig. 2, which confirms desired dual-band operation.

[00268] The scattering parameters of the final simulated coupler are shown in Fig. 63, where it can be observed that there are slightly higher insertion losses at the upper frequency (4.3 and 4.3 dB) than at the lower frequency (3.2 and 3.5 dB); however, the circuit still performs with acceptable losses over both bands. The phase difference between the two

output ports 2 and 3 is 90 degrees at the lower frequency and 85 degrees at the higher frequency.

[00269] The coupler was then fabricated and the 1005 (metric) capacitors manually soldered in place. The experimentally determined scattering parameters are also shown in Fig. 63, in which it can be seen that there is qualitative agreement with the simulated scattering, with a small frequency shift contributing to insertion losses at the higher frequency of 6.5 dB (S21) and 5.8 dB (S31), and 2.9 dB (S21) and 3.9 dB (S31) at the lower frequency. The phase difference between the two output ports 2 and 3 is very good with 89 degrees at the lower frequency and 90 degrees at the higher frequency. The observed discrepancies between simulated and measured data sets are most likely due to the placement and soldering of the capacitors, which was done manually; as well as the fact that a simplified capacitor model was used in simulation.

[00270] A novel dual-band quadrature hybrid coupler was designed and experimentally validated for operation at both GPS L1 and L2 frequencies. The dual-band functionality was achieved through embedding MTM-EBG unit cells into the host microstrip lines, which provide a given phase shift at two different, and designable, operating frequencies.

[00271] Immaterial modifications may be made to the embodiments described here without departing from what is covered by the claims. In the claims, the word “comprising” is used in its inclusive sense and does not exclude other elements being present. The indefinite articles “a” and “an” before a claim feature do not exclude more than one of the feature being present. Each one of the individual features described here may be used in one or more embodiments and is not, by virtue only of being described here, to be construed as essential to all embodiments as defined by the claims.

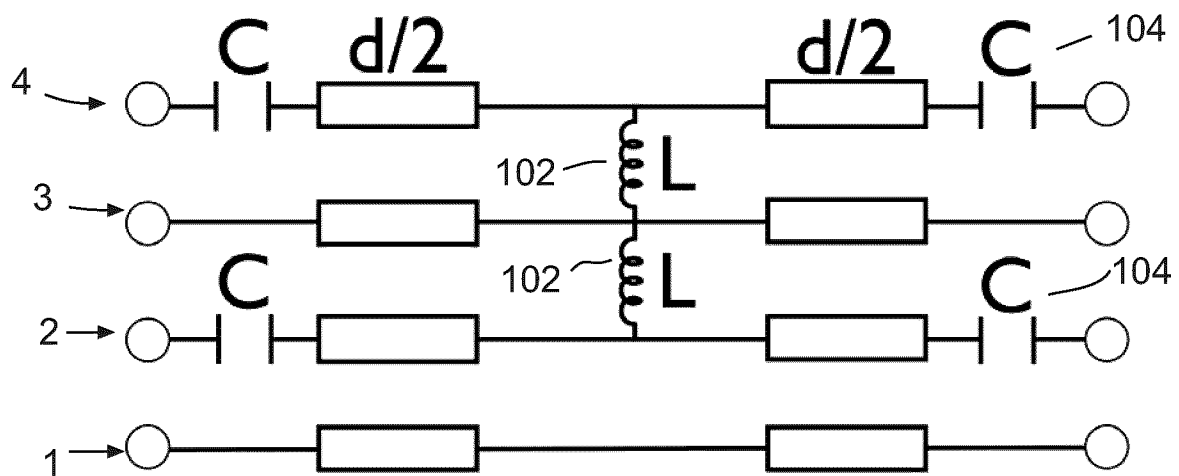
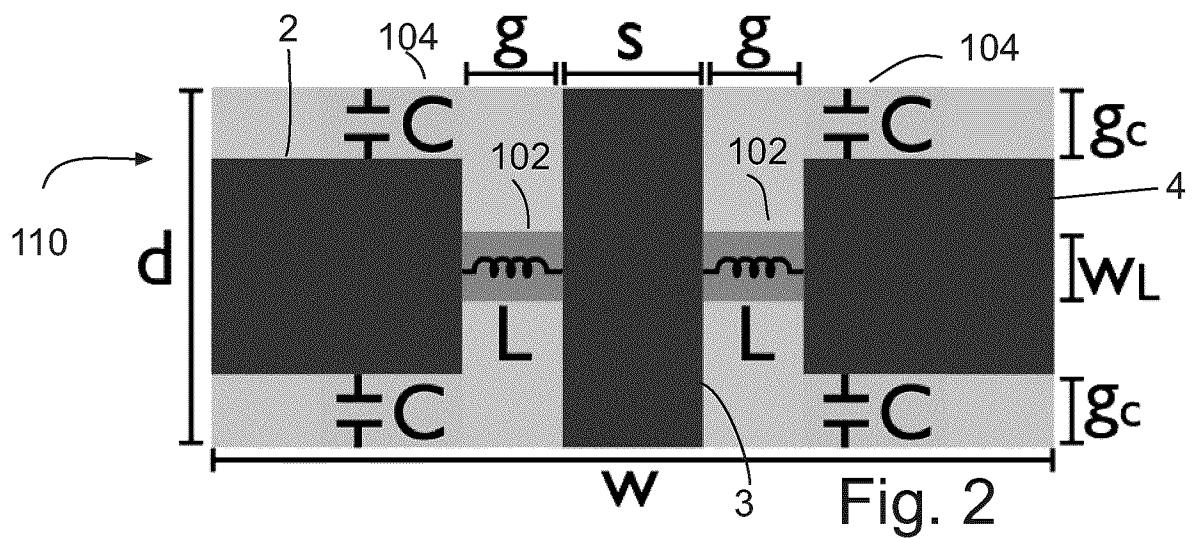
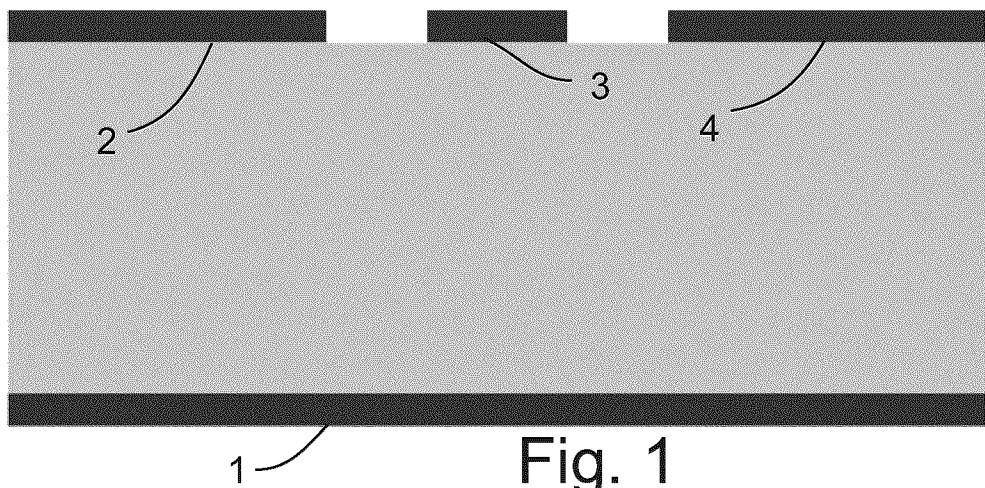
THE EMBODIMENTS OF THE INVENTION IN WHICH AN EXCLUSIVE PROPERTY OR PRIVILEGE IS CLAIMED ARE DEFINED AS FOLLOWS:

1. An electromagnetic structure comprising:
a conductor-backed coplanar waveguide structure including a central line, two side lines and a conductor backing;
the conductor-backed coplanar waveguide structure being loaded with one or more capacitors on the side lines or on the central line and inductors connecting the side lines to the central line, the capacitors and inductors having capacitance and inductance selected to cause a frequency-dependent coupling between a parallel-plate waveguide mode and a coplanar waveguide mode to form an electromagnetic bandgap.
2. The electromagnetic structure of claim 1 further comprising a shielding plane, the conductor-backed coplanar waveguide structure and the shielding plane together forming a shielded conductor-backed coplanar waveguide structure.
3. The electromagnetic structure of claim 1 or claim 2 in which the one or more capacitors are on the side lines only.
4. The electromagnetic structure of claim 1 or claim 2 in which the one or more capacitors are on the central line only.
5. The electromagnetic structure of claim 1 or claim 2 in which the capacitors are on both the side line and the central line.
6. The electromagnetic structure of any one of claims 1-5 in which the inductors and capacitors are formed of a uniplanar conductive layer.
7. The electromagnetic structure of claim 6 in which the inductors and capacitors are formed by printing.

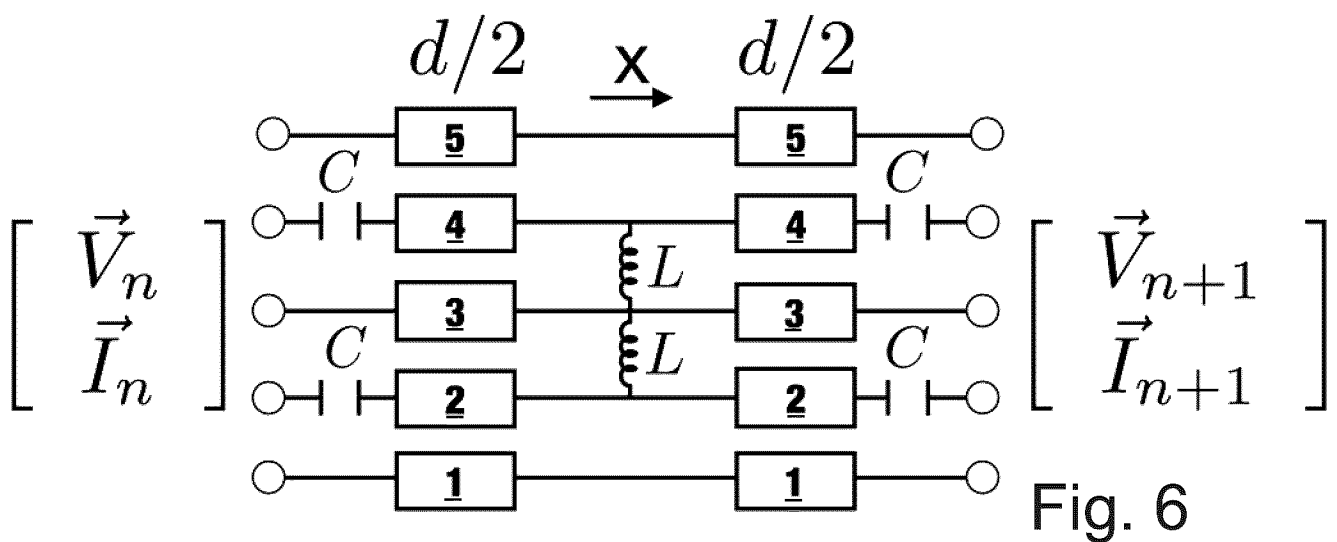
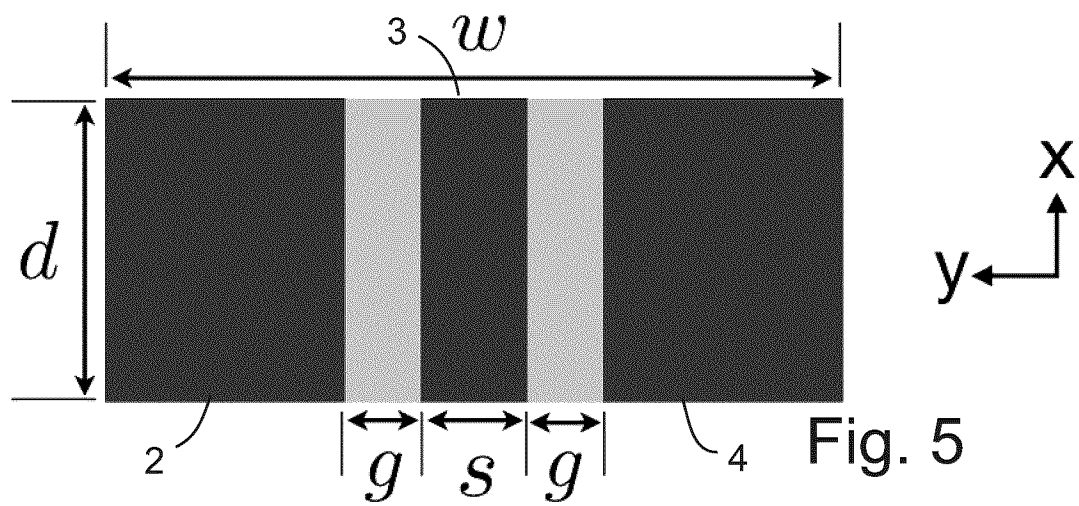
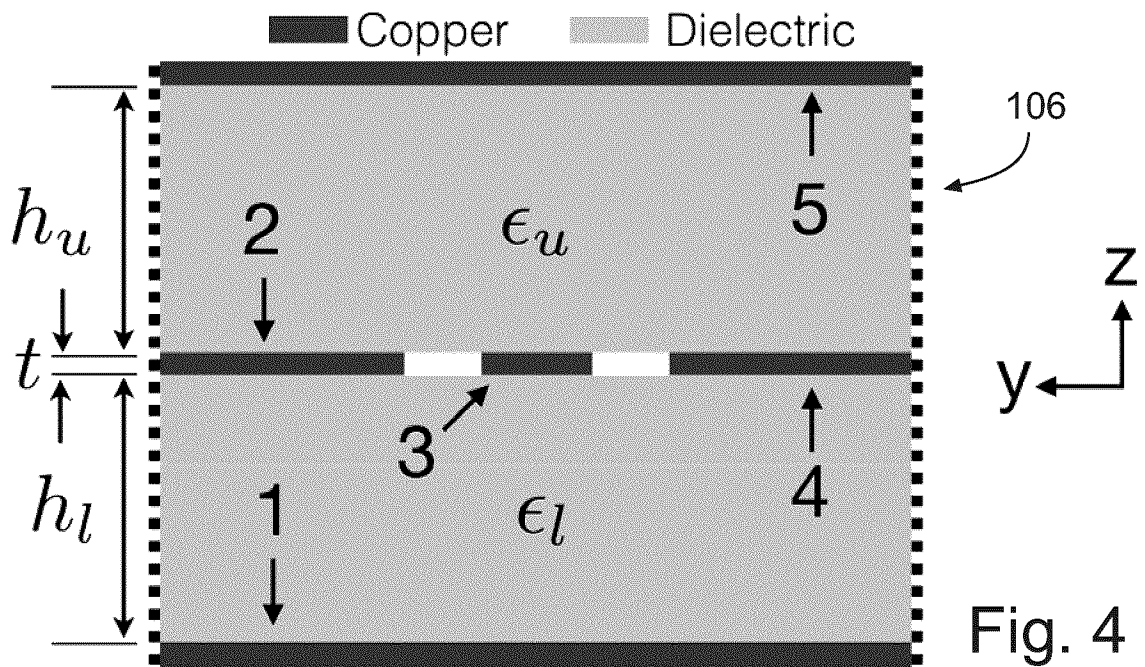
8. The electromagnetic structure of claim 6 or claim 7 in which the central line and side lines are formed of the uniplanar conductive layer by printing.
9. An enclosure structure comprising plural electromagnetic structures as claimed in any one of claims 1-8 arranged in parallel to surround a central area.
10. The enclosure structure of claim 9 arranged around a via to suppress parallel-plate noise.
11. The enclosure structure of claim 9 arranged around a ground plane for an antenna to suppress surface waves in the ground plane.
12. A hybrid coupler having four branches and four ports, and each branch comprises a structure formed according to any one of claims 1-8.
13. A patch antenna comprising a patch and electromagnetic structures as claimed in any one of claims 1-8, the electromagnetic structures arranged at edge portions of the patch.
14. The patch antenna of claim 13 in which the electromagnetic structures have different values of design parameters from each other in order to effect multiple operating frequencies.
15. The patch antenna of claim 13 or claim 14 in which the patch generally defines a rectangle and the electromagnetic structures are arranged at one side of the rectangle.
16. The patch antenna of claim 13 or claim 14 in which the patch generally defines a rectangle and the electromagnetic structures are arranged at two opposing sides of the rectangle.

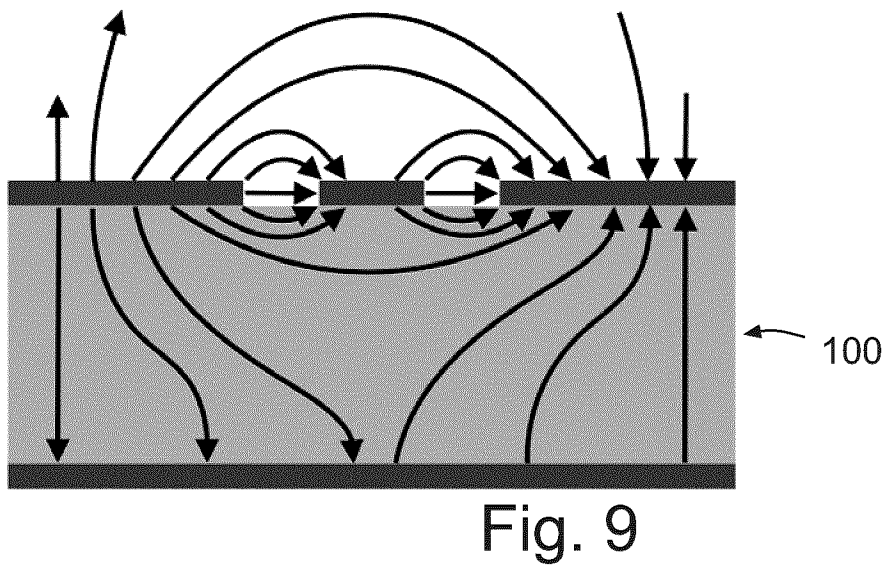
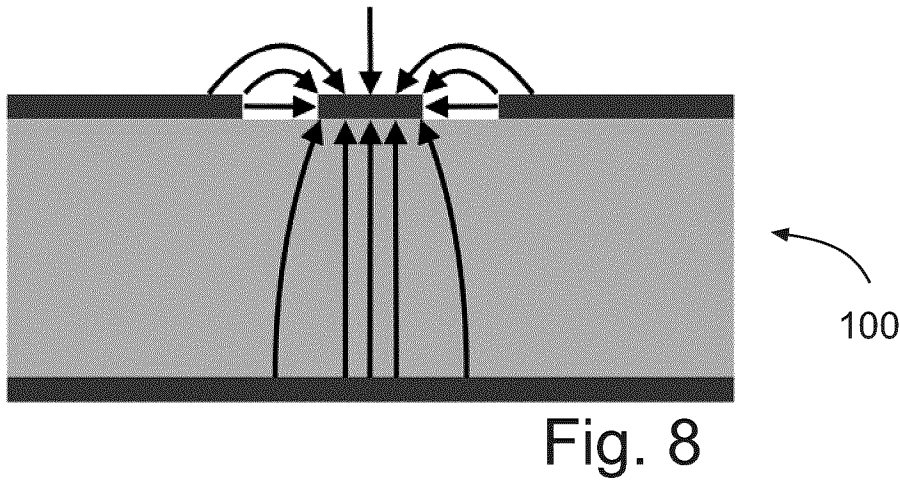
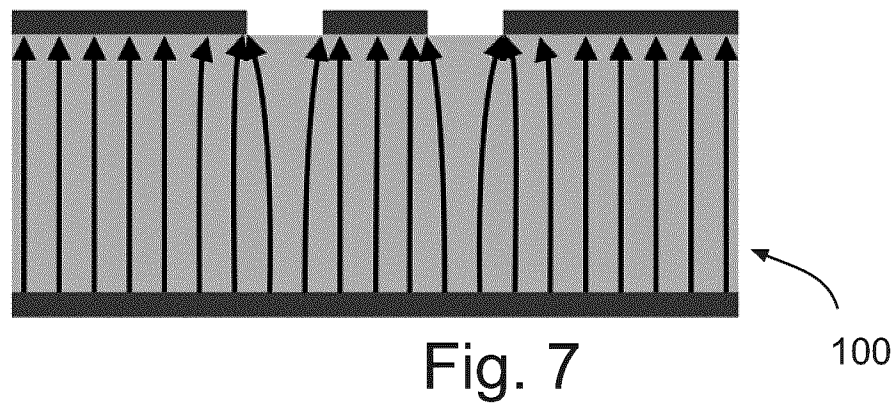
17. The patch antenna of claim 13 or claim 14 in which the patch generally defines a rectangle and the electromagnetic structures are arranged at two adjacent sides of the rectangle.
18. The patch antenna of claim 13 or claim 14 in which the patch generally defines a rectangle and the electromagnetic structures are arranged at three sides of the rectangle.
19. The patch antenna of claim 13 or claim 14 in which the patch generally defines a rectangle and the electromagnetic structures are arranged at all sides of the rectangle.
20. The patch antenna of claim 13 or claim 14 in which the patch is generally circular and the electromagnetic structures are arranged around a perimeter of the patch.
21. A dual band power divider having an input line and a pair of output lines, in which a structure according to any one of claims 1-8 is formed at the junction of the input line and the pair of output lines.
22. The dual band power divider of claim 21 in which the structure has characteristic impedance equal to that of a quarter wavelength transformer designed to enable impedance matching at the input.
23. The dual band power divider of claim 22 in which the structure is formed as a varactor.
24. A method of using a structure according to any one of claims 1-8 as a sensor comprising placing a material in proximity to or in contact with the structure and comparing an electrical characteristic of the structure when the material is in proximity to or in contact with the structure to a reference.

