



Universitat Autònoma de Barcelona

ESCOLA TÈCNICA SUPERIOR D'ENGINYERIA

Departament d'Enginyeria Electrònica

**Caracterización de nuevos
resonadores metamaterial,
líneas de transmisión artificiales
y aplicaciones en el diseño de
circuitos de comunicaciones**

**Tesis doctoral realizada por
Francisco Aznar Ballesta**

**Bajo la dirección de
Ferran Martín Antolín**

y

Jordi Bonache Albacete



Bellaterra (Cerdanyola del Vallès), mayo de 2009

Dr. Joan Ferran Martín Antolín y Dr. Jordi Bonache Albacete, profesores del Departament d'Enginyeria Electrònica de la Universitat Autònoma de Barcelona.

CERTIFICAN

Que la memoria titulada "Caracterización de nuevos resonadores metamaterial, líneas de transmisión artificiales y aplicaciones en el diseño de circuitos de comunicaciones", presentada por Francisco Aznar Ballesta para optar al grado de Doctor en Ingeniería Electrónica ha sido realizada bajo su dirección.

Bellaterra (Cerdanyola del Vallès), 26 de mayo de 2009



Dr. Ferran Martín Antolín



Dr. Jordi Bonache Albacete

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Presentación y Motivación

Al comienzo del siglo XXI ha surgido un nuevo campo de investigación en el área del electromagnetismo aplicado y de la ingeniería de microondas, basado en el control de las propiedades electromagnéticas de estructuras artificiales, conocidas como **metamateriales**. Este nuevo concepto de metamaterial agrupa diversas estructuras que suelen ser periódicas. En el caso de los **cristales electromagnéticos**, abarca a los medios en el que el periodo de la estructura es comparable a la longitud de onda de la radiación que se propaga por dicho medio. Para los **medios efectivos**, las celdas que forman el medio tienen un tamaño mucho menor que la longitud de onda a la frecuencia de operación y de esta forma la señal ve dicho medio como un continuo, sin diferenciar las partes individuales que forman dicho medio. Con este concepto de medio efectivo podemos combinar diferentes estructuras que en conjunto muestren características electromagnéticas inexistentes en la naturaleza, y cuyas propiedades son diferentes a las que tendrían por separado. Este tipo de medio puede estar formado por elementos resonantes cuyas dimensiones son inferiores a la longitud de onda a su frecuencia de resonancia, por lo que se les llama **resonadores eléctricamente pequeños**, lo que nos permite diseñar celdas de reducidas dimensiones en términos de longitud de onda. Debido a que el tamaño de la celda unidad que forma el medio vendrá determinado por el tamaño de los elementos que la componen, reducir el tamaño de estos elementos implica una reducción en el tamaño de la estructura completa. El **objetivo principal** de esta tesis será la **obtención de nuevos resonadores eléctricamente pequeños cuyas dimensiones sean menores a las de los ya existentes, su utilización en líneas de transmisión planares, la caracterización tanto de los resonadores como de las líneas donde se utilicen y su aplicación en el diseño de circuitos de microondas planares de reducidas dimensiones**. La finalidad de esto es la **miniaturización** de circuitos de microondas utilizados en sistemas de comunicación, sin reducir prestaciones e incluso mejorándolas.

Esta tesis se ha redactado siguiendo el formato de “**Compendio de Artículos**”, por lo que para muchos de los resultados se hará referencia a los artículos correspondientes, los cuales se encuentran adjuntados al documento. A continuación se detallan los puntos tratados en cada uno de los capítulos que forman esta tesis:

- En el capítulo 1 se hará una introducción a los metamateriales, en la que se exponen las principales propiedades electromagnéticas que caracterizan este tipo de medios, además de las características y principios de funcionamiento de los resonadores de anillos abiertos. También se presentan las configuraciones más conocidas en el diseño de metamateriales.
- En el capítulo 2 se presentarán y caracterizarán los resonadores eléctricamente pequeños más utilizados en el diseño de estructuras metamaterial, así como los nuevos resonadores que se proponen en esta tesis.
- En el capítulo 3 se muestran los modelos circuitales, el método de extracción de parámetros y la metodología empleada en la caracterización de líneas de transmisión basadas en resonadores eléctricamente pequeños como los CSRRs y SRRs. Se hará hincapié en la dualidad existente entre ambas líneas de transmisión artificiales.
- En el capítulo 4 se presentan diferentes aplicaciones en las que se demuestra la utilidad y posibilidades de aplicación en el diseño y miniaturización de dispositivos de microondas de los nuevos resonadores propuestos.
- Para finalizar se exponen las conclusiones y líneas futuras de investigación.