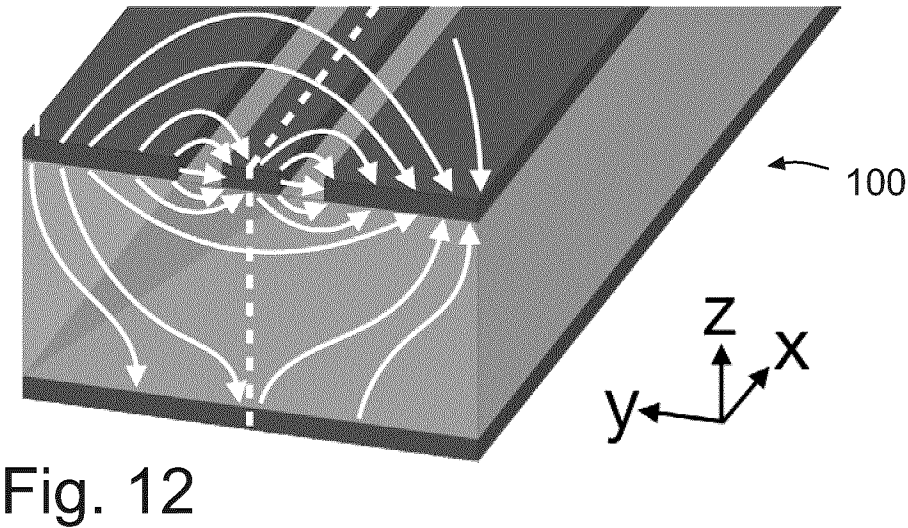
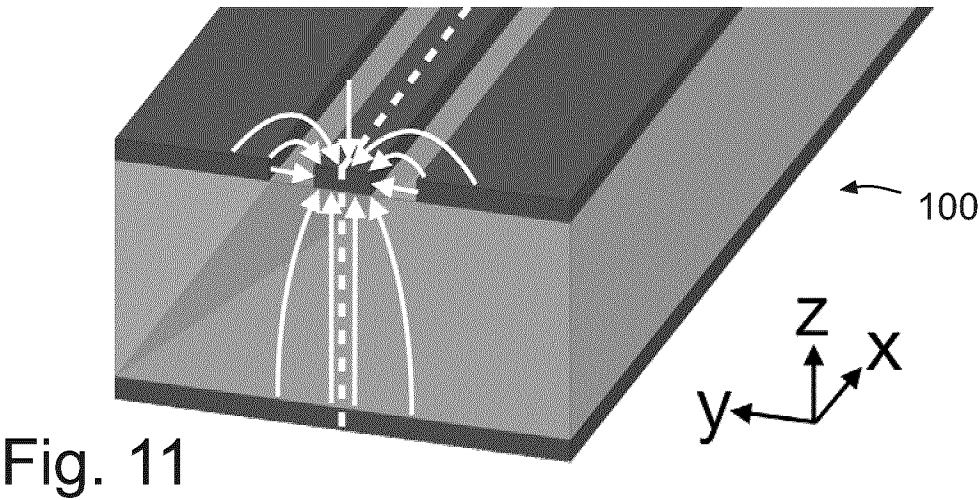
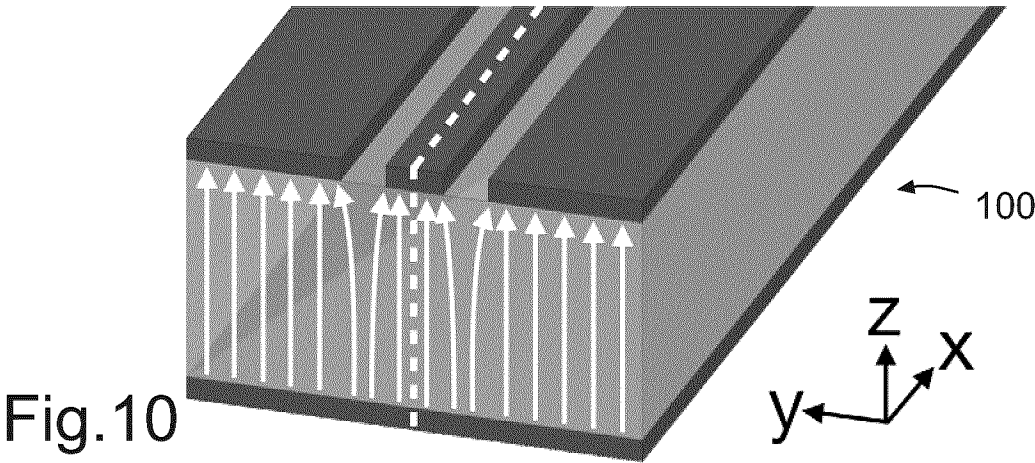
25. The method of claim 24 in which the electrical characteristic is the resonance of the structure.
26. An array formed of structures in accordance with any one of claims 1-8.
27. The array of claim 26 in which the array forms a grid or circular pattern.
28. A beamformer comprising at least a structure according to any one of claims 1-8 formed on a first side of a dielectric slab.
29. The beamformer of claim 28 further comprising at least an additional structure according to any one of claims 1-8 formed on the second side of the dielectric slab.
30. An absorber comprising at least a structure according to any one of claims 1-8 formed on a first side of a lossy dielectric slab.



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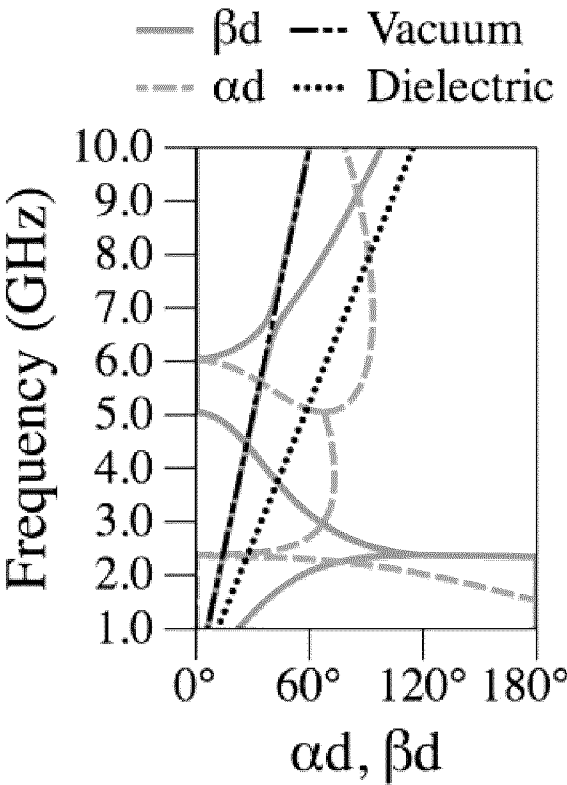


Fig. 13A

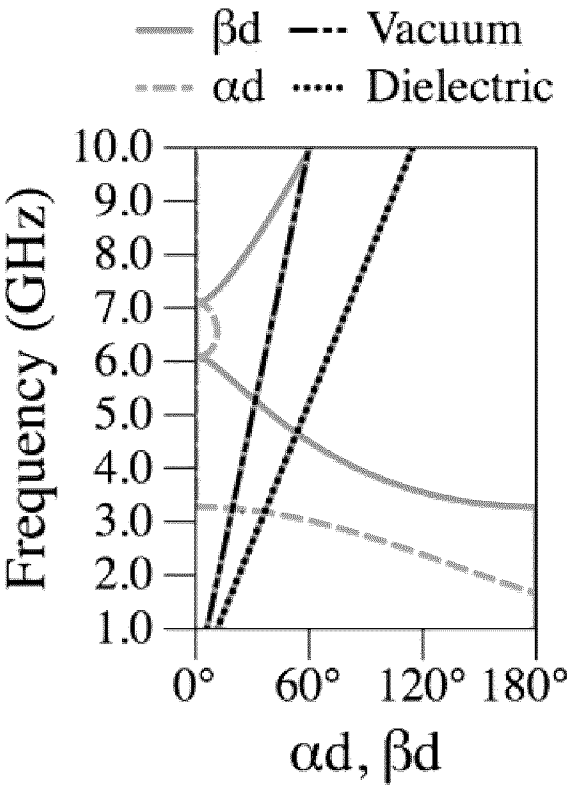


Fig. 13B

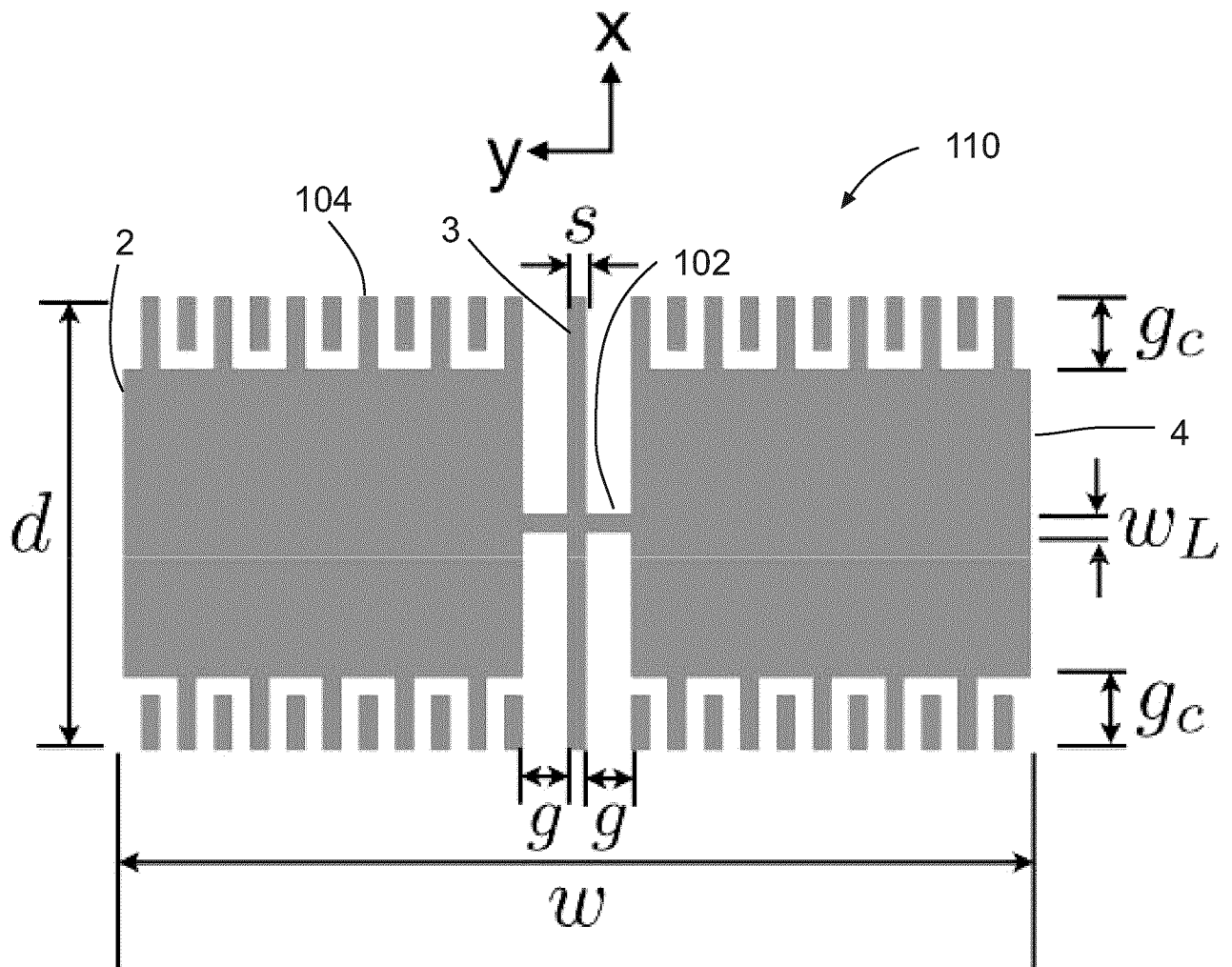


Fig. 14

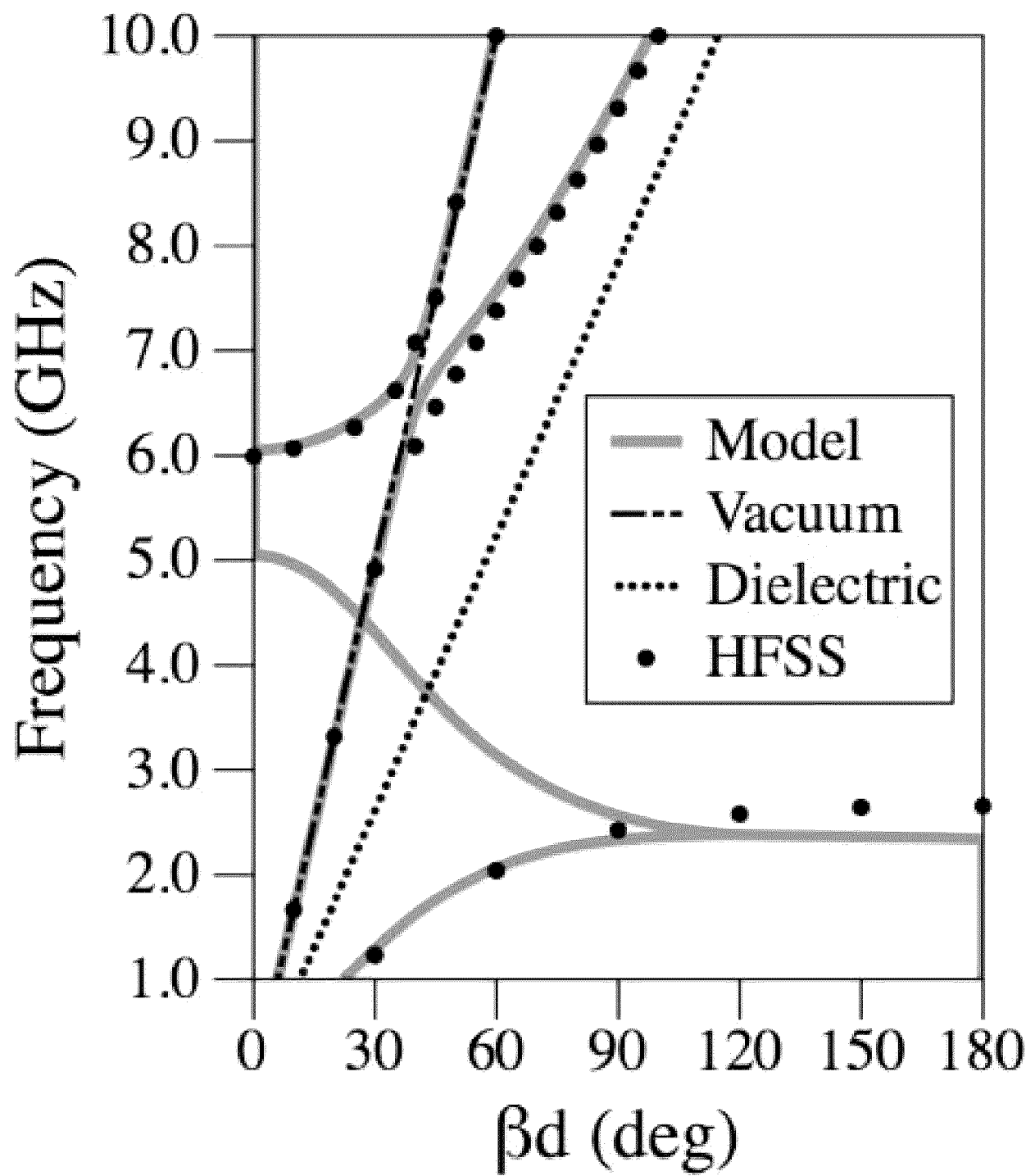


Fig. 15

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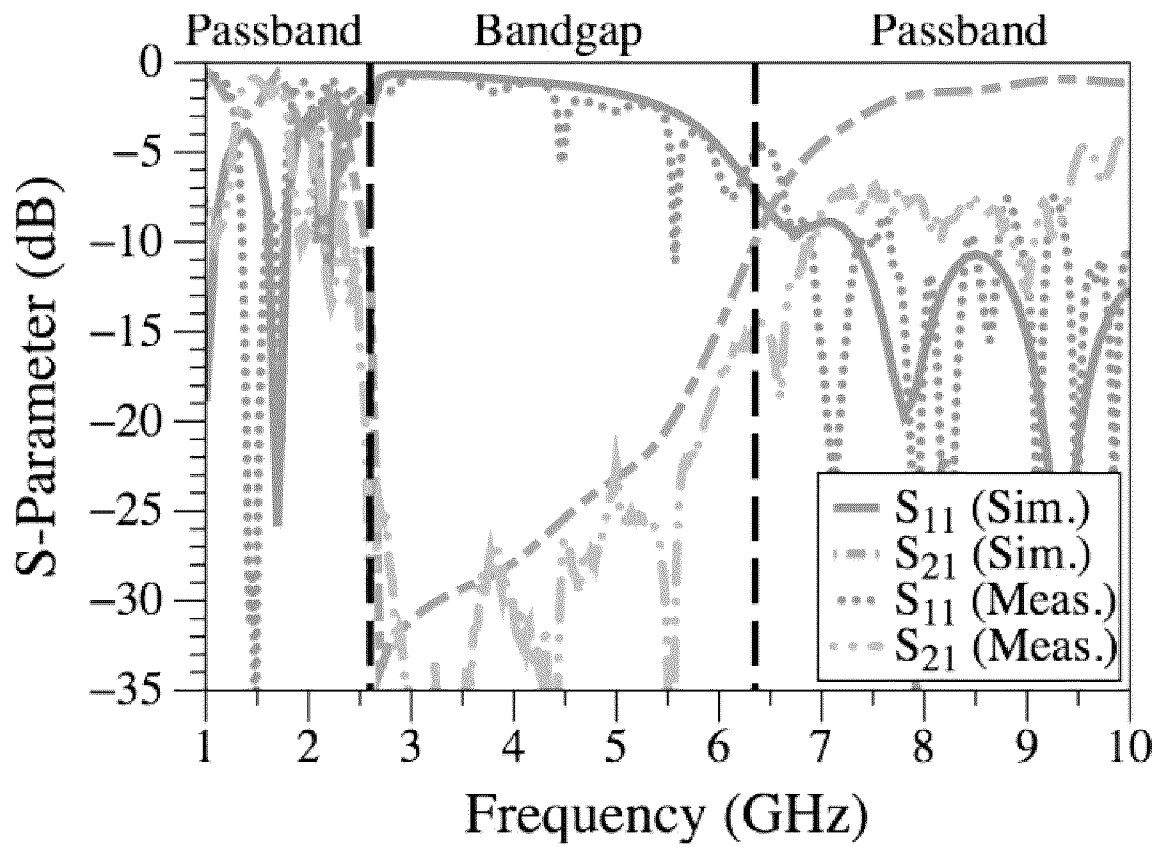


Fig. 16

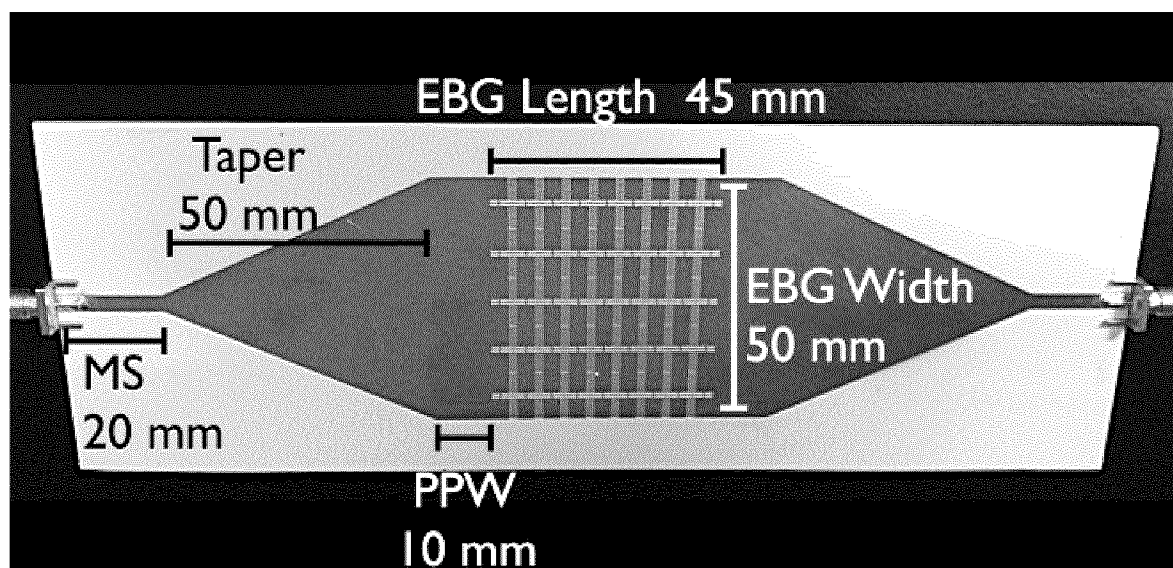


Fig. 17

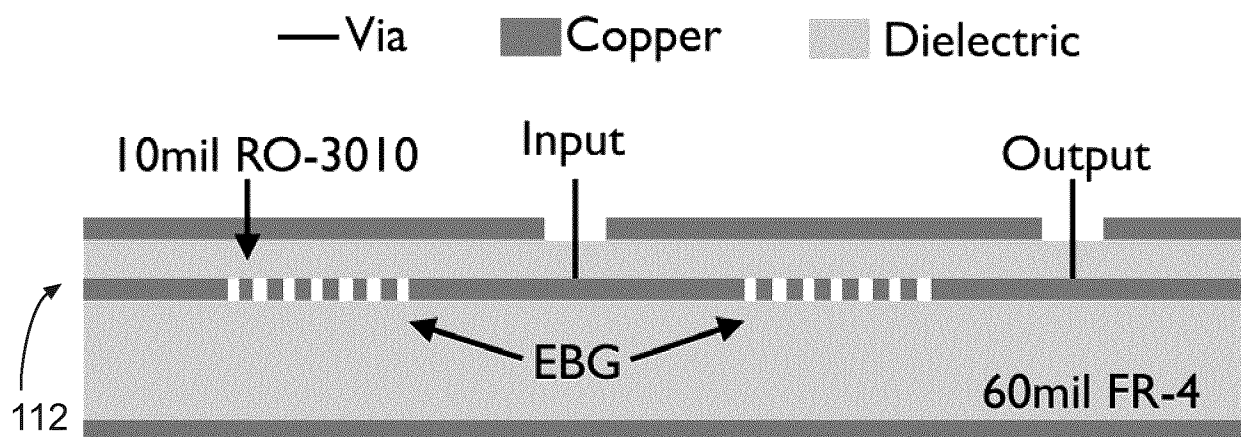
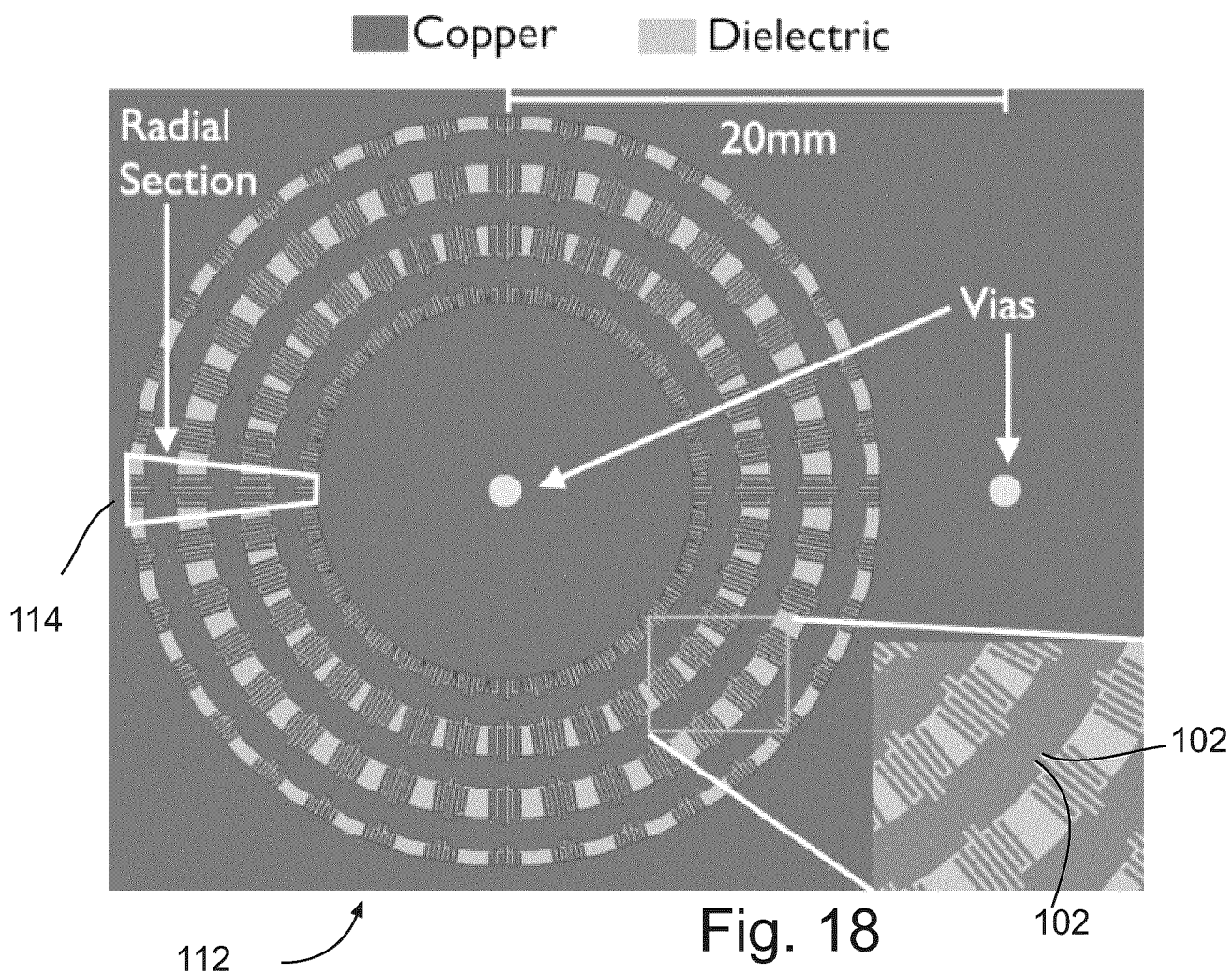
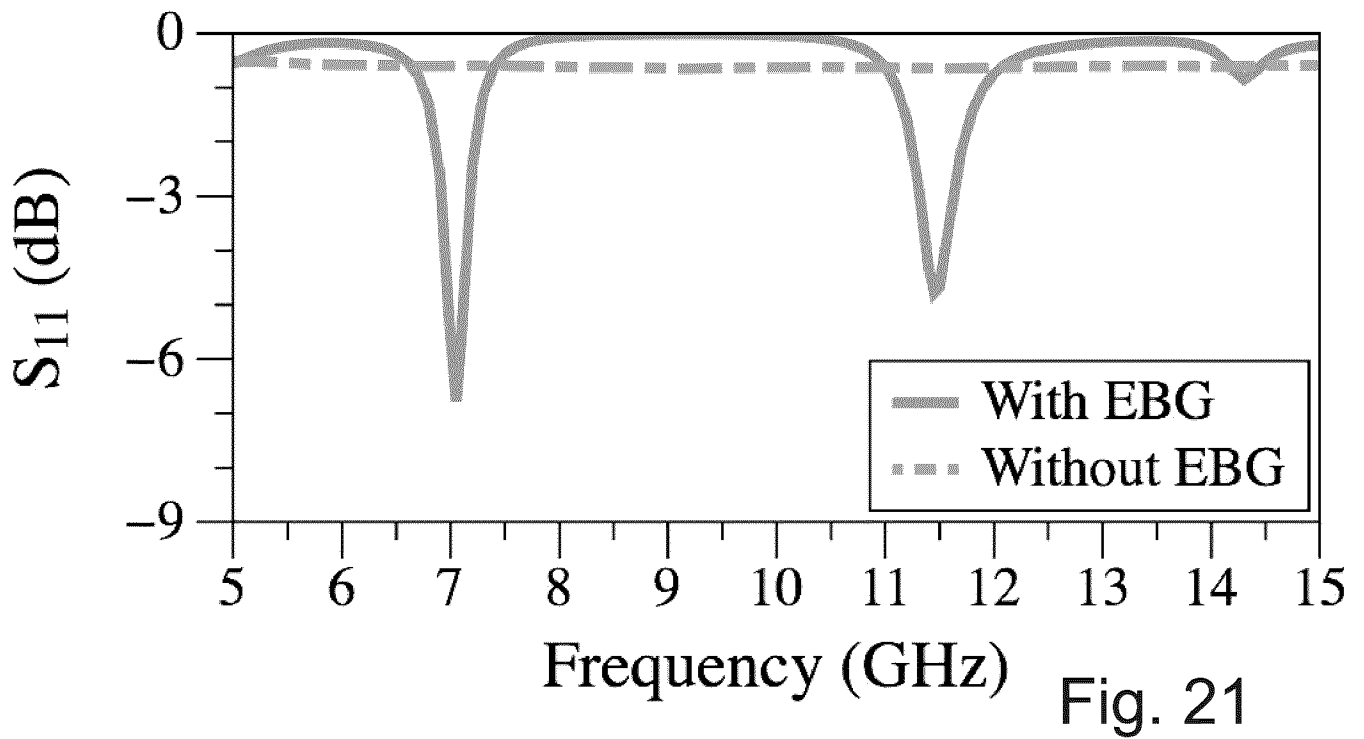
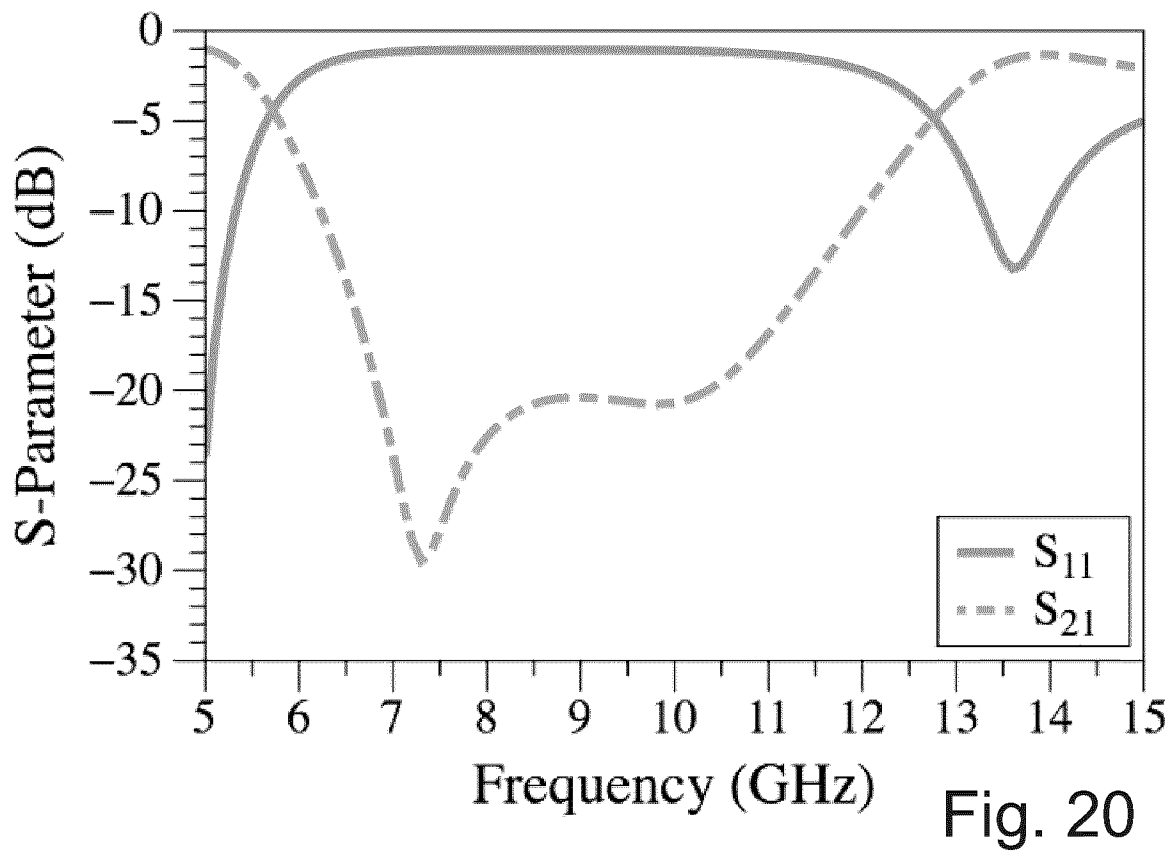


Fig. 19

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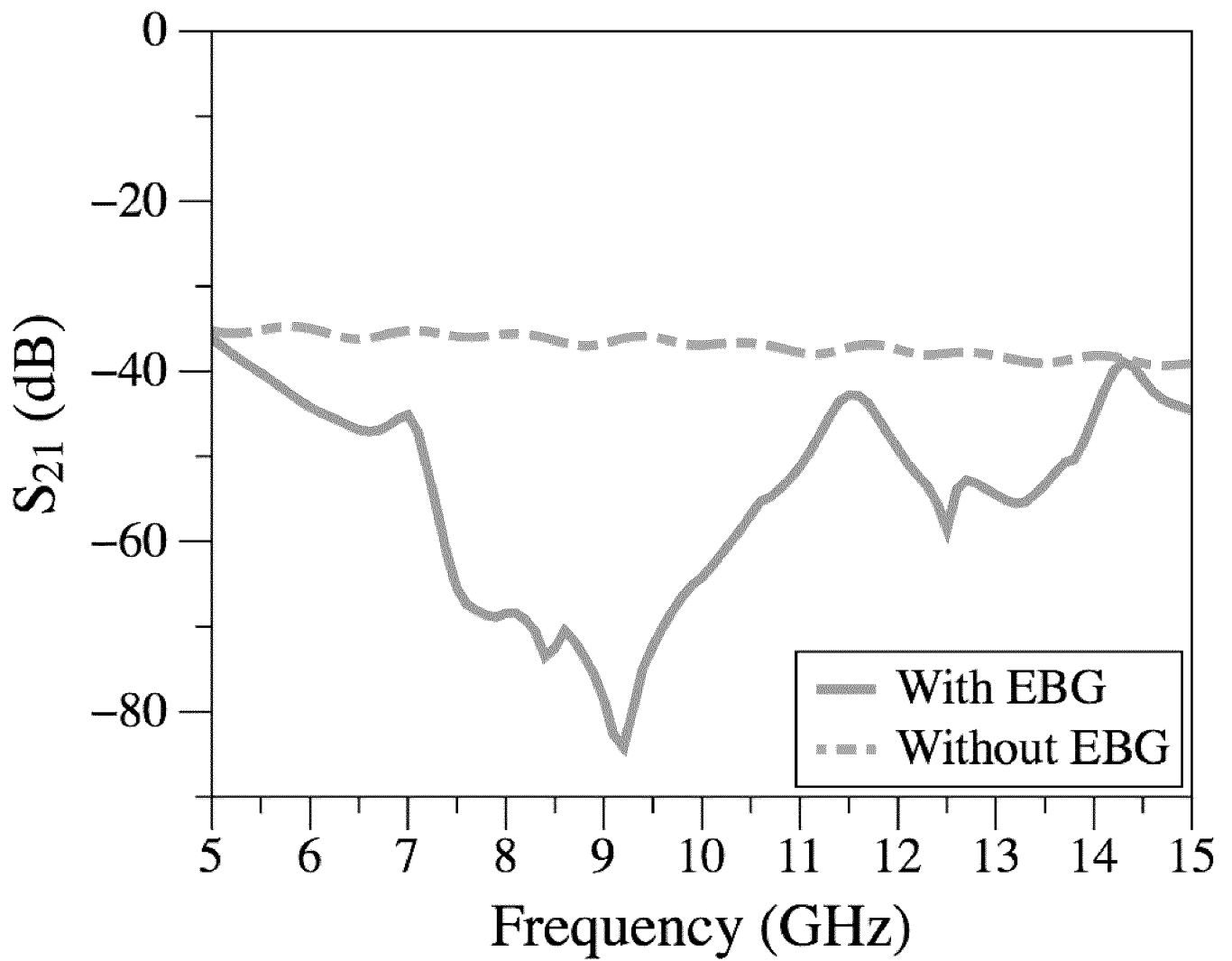