El trabajo llevado a cabo durante la realización de esta tesis se ha realizado dentro del Grupo de Investigación GEMMA (Grup d'Enginyeria de Microones i Mil·limètriques Aplicat) del Departamento de Ingeniería Electrónica de la Universidad Autónoma de Barcelona. Asimismo, parte de este grupo, GEMMA, constituye el centro CIMITEC (Centre d'Investigació en Metamaterials per a la Innovació en les Tecnologies Electrònica i de Comunicacions) que forma parte de la Red de Centros de Apoyo a la Innovación Tecnológica del CIDEM/COPCA (Centro para la Innovación y el Desarrollo Empresarial) de la Generalitat de Catalunya. CIMITEC desarrolla trabajos de investigación y transferencia de tecnología en el campo de los metamateriales con aplicaciones en ingeniería de microondas. GEMMA y CIMITEC también han pertenecido a la Red de Excelencia de la Unión Europea NoE 500252-2 METAMORPHOSE (*Metamaterials organized for radio, millimeter wave and photonic superlattice*

engineering, Junio 2004 - Mayo 2008), orientada a la investigación y fomento de los metamateriales en Europa y que recientemente ha dado lugar al *Virtual Institute for Artificial Electromagnetic Materials and Metamaterials* (METAMORPHOSE VI AISBL) del cual somos miembros de pleno derecho (*Full Member*). GEMMA y CIMITEC pertenecen a la REME (Red Española de Metamateriales). El desarrollo de las actividades de GEMMA, CIMITEC y en particular el trabajo desarrollado en esta tesis, han sido y son financiadas por diferentes proyectos y contratos con empresas entre los que cabe destacar:

- Proyecto del Plan Nacional METASYSTEMS (TEC2004-04249-C02-01) de la Dirección General de Investigación. Coordinado entre la Universidad Autónoma de Barcelona y la Universidad de Sevilla. Titulado: *Diseño, caracterización y aplicación de estructuras basadas en metamateriales al desarrollo de subsistemas de microondas y milimétricas*.
- Proyecto del Plan Nacional METAINNOVA (TEC2007-68013-C02-02) de la Dirección General de Investigación. Coordinado entre la Universidad Autónoma de Barcelona y la Universidad de Sevilla. Titulado: *Tecnologías basadas en metamateriales y su aplicación a la innovación en componentes y subsistemas de RF microondas y milimétricas: circuitos de radiocomunicación*.
- Proyecto Europeo Eureka METATEC: *METAmaterial-based TEchnology for broadband wireless Communications and RF identification*. Consorcio formado por la University of Novi Sad (Serbia), la empresa IMTEL Mikrotalasi A.D. (Serbia), el *Institute of Microwave Techniques and Electronics - IMTEL Institute* (Serbia), la empresa AIDA Centre s.l. (España) y la Universidad Autónoma de Barcelona – CIMITEC (España).
- Proyecto CONSOLIDER INGENIO 2010 (CSD2008-00066) del Ministerio de Ciencia e Innovación. Titulado: *Ingeniería de Metamateriales (EMET)*.
- Varios contratos con SEIKO EPSON Corporation y EPSON Europe Electronics GmbH para el desarrollo de filtros paso banda de banda ancha.

ÍNDICE

Capítulo 1

Introducción

1.1 Metamateriales

A la hora de hablar de Metamateriales, la primera pregunta que surge es: ¿Qué son los metamateriales? Si buscamos en la bibliografía, vemos que no existen muchos libros especializados sobre este tema y los que hay son muy recientes [1–8]. Esto es debido a que no hace mucho (desde principios del siglo XXI) que se acuñó este término para referirse a estructuras con características electromagnéticas u ópticas que no se observan en la naturaleza (como por ejemplo estructuras con índice de refracción negativo, o estructuras en las que la propagación de las ondas es contraria al del flujo de propagación de la energía, propagación zurda). Debido a que hace poco tiempo que se usa este término, la comunidad científica aún no se ha puesto de acuerdo en una definición universal para el término metamaterial. Aun con la ausencia de una definición oficial y universal, podemos decir que un metamaterial es un material creado artificialmente que presenta propiedades electromagnéticas inusuales y que no suelen aparecer en la naturaleza. Estas propiedades son debidas al diseño de la estructura y no a los elementos que la forman. Es decir, la estructura completa, en su conjunto, tiene un comportamiento diferente al que tendrían sus elementos por separado. A partir de este punto aparece cierta controversia con la definición de metamaterial. Hay quien considera que un metamaterial es una estructura periódica como la de la referencia [9], en la cual la dimensión de la estructura que se repite es netamente menor que la longitud de onda con la que se trabaja. De esta forma, con una arquitectura adecuada de las inclusiones, la estructura se comporta como un medio continuo y se puede controlar su comportamiento electromagnético. También hay quien dice que el tamaño no es un condicionante para los metamateriales, aceptando como metamateriales estructuras con dimensiones de celda unidad comparables a la longitud de onda, como los cristales fotónicos [10–13]. Otros incluyen dentro de

los metamateriales a estructuras aleatorias [14]. Existe también quien restringe la definición de los metamateriales a aquellos cuyo índice de refracción es negativo [15]. En nuestro caso, consideraremos metamateriales aquellas estructuras artificiales cuyo tamaño de celda es menor que la longitud de onda de la frecuencia de trabajo, cuyas propiedades electromagnéticas podemos controlar y que nos permite, en el caso de que sea conveniente, obtener transmisión zurda. La característica, por tanto, más importante de estas líneas es la controlabilidad de las propiedades electromagnéticas además de la escalabilidad