Fig. 22

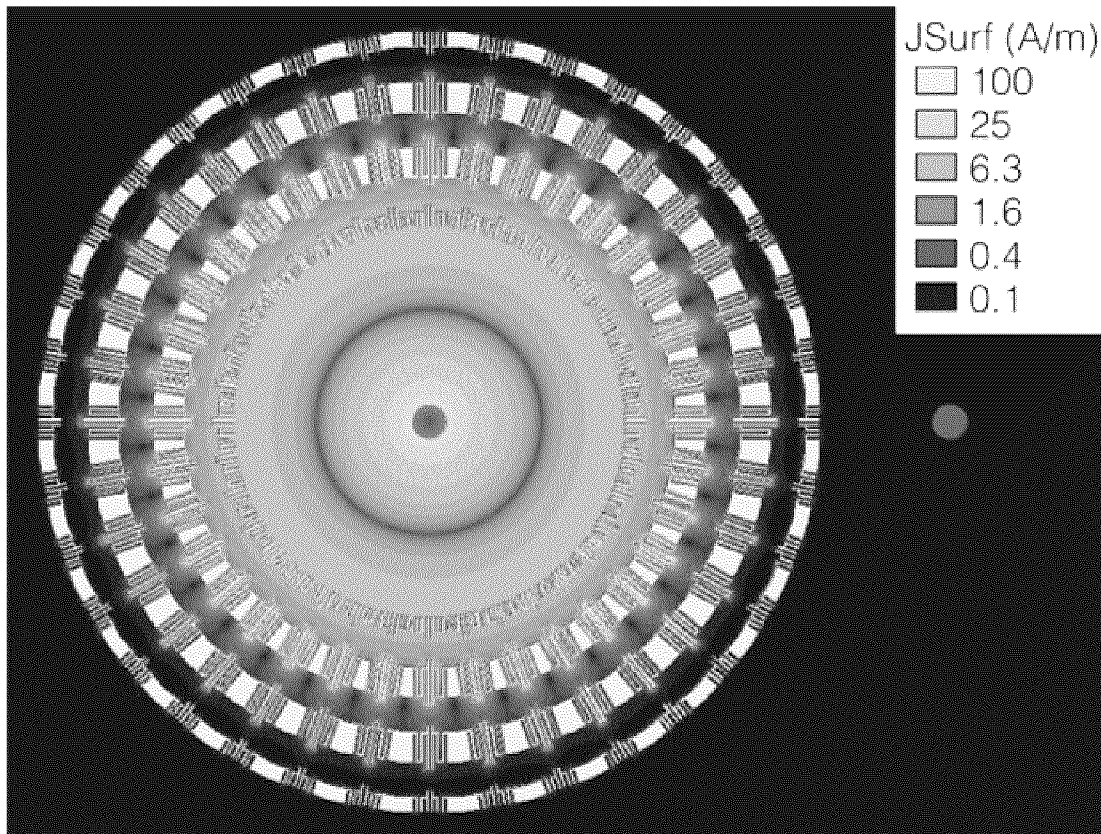


Fig. 23

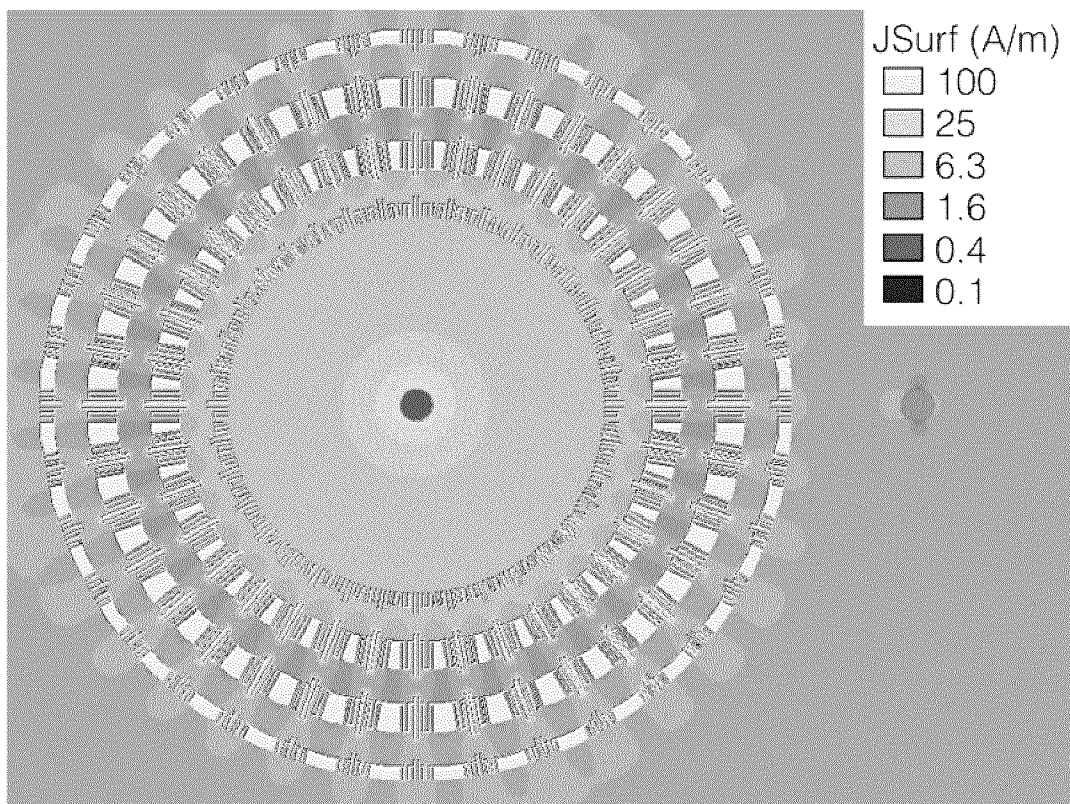
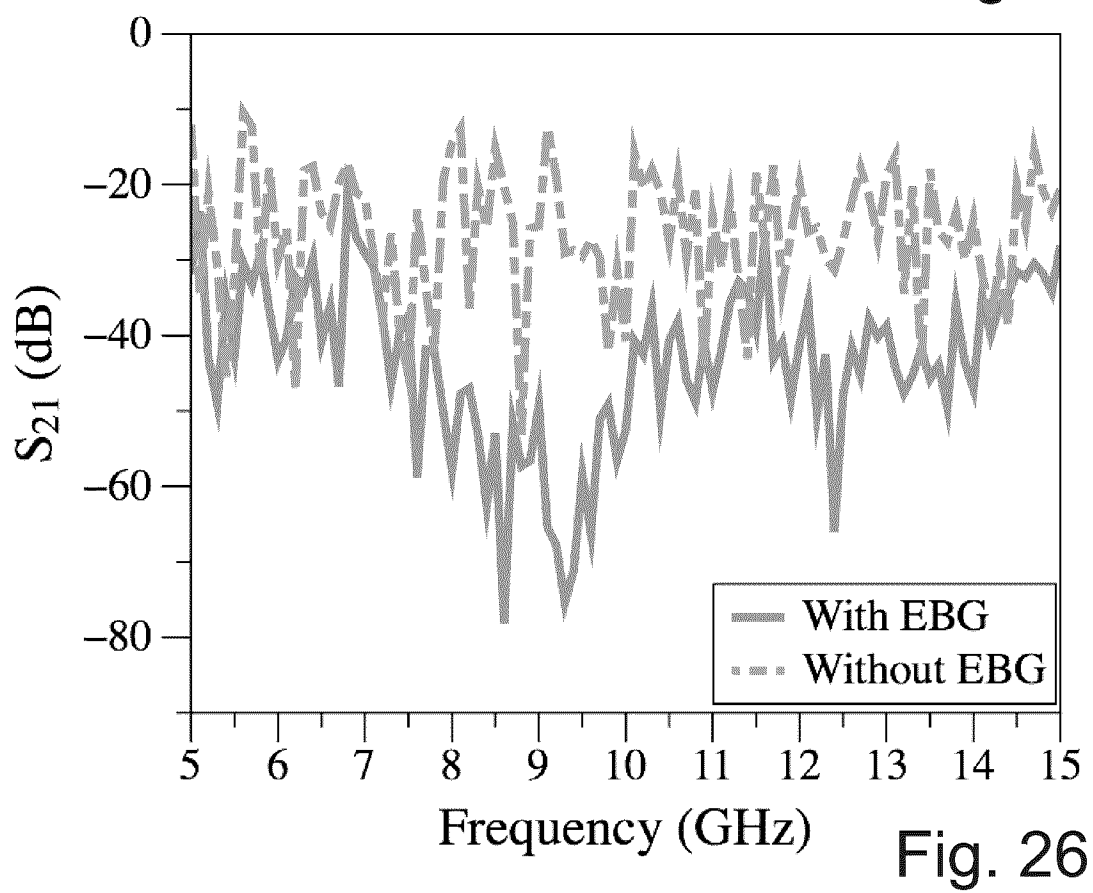
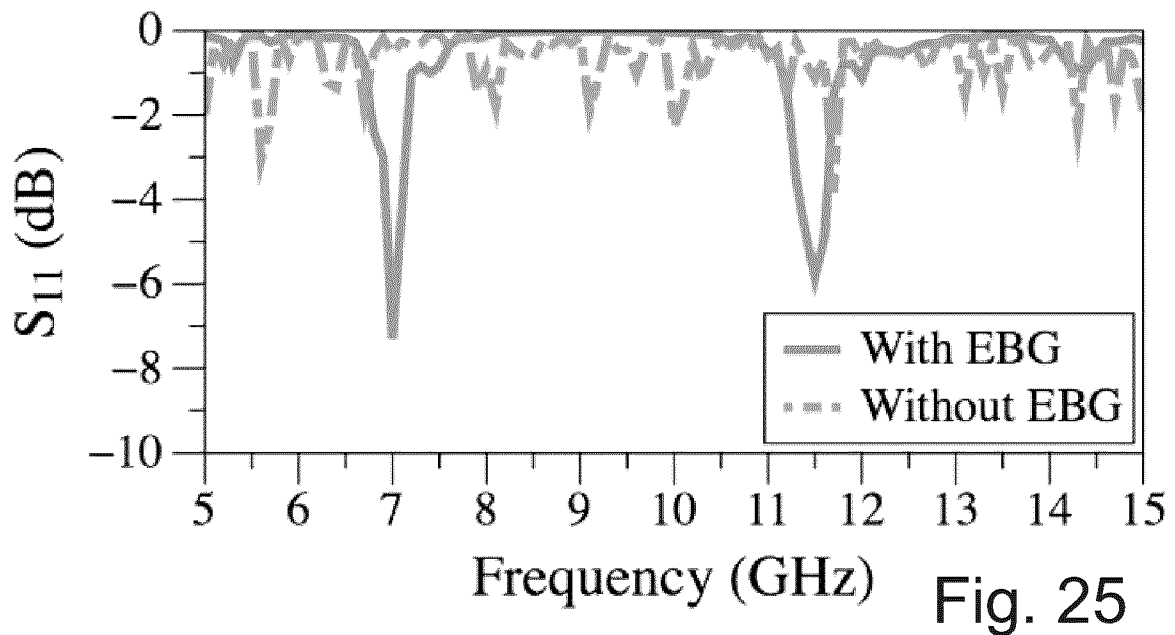


Fig. 24



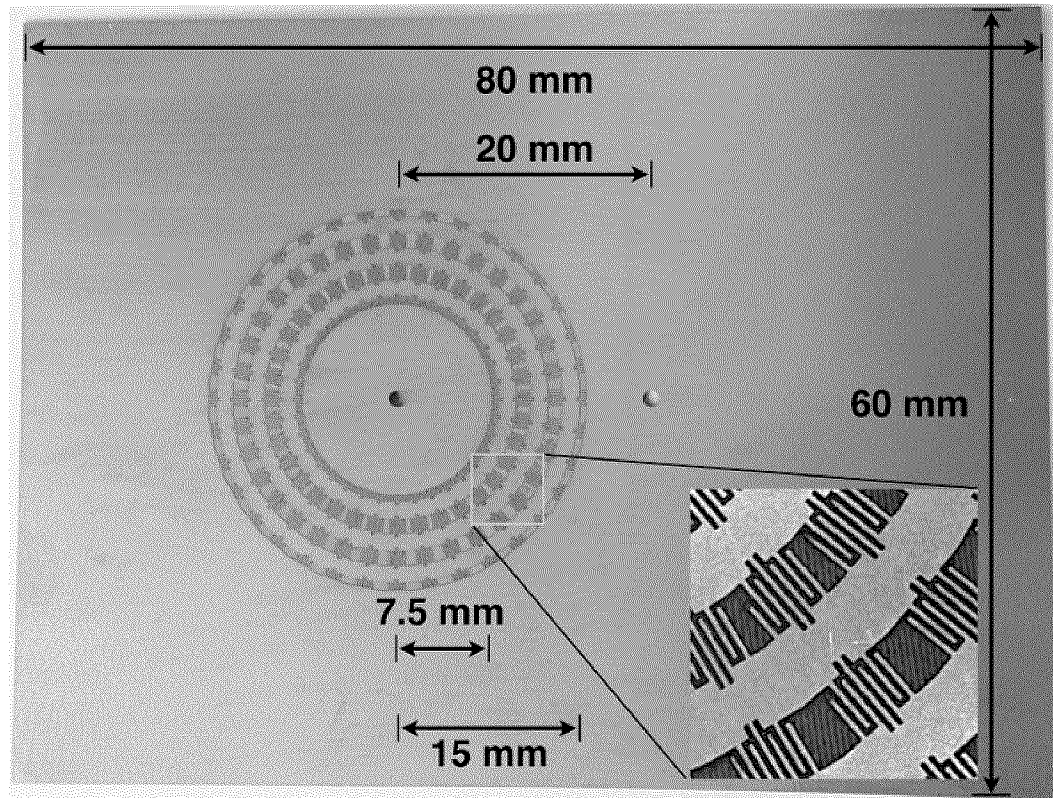


Fig. 27

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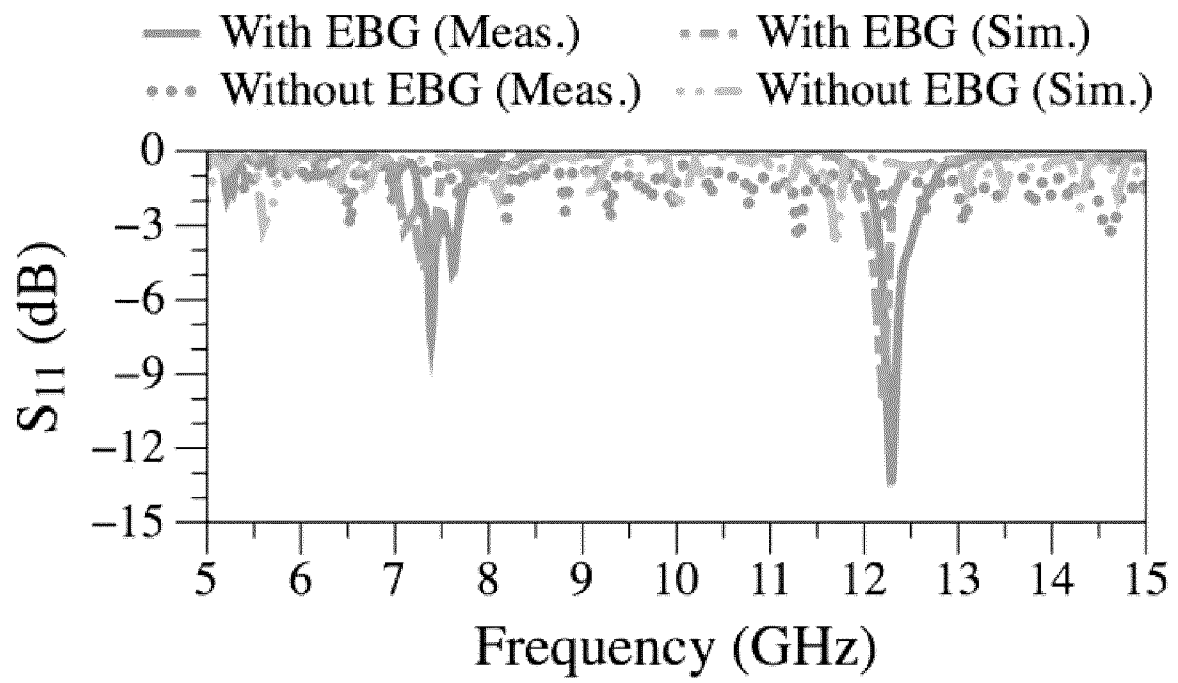


Fig. 28

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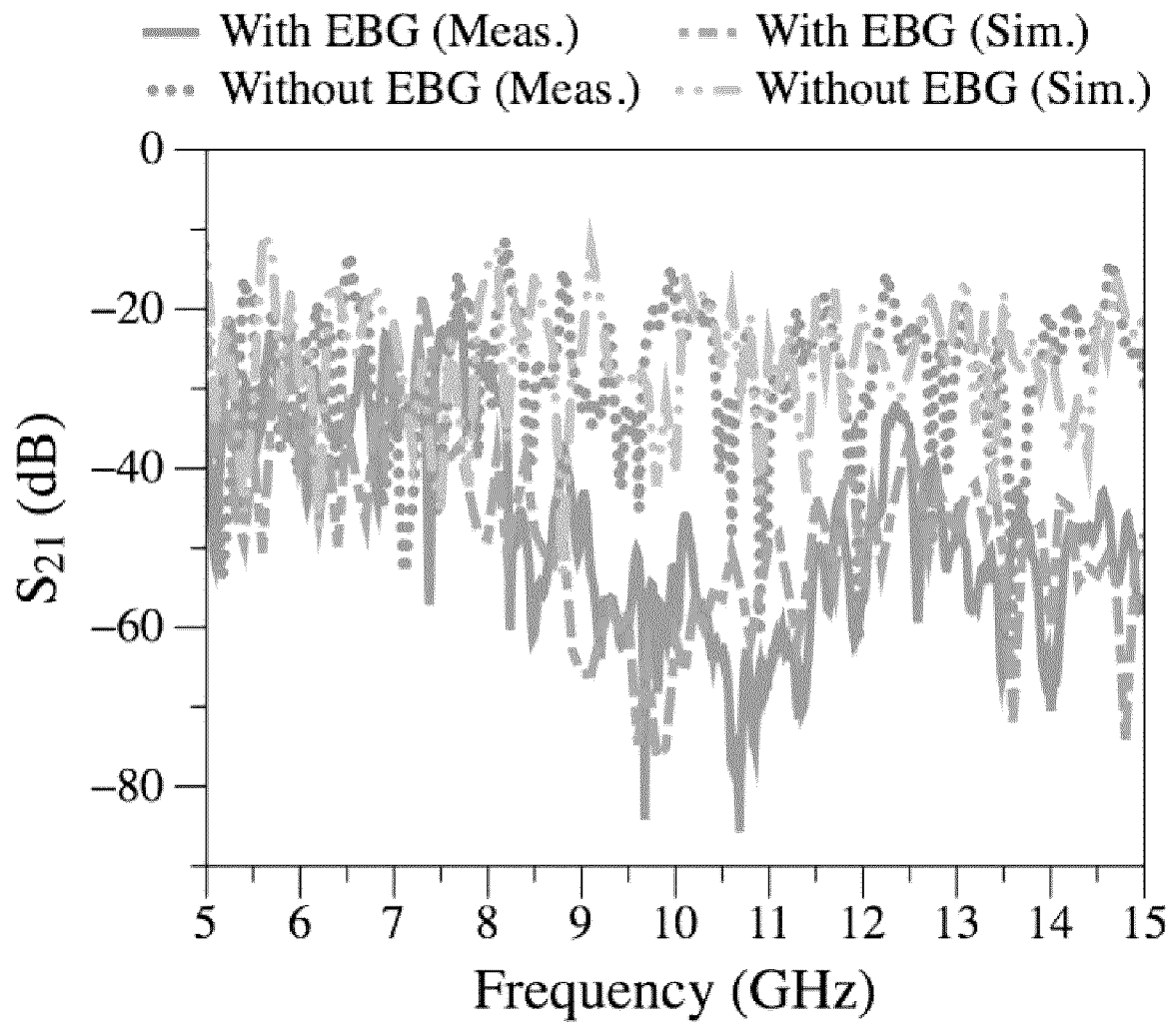


Fig. 29

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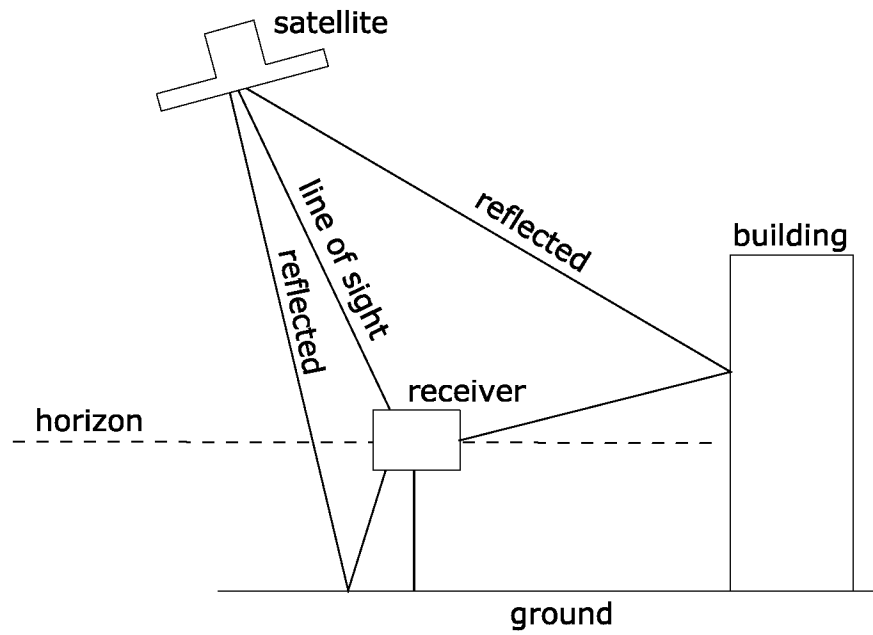


Fig. 30

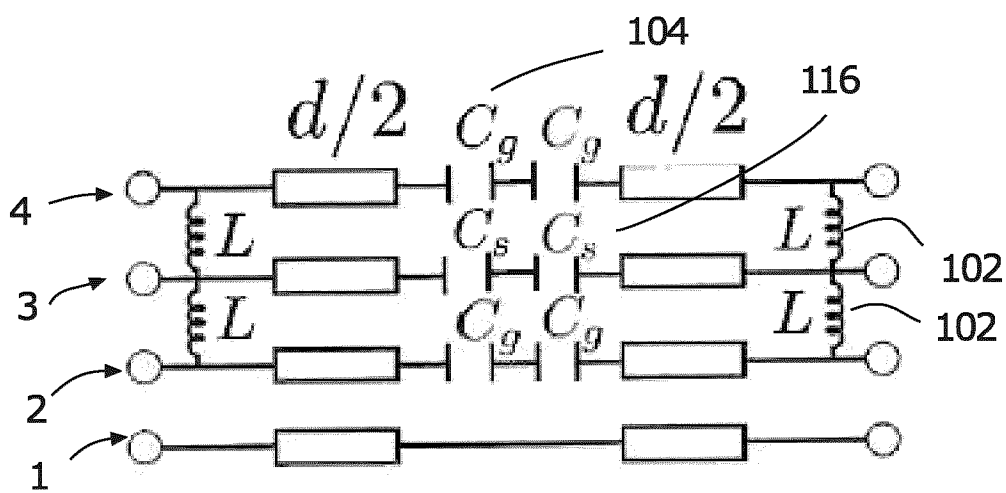


Fig. 31

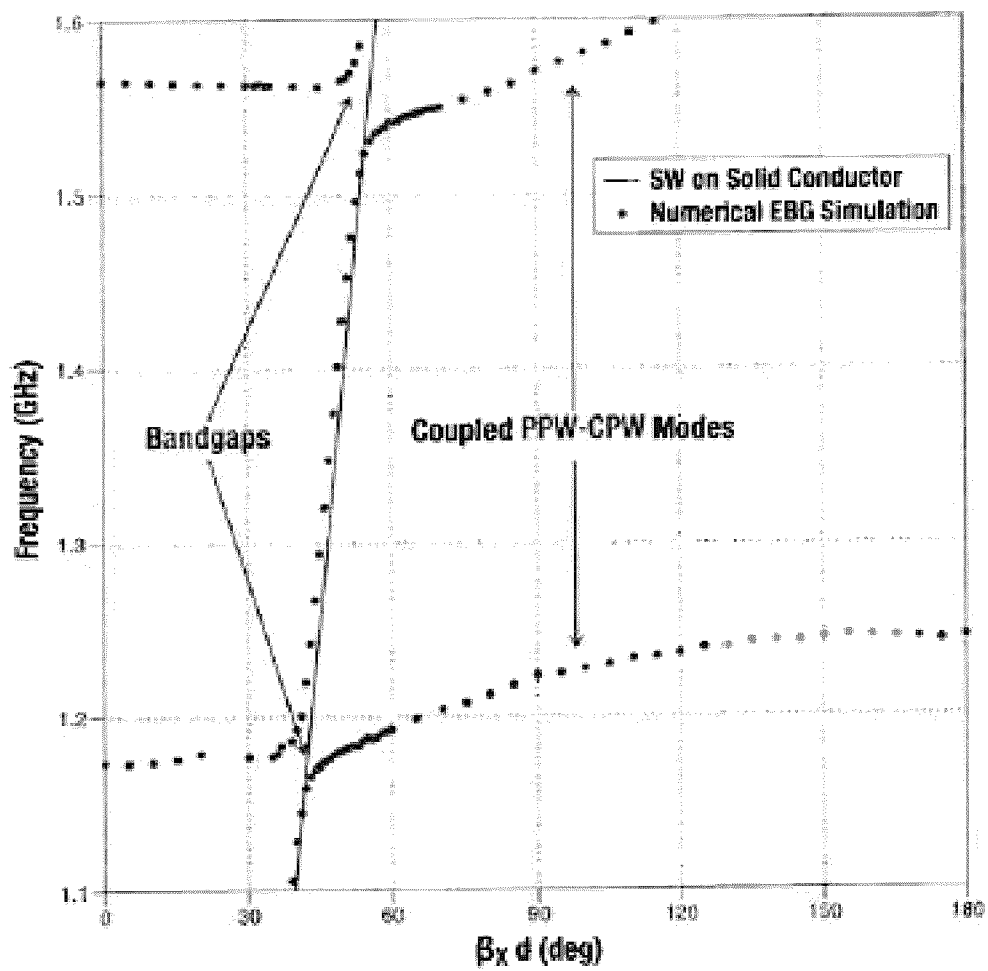
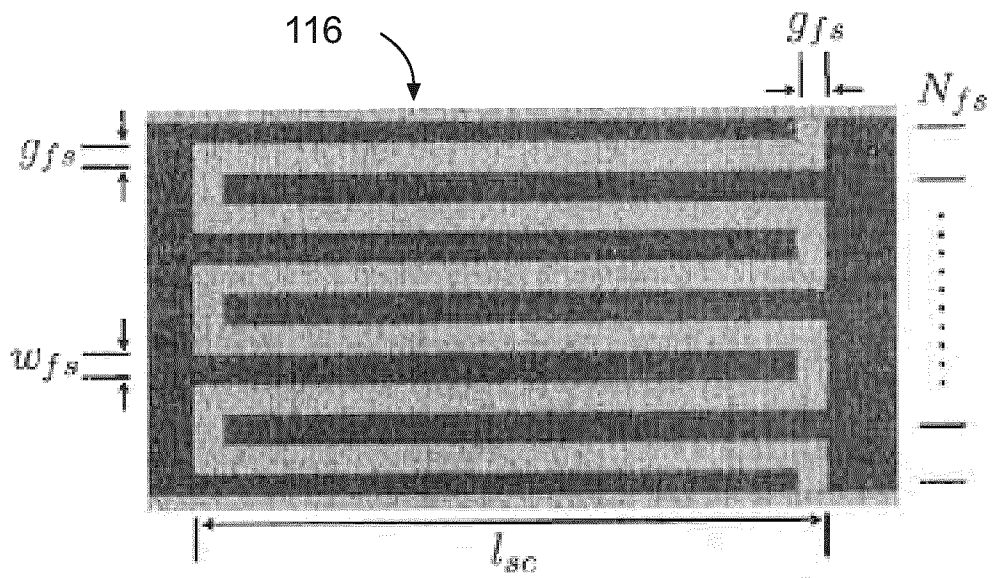
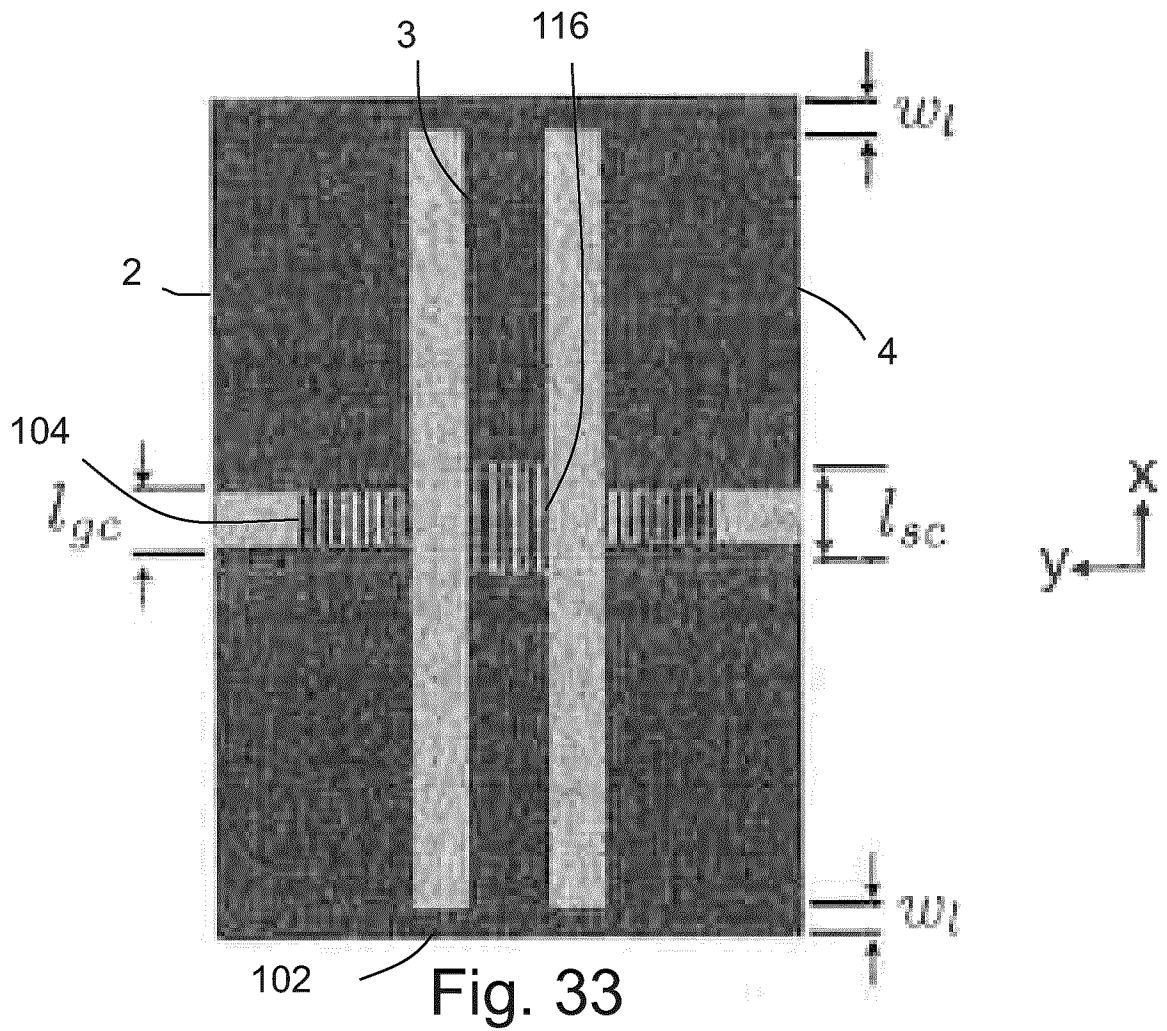


Fig. 32

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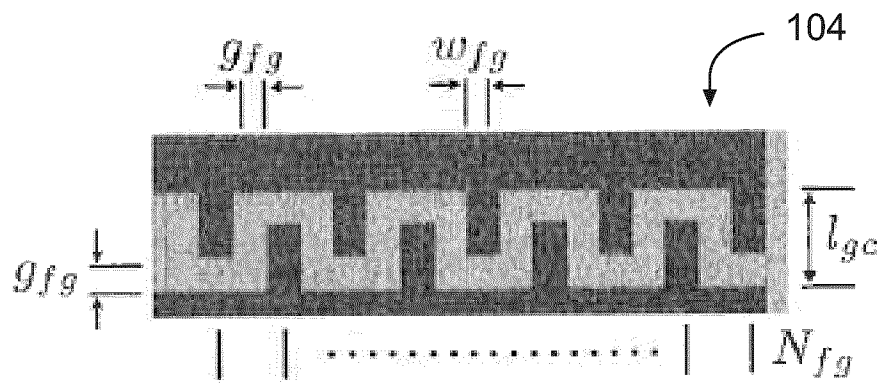


Fig. 35

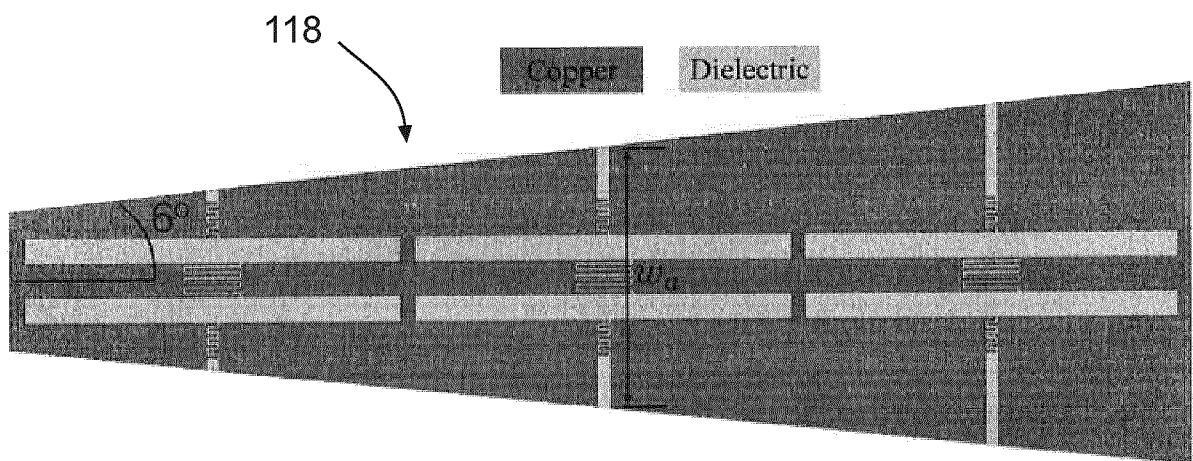


Fig. 36

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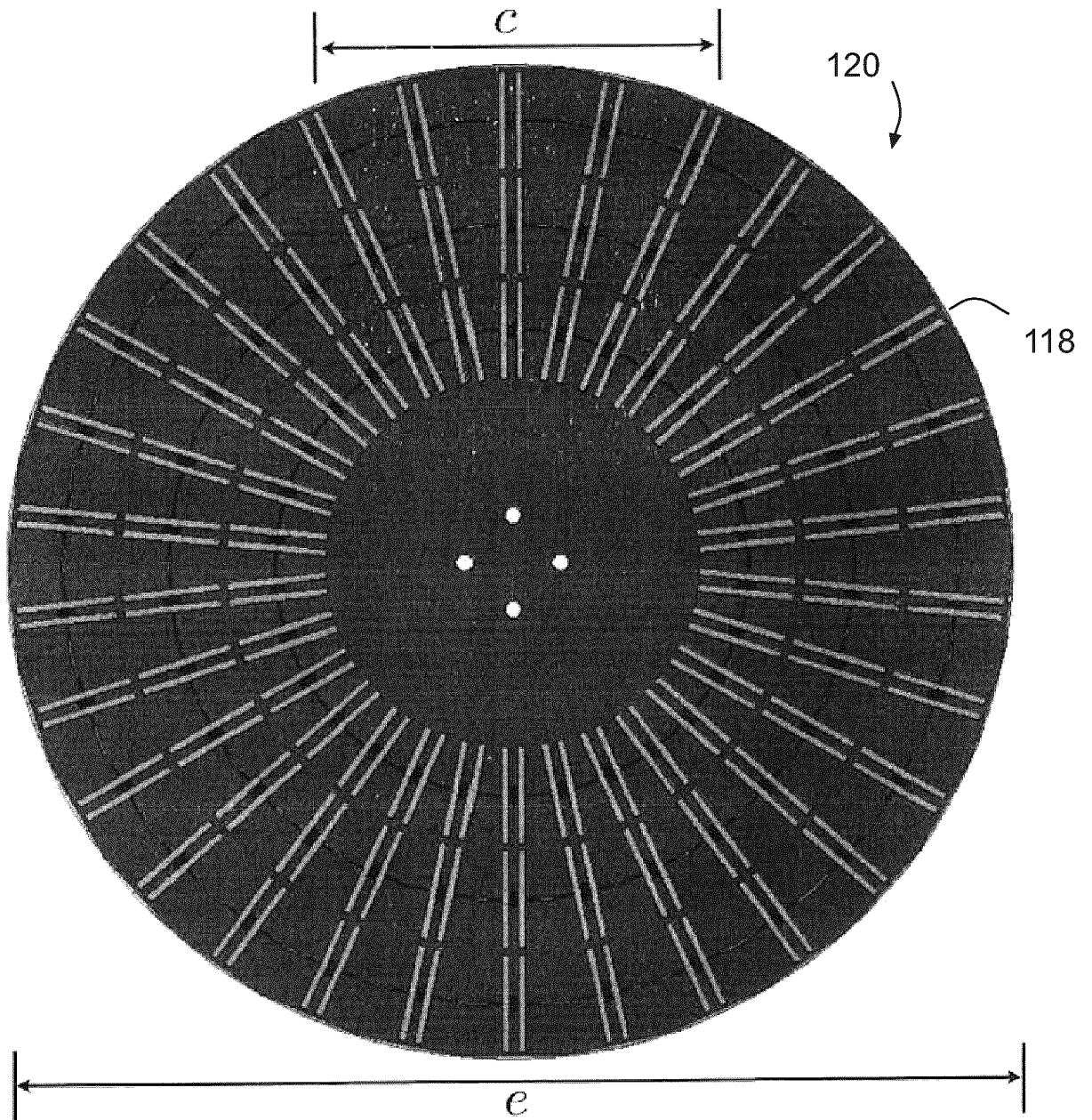


Fig. 37

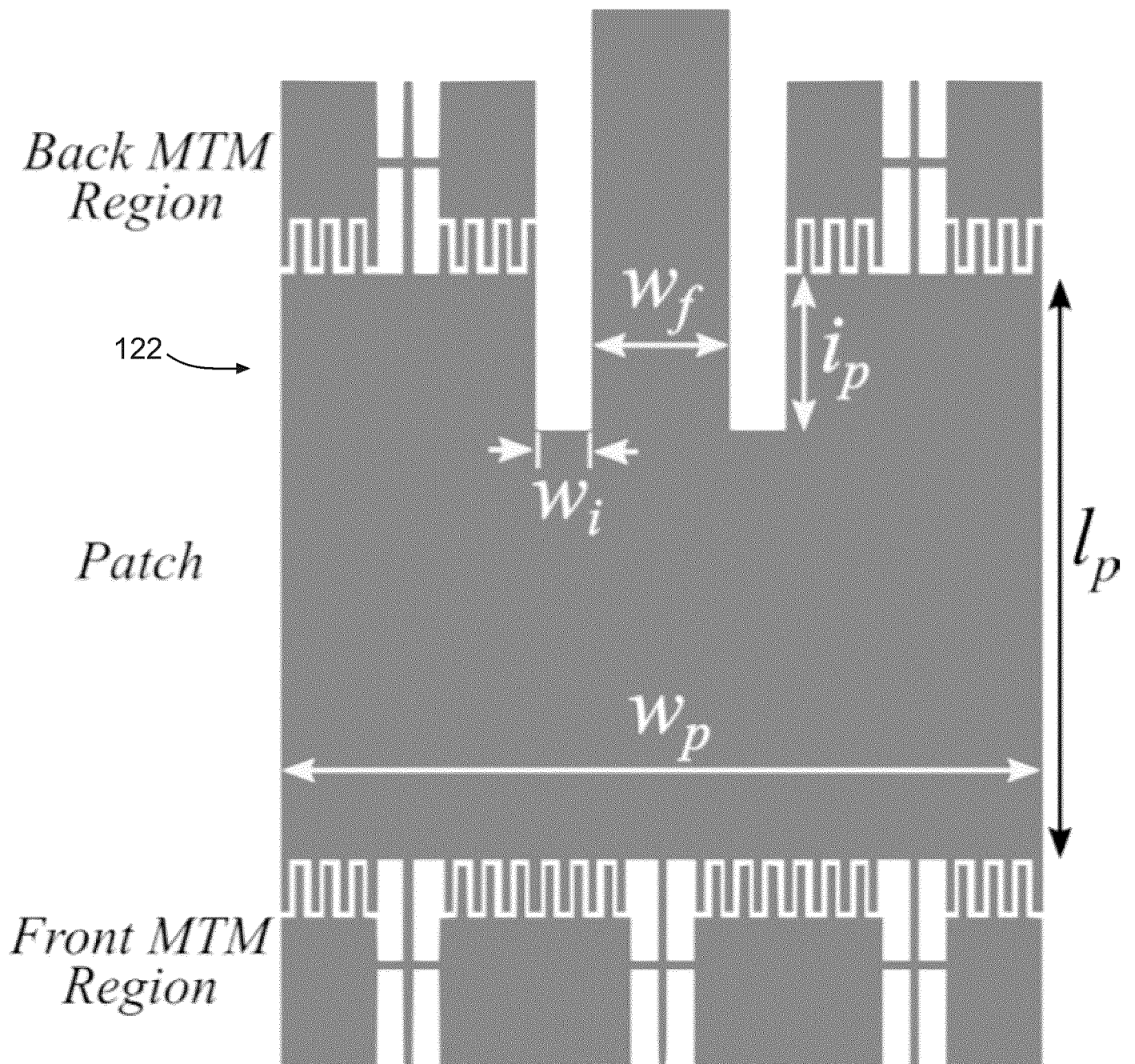
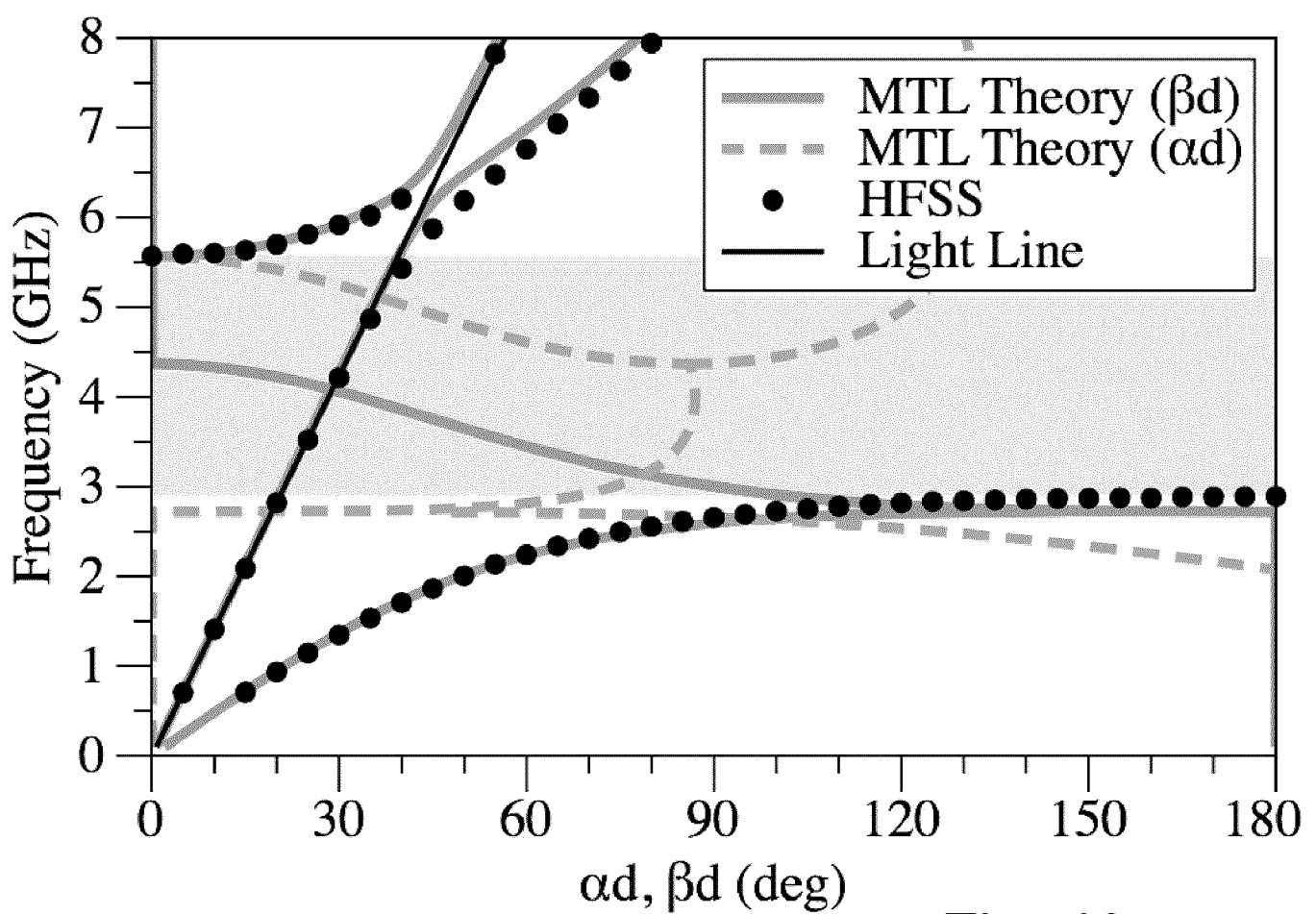
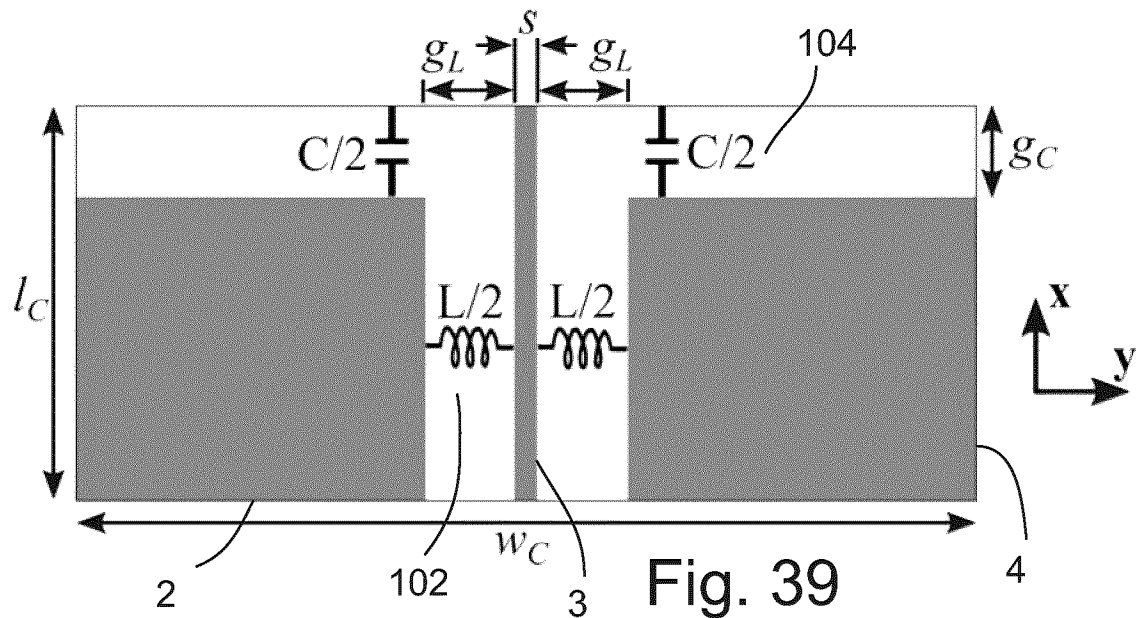


Fig. 38

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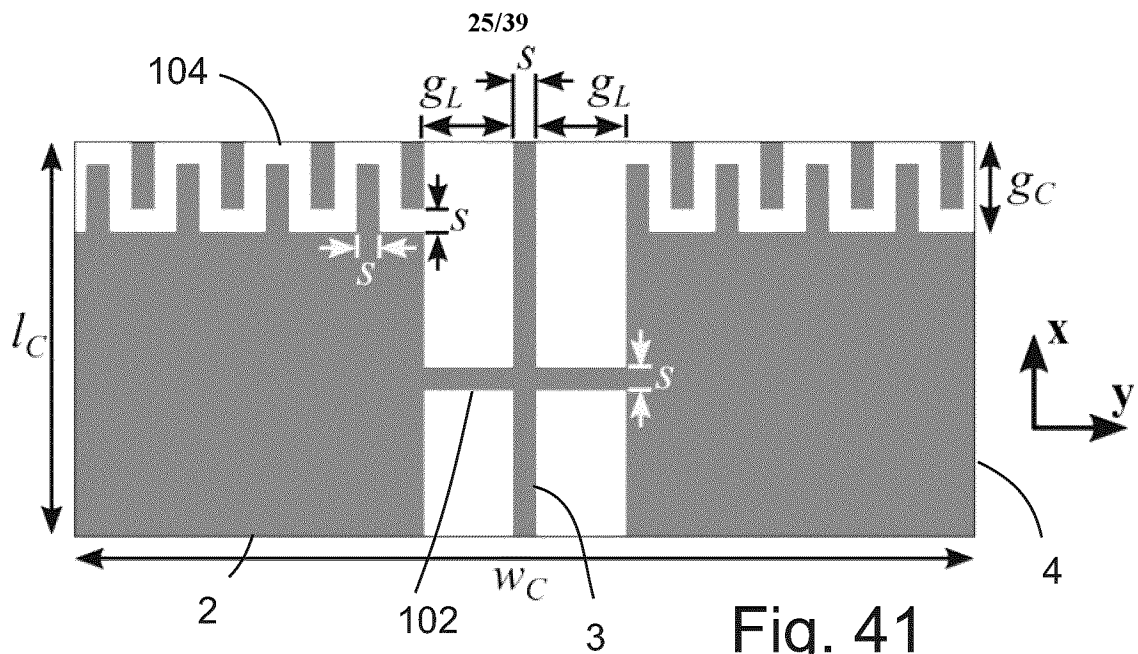


Fig. 41

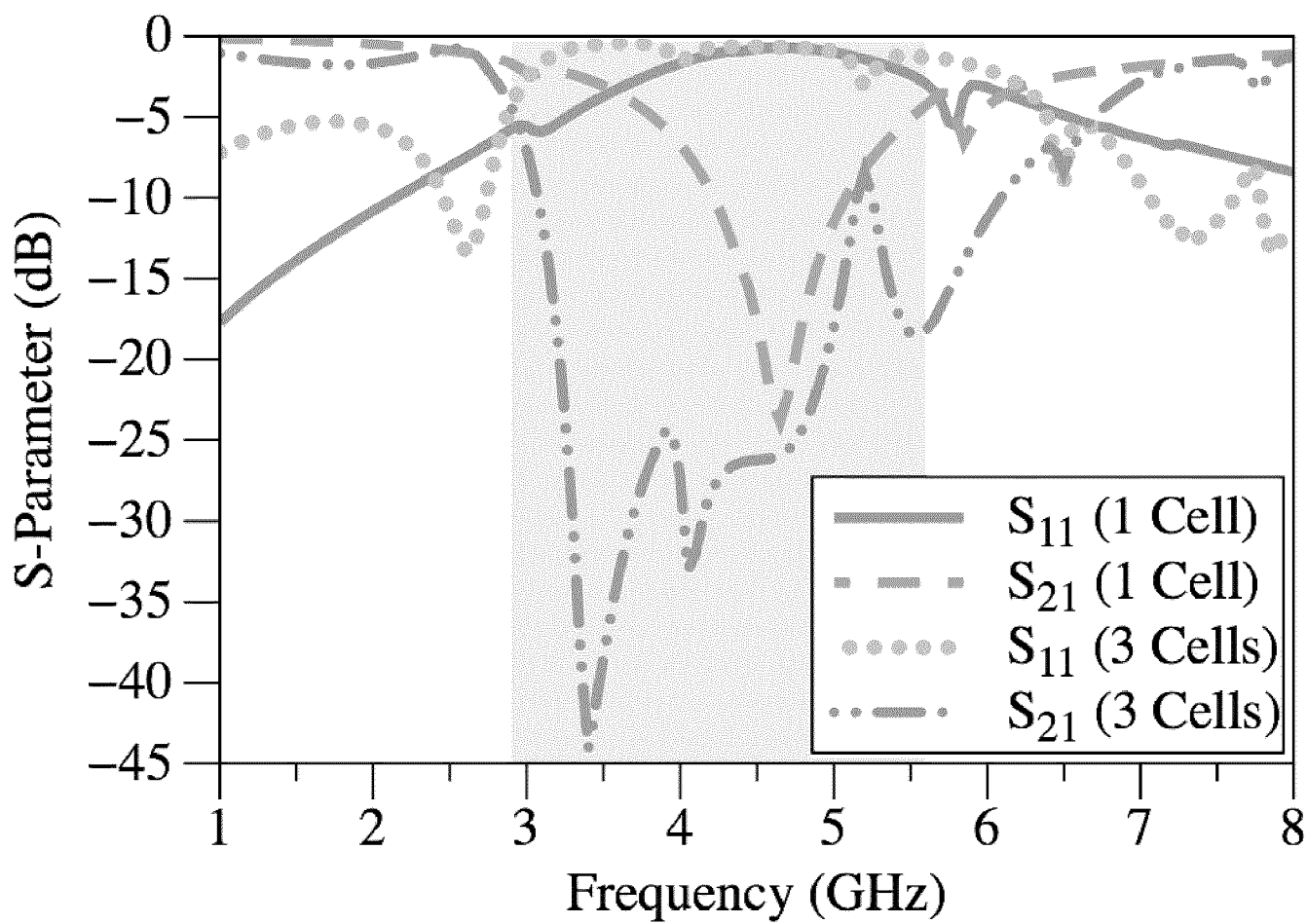


Fig. 42

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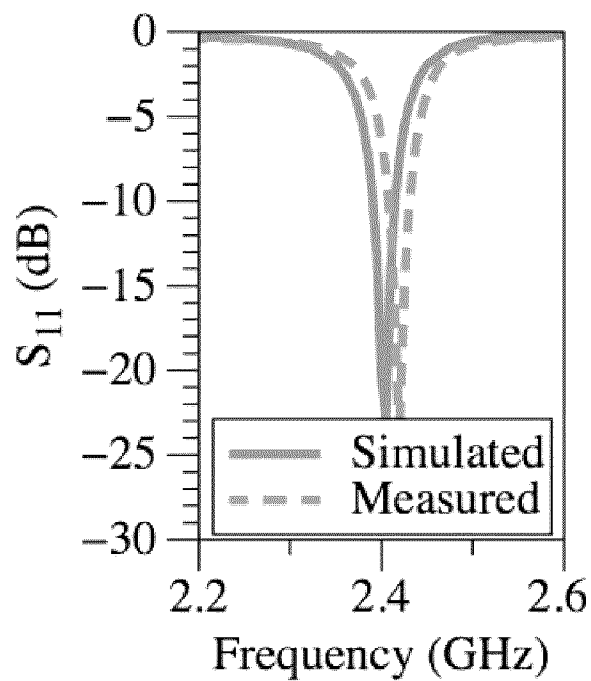


Fig. 43A

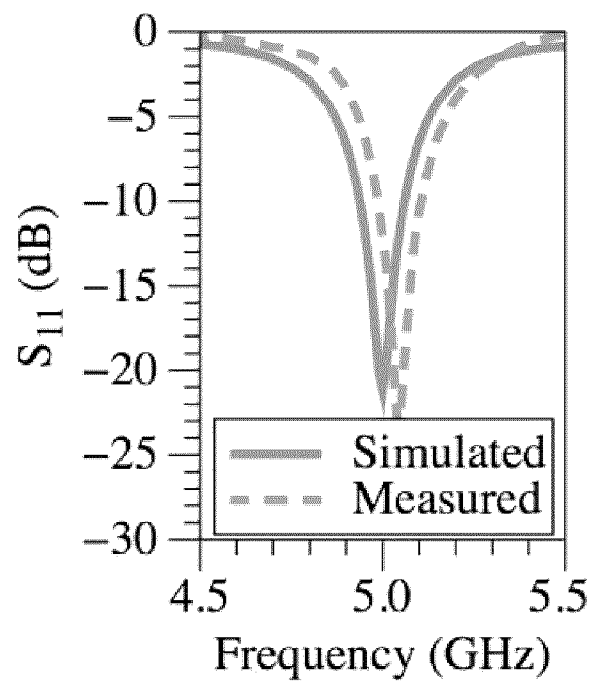


Fig. 43B

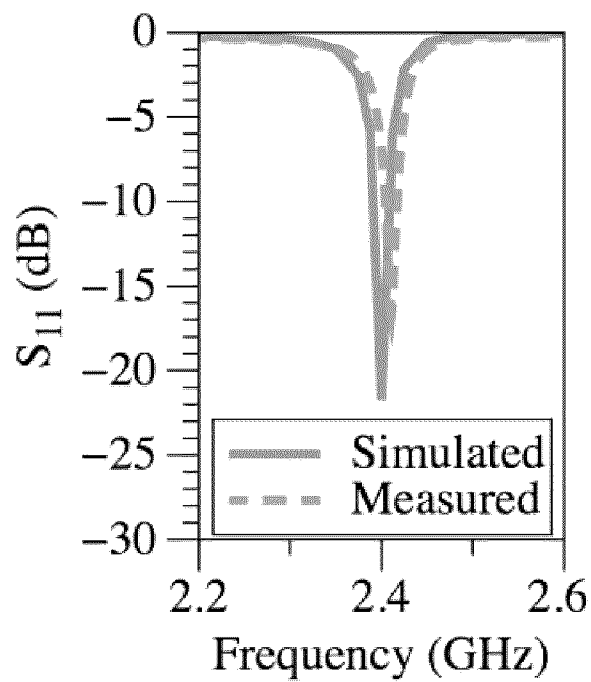


Fig. 43C

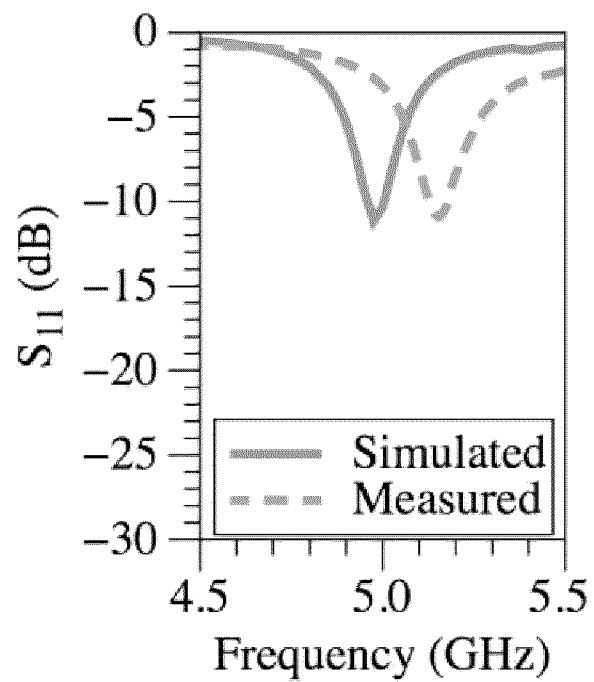


Fig. 43D

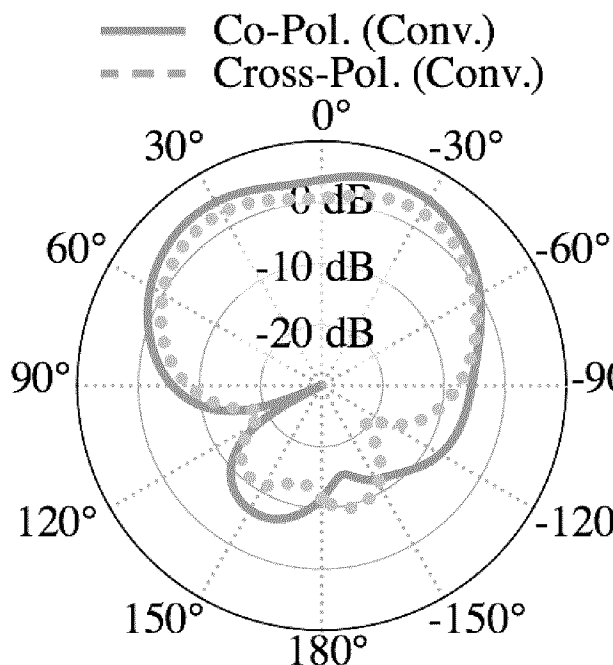


Fig. 44A

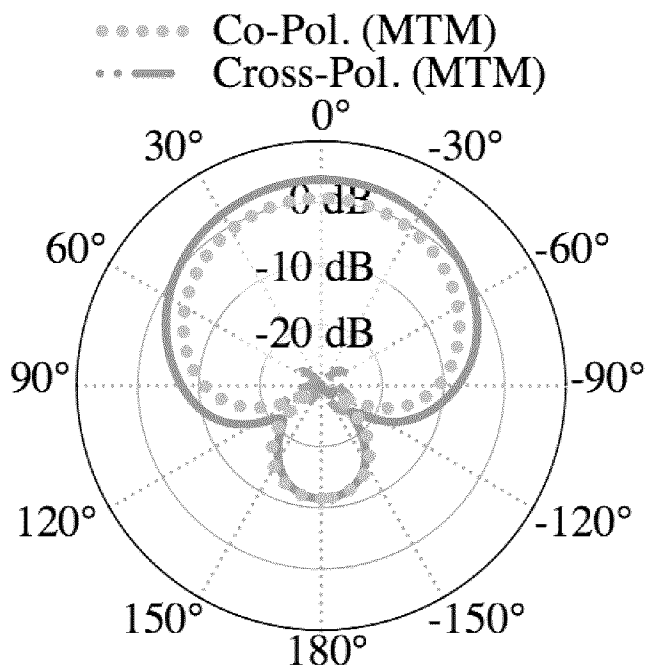


Fig. 44B

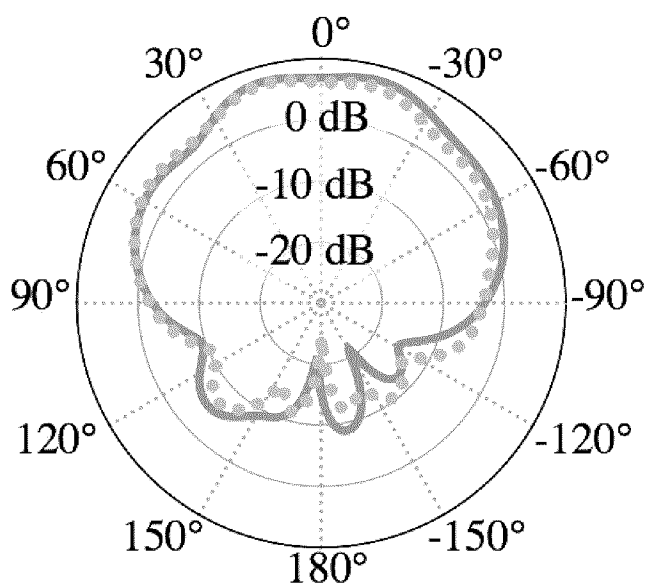


Fig. 44C

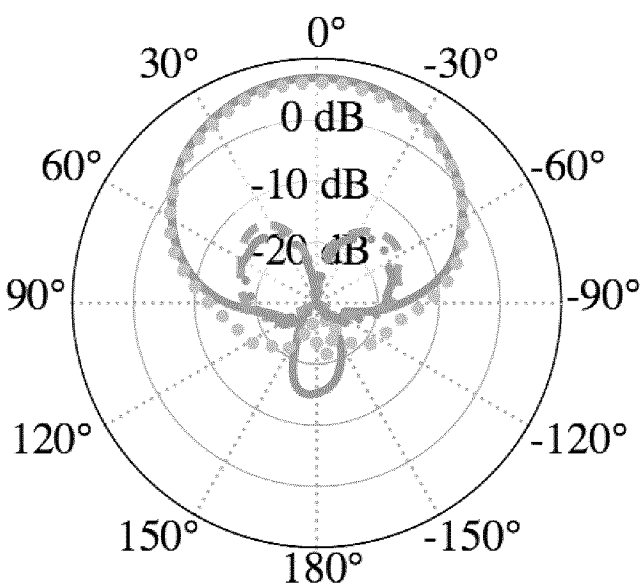


Fig. 44D

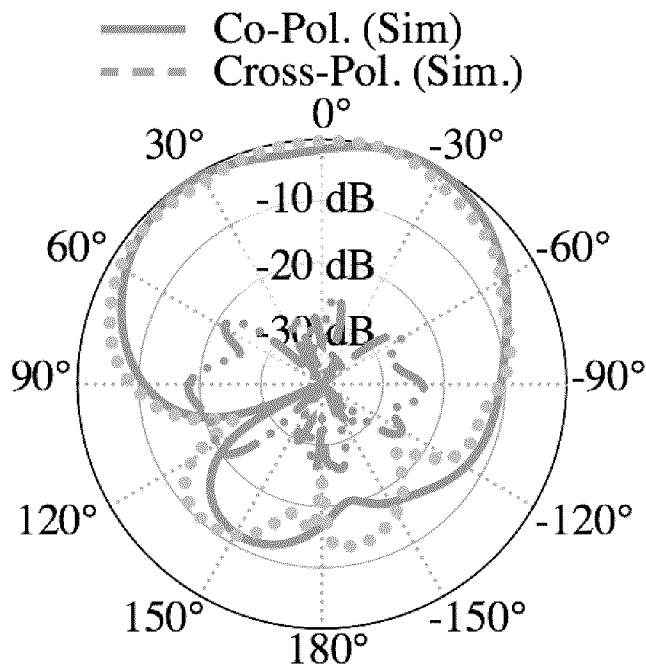


Fig. 45A

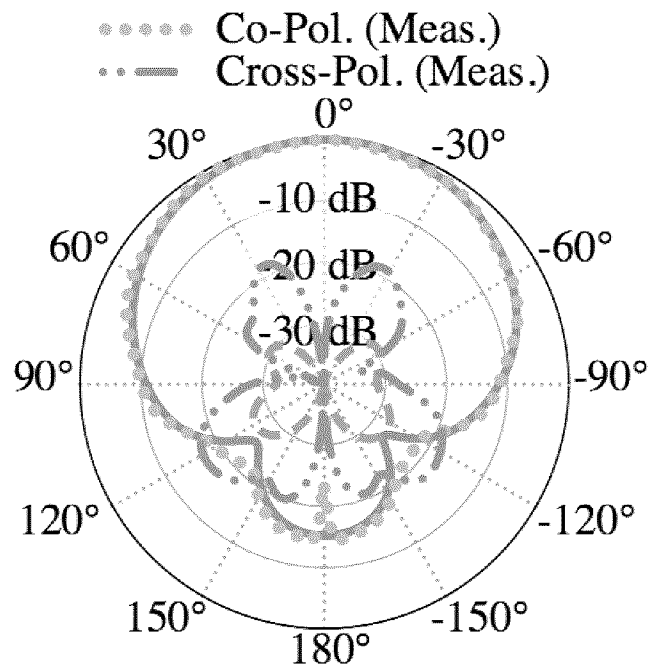


Fig. 45B

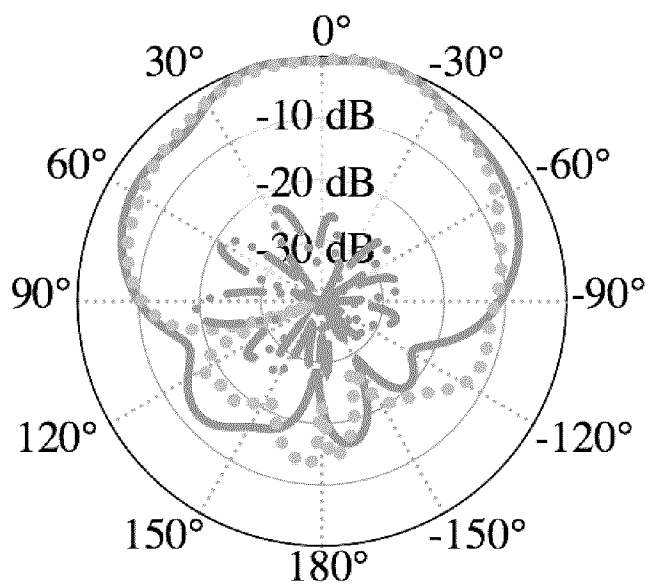


Fig. 45C

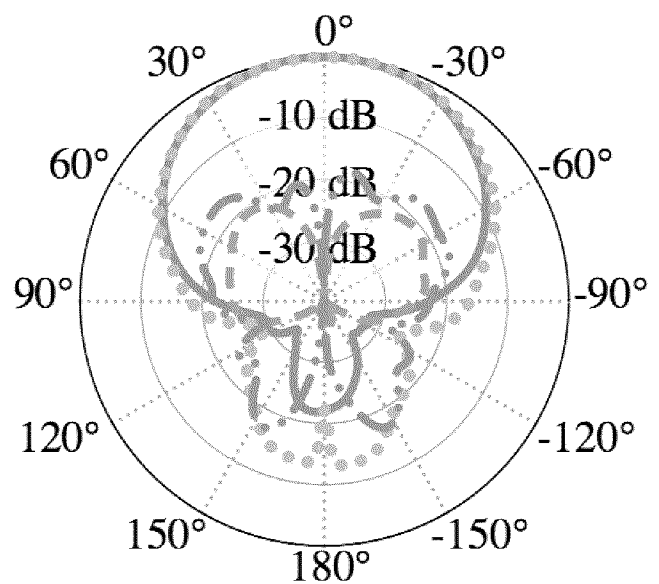


Fig. 45D

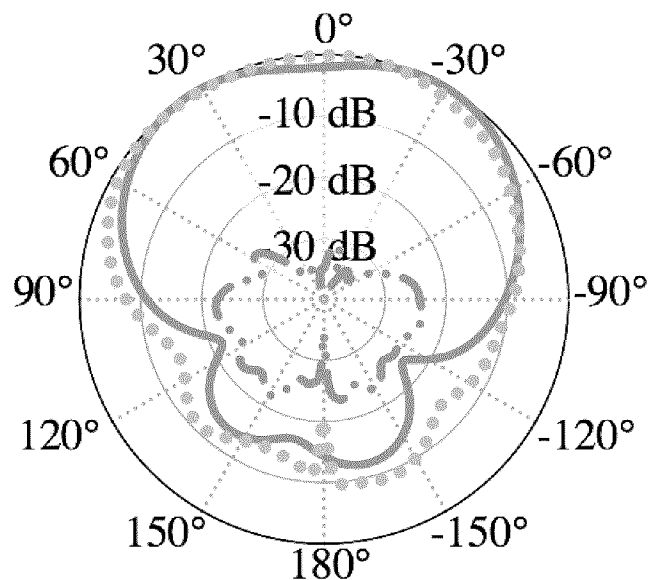


Fig. 45E

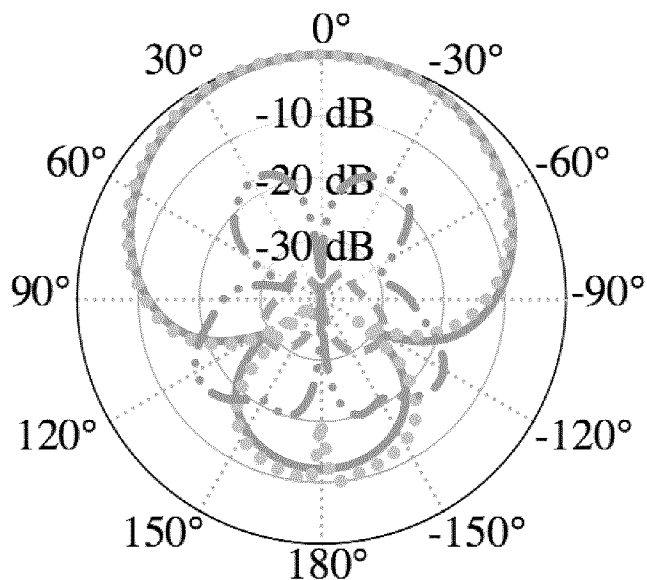


Fig. 45F

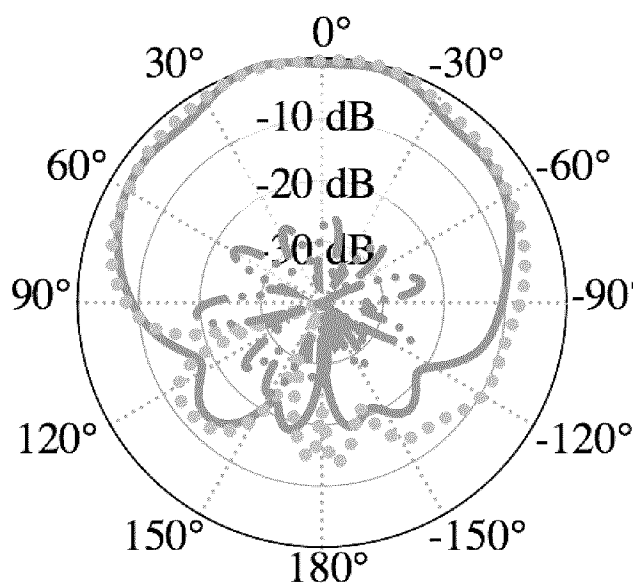


Fig. 45G

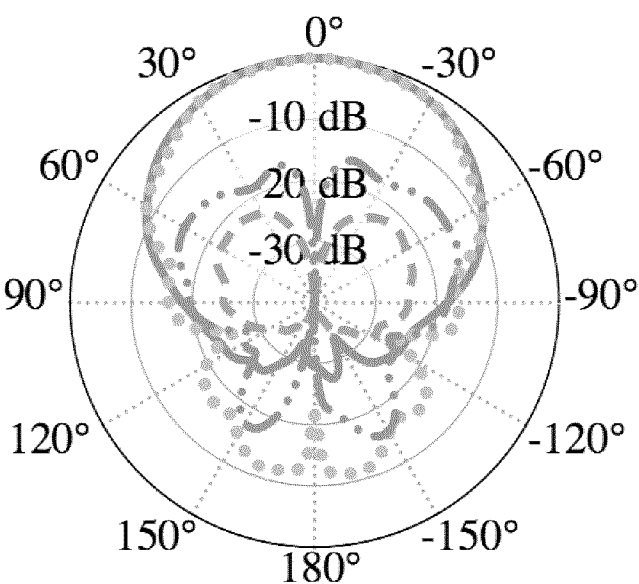


Fig. 45H

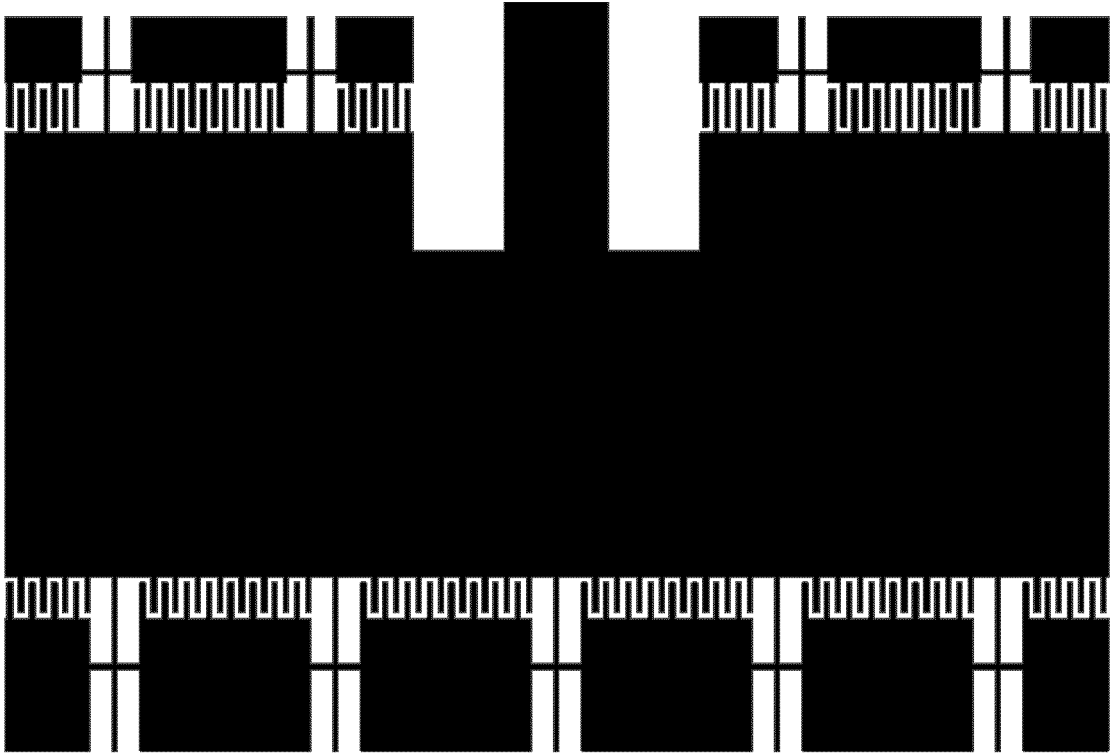


Fig. 46

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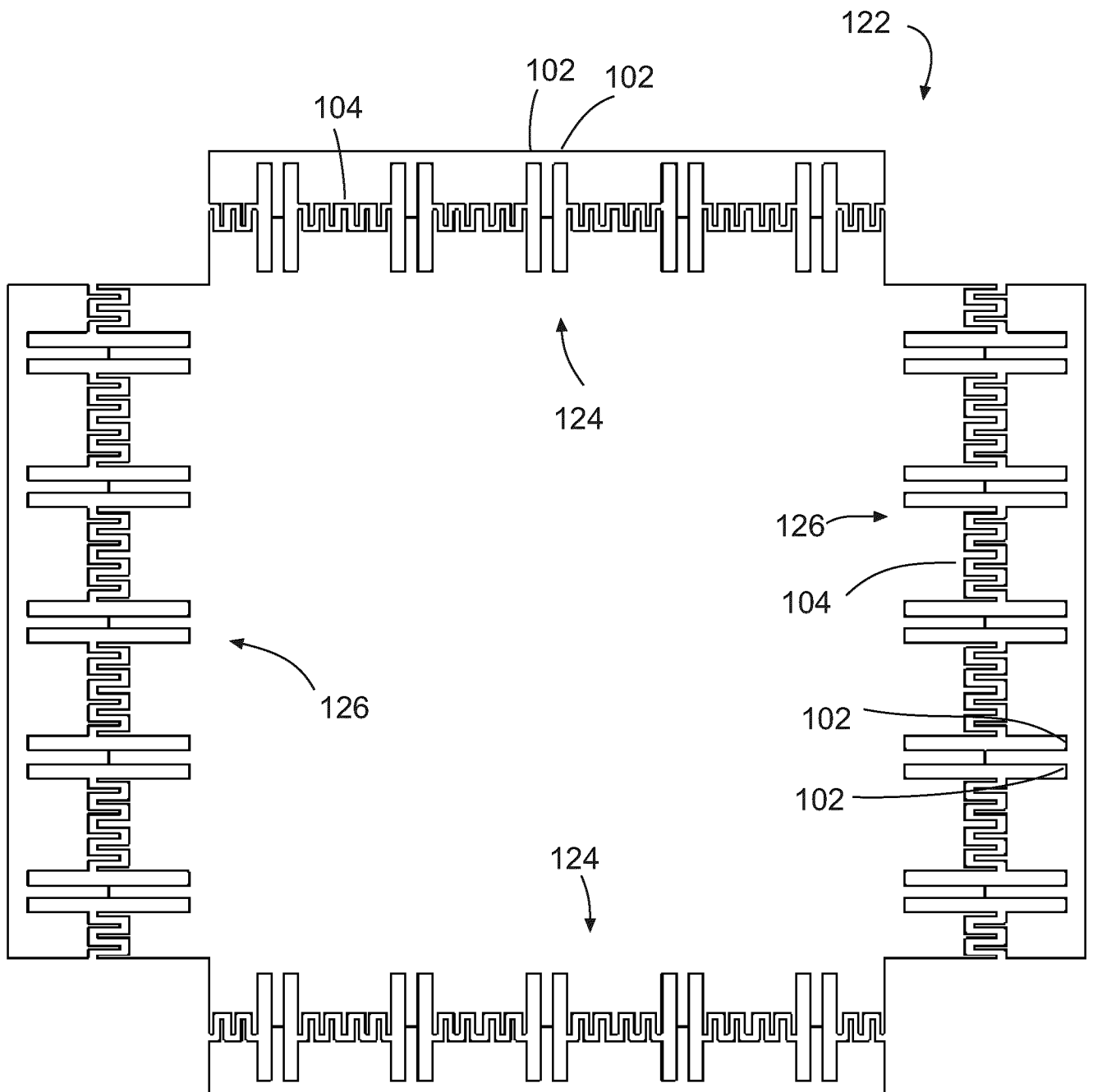


Fig. 47

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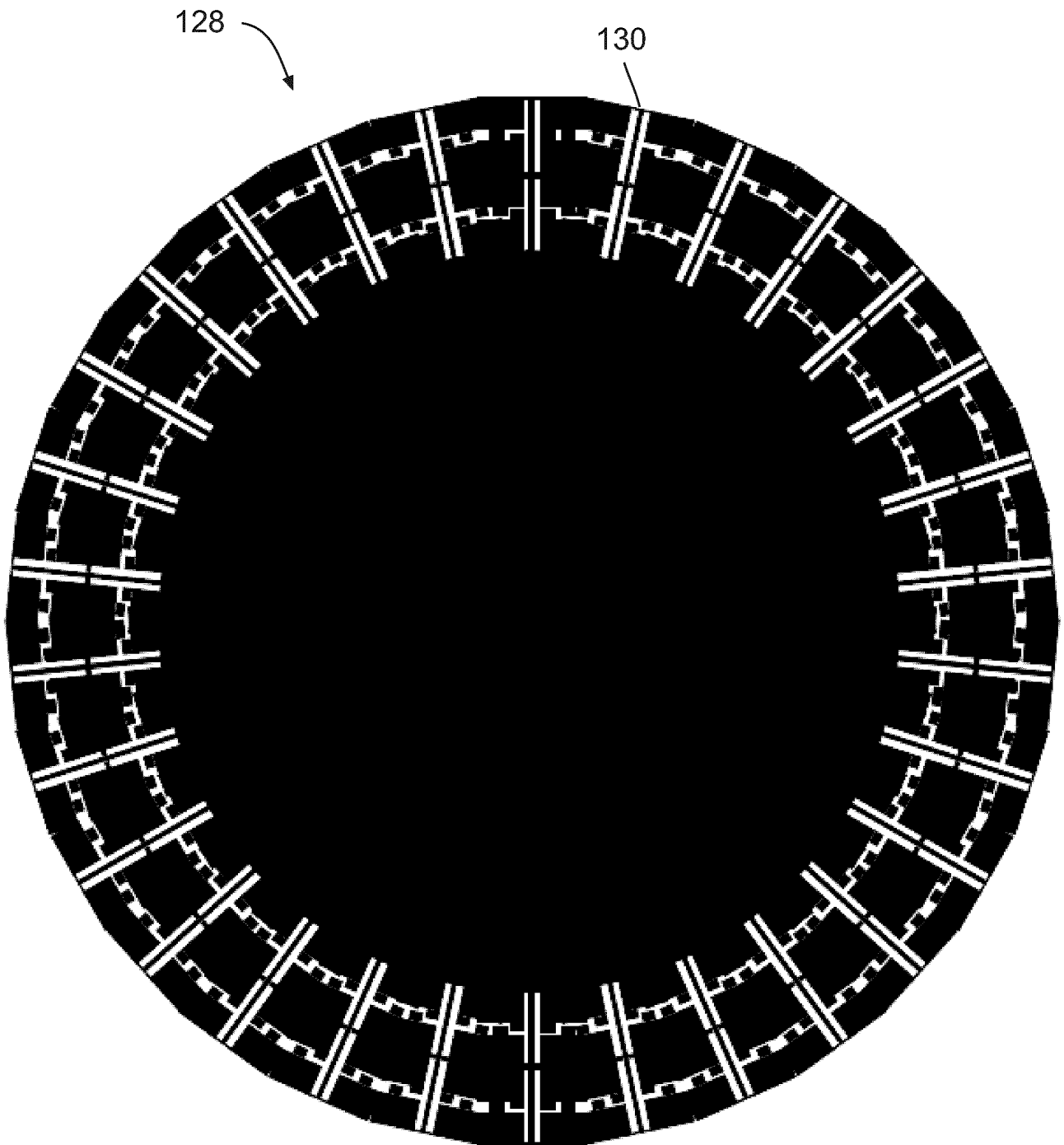


Fig. 48

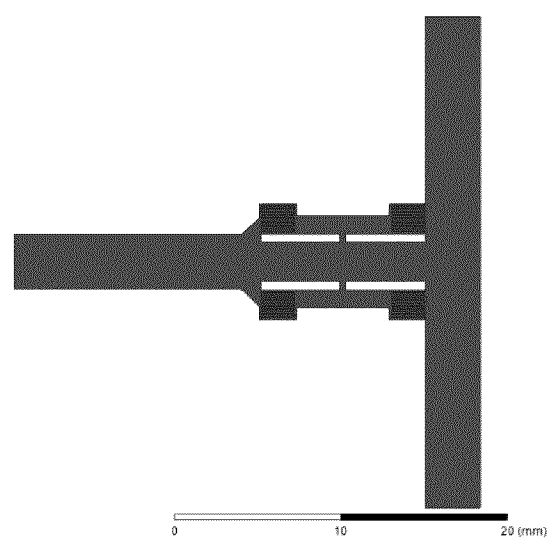


Figure 49: Dual Band power divider with fully printed embedded MTM-EBG.

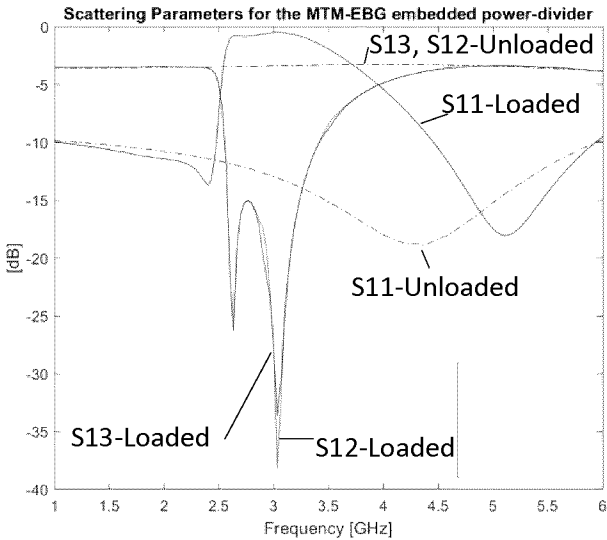


Figure 50: Simulated response

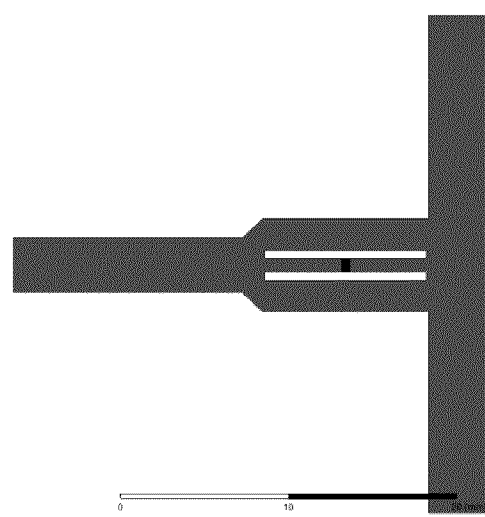


Figure 51: Variable Embedded MTM-EBG in power divider

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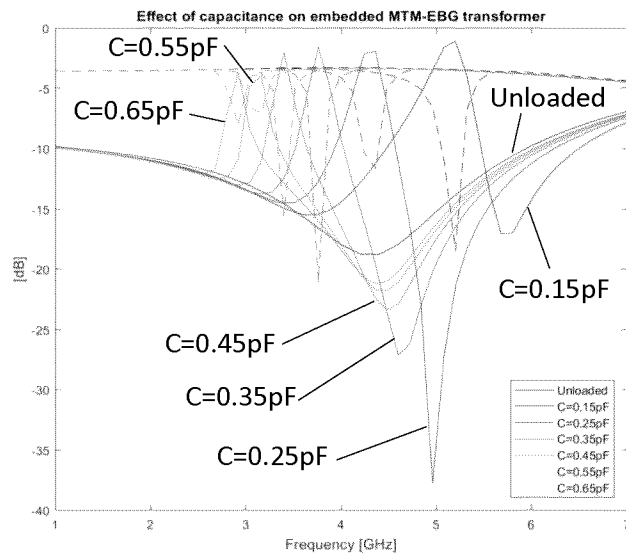
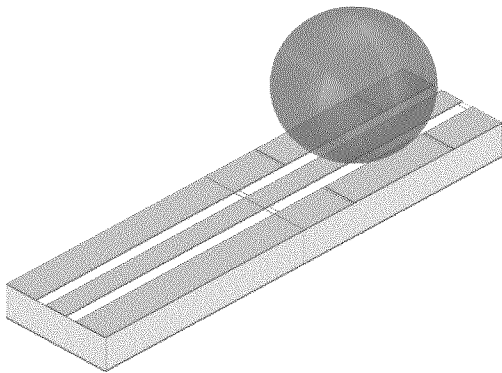
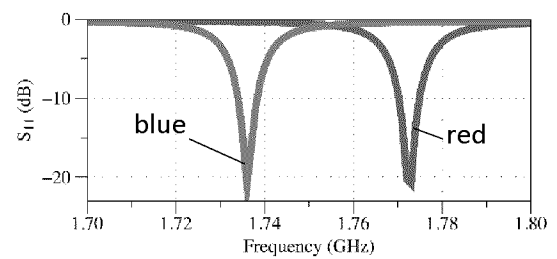


Figure 52: Response to Capacitance variation



(A)



(B)

Fig. 53A: layout of the proposed MTM-EBG-based sensor with a test sample in contact with the structures, 53 B) S_{11} response of the sensor, showing a resonant dip without the sample (red curve) and with the sample present (blue curve).

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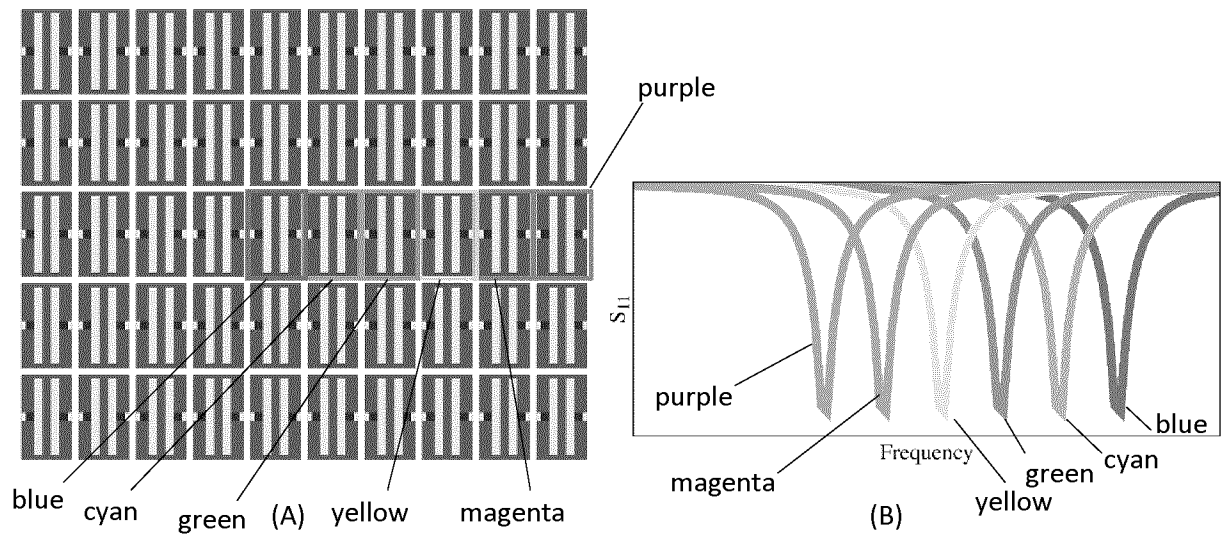


Fig. 54A: layout of a hypothetical two-dimensional sensor array, with the highlighted elements mapped uniquely by color to the resonant frequencies in Fig.54B.

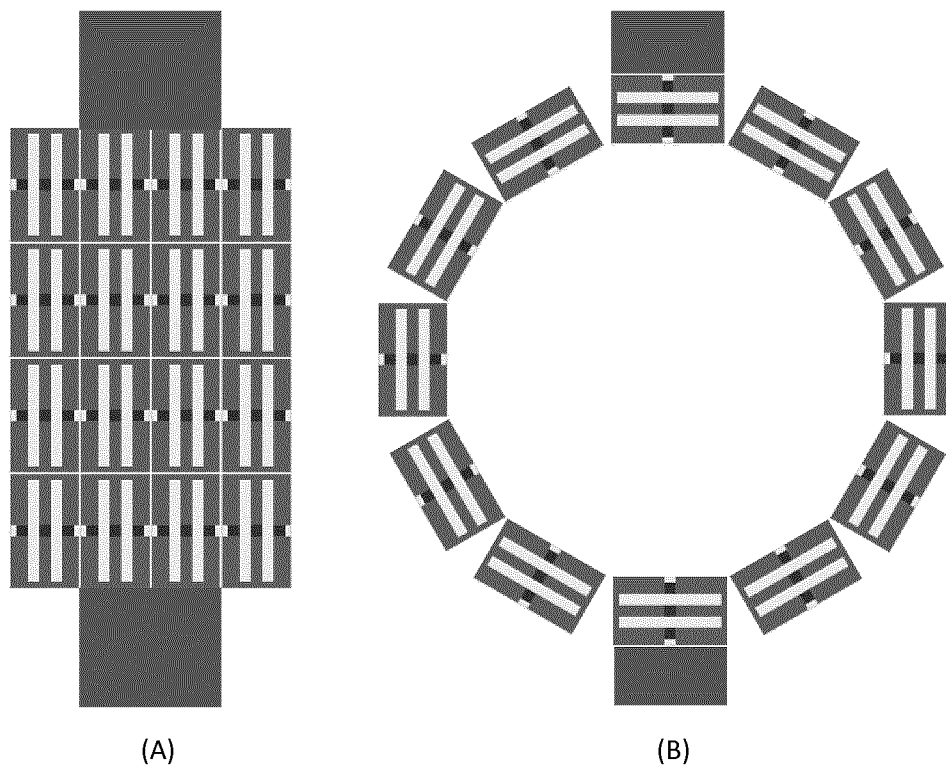


Fig 55: suggested layout of planar resonators using MTM-EBG structures, with feed lines from the top and bottom: 55A) rectangular cavity resonator to replace a SIW cavity, and 55B) MTM-EBG based ring resonator.

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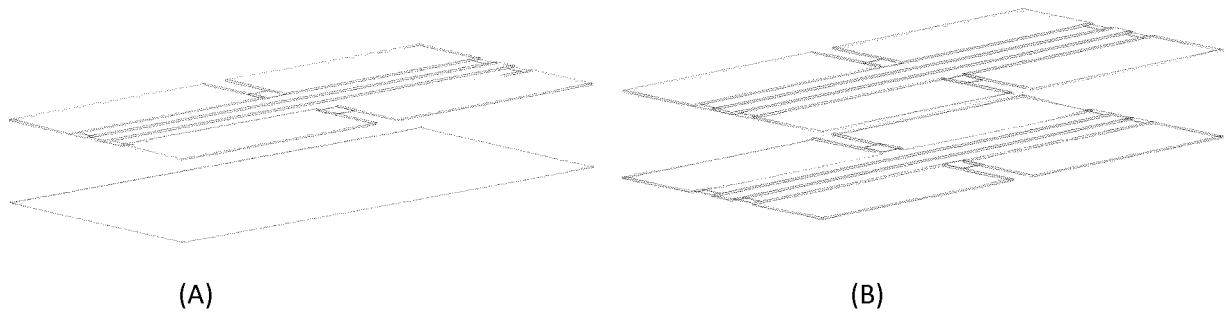


Fig. 56A) Original MTM-EBG structure with a solid conductor ground plane on the back side of the dielectric, and 56B) augmented MTM-EBG structure with patterning on both sides of the dielectric.

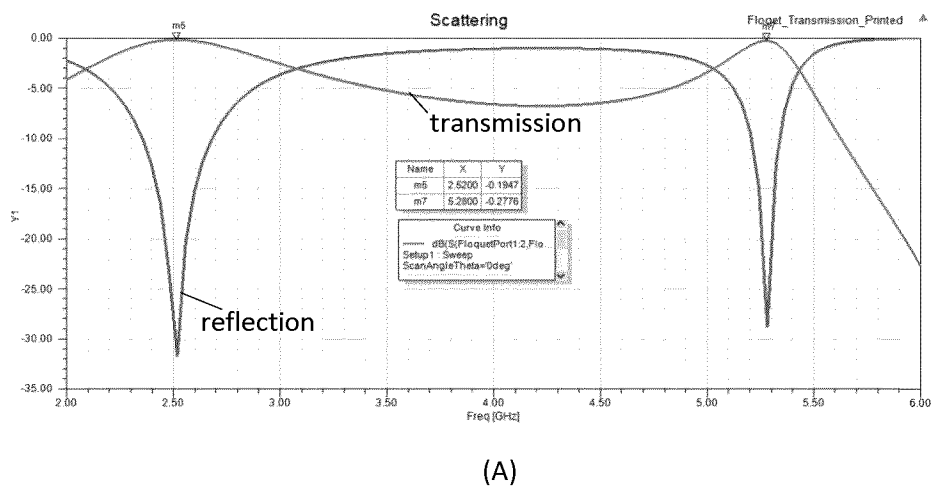


Fig. 57 A) transmission (yellow) and reflection (red) of a proposed structure. Bandwidths and frequencies are bear suitable for the 2.4 and 5.8 GHz ISM bands. 57B) Beamforming occurring as a result of the proposed structure, with incident and outgoing angles being different and indicated with the solid black arrows.

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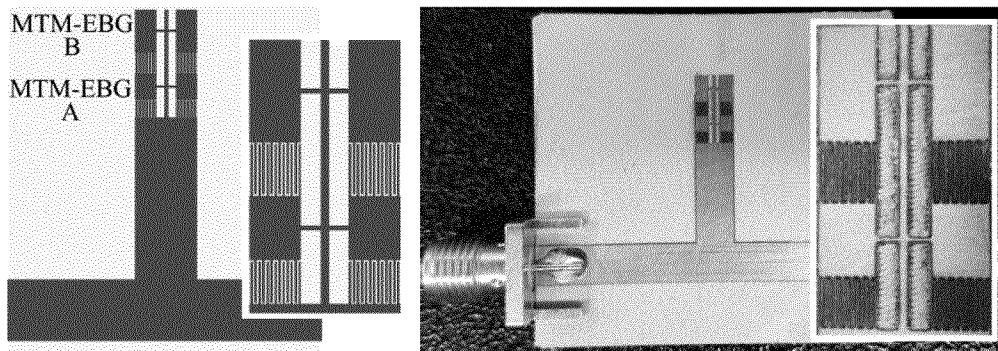


Figure 58: Layout of triple-band stub, left, with fabrication on the right. Close-up of MTM-EBG sections inset.

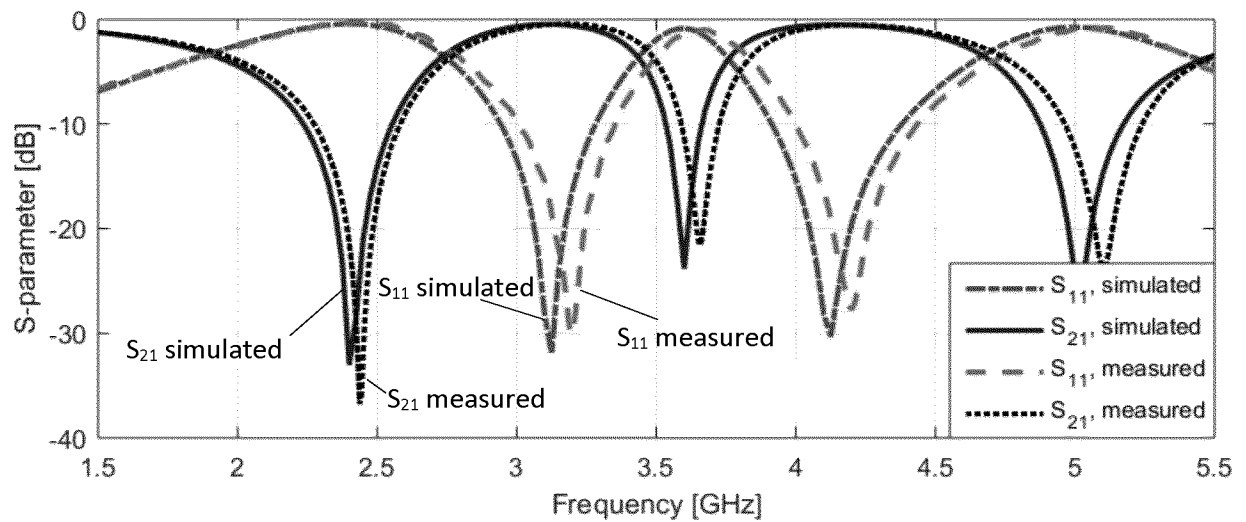


Figure 59: Simulated and measured frequency response of triple-band stub.

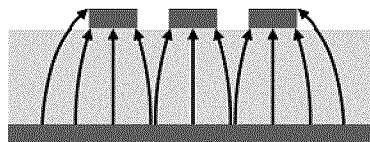


Fig. 60 Example diagrams of the cross-section of the truncated conductor-backed CPW, with fields corresponding to a MS-like mode.

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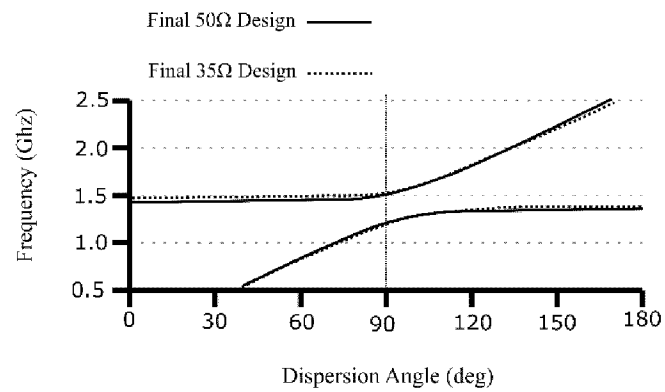


Fig. 61. Dispersion curves of the MS-like mode for the 50 Ω cell (red, solid) and 35 Ω cell (blue, dashed).

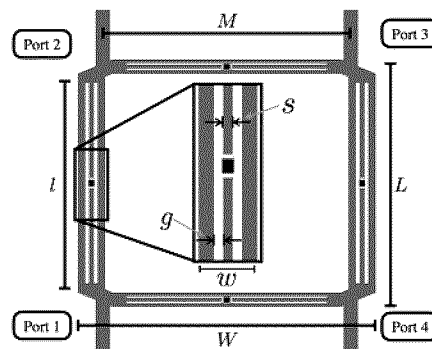


Fig. 62 Layout of the proposed couplers: a) simulated, inset: relevant dimensions of the MTM-EBG unit cell, and b) fabricated coupler.

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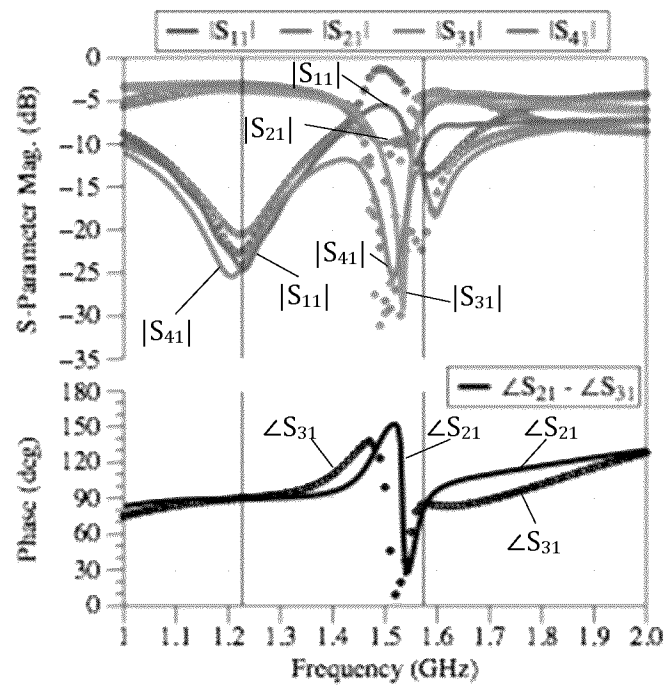


Fig. 63. Scattering parameters (magnitude, top, and phase, bottom) of the simulated (dotted) and measured (solid) scattering parameters. The two vertical black lines correspond to the operating frequencies of GPS L2 (1.228 GHz) and GPS L1 (1.575 GHz).

INTERNATIONAL SEARCH REPORT

International application No.

PCT/CA2019/050638

A. CLASSIFICATION OF SUBJECT MATTER

IPC: **H01P 3/00** (2006.01), **H01F 5/00** (2006.01), **H01G 4/33** (2006.01), **H01P 1/162** (2006.01),
H01P 1/20 (2006.01)

According to International Patent Classification (IPC) or to both national classification and IPC

B. FIELDS SEARCHED

Minimum documentation searched (classification system followed by classification symbols)

Keywords used across the whole IPC

Documentation searched other than minimum documentation to the extent that such documents are included in the fields searched

Electronic database(s) consulted during the international search (name of database(s) and, where practicable, search terms used)

Databases: Questel Orbit, Canadian Patent Database.

Keywords: Electromagnetic; waveguide/waveguides; coplanar/co?planar; capacitor/capacitors; inductor/inductors; conductor/conductors; back/back; electromagnetic; bandgap/bandgaps; parallel; metamaterial

C. DOCUMENTS CONSIDERED TO BE RELEVANT

Category*	Citation of document, with indication, where appropriate, of the relevant passages	Relevant to claim No.
X, Y	CA 2936482 A1 (<i>Iyer et al.</i>) – 19 January 2018 (19-01-2018) * [0021]-[0022]; [0064]; [0072]; [0086]; [0089]-[0093]; [0095]; [0097]-[0098]; [00118]; [00168]; [00176]; [00187]; Figures 1-5; claims 1 and 7 *	1-30
A	US 2007/0090398 A1 (<i>McKinzie</i>) – 26 April 2007 (26-04-2007) * the whole document *	
A	Mao, S. and Chen, M., <i>Propagation Characteristics of Finite-Width Conductor-Backed Coplanar Waveguides With Periodic Electromagnetic Bandgap Cells</i> , In IEEE TRANSACTIONS ON MICROWAVE THEORY AND TECHNIQUES, vol. 50, No. 11, Pages 2624-2628, November 2002 * * the whole document *	
A	US 7,030,463 B1 (<i>Subramanyam et al.</i>) – 18 April 2006 (18-04-2006) * the whole document *	

☐ Further documents are listed in the continuation of Box C.

☒ See patent family annex.

* "A" "D" "E" "L" "O" "P"	Special categories of cited documents: document defining the general state of the art which is not considered to be of particular relevance document cited by the applicant in the international application earlier application or patent but published on or after the international filing date document which may throw doubts on priority claim(s) or which is cited to establish the publication date of another citation or other special reason (as specified) document referring to an oral disclosure, use, exhibition or other means document published prior to the international filing date but later than the priority date claimed	"T" "X" "Y" "&"	later document published after the international filing date or priority date and not in conflict with the application but cited to understand the principle or theory underlying the invention document of particular relevance; the claimed invention cannot be considered novel or cannot be considered to involve an inventive step when the document is taken alone document of particular relevance; the claimed invention cannot be considered to involve an inventive step when the document is combined with one or more other such documents, such combination being obvious to a person skilled in the art document member of the same patent family
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Date of the actual completion of the international search
17 July 2019 (17-07-2019)

Date of mailing of the international search report
24 July 2019 (24-07-2019)

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Authorized officer

Reginald Linco (819) 639-8198

INTERNATIONAL SEARCH REPORT
Information on patent family members

International application No.
PCT/CA2019/050638

Patent Document Cited in Search Report	Publication Date	Patent Family Member(s)	Publication Date
CA2936482A1	19 January 2018 (19-01-2018)	None	
US2007090398A1	26 April 2007 (26-04-2007)	US2007090398A1 US7626216B2 CA2622456A1 CA2622456C EP1938674A2 US2010180437A1 US8595924B2 WO2007047891A2 WO2007047891A3	26 April 2007 (26-04-2007) 01 December 2009 (01-12-2009) 26 April 2007 (26-04-2007) 27 January 2015 (27-01-2015) 02 July 2008 (02-07-2008) 22 July 2010 (22-07-2010) 03 December 2013 (03-12-2013) 26 April 2007 (26-04-2007) 28 February 2008 (28-02-2008)
US7030463B1	18 April 2006 (18-04-2006)	None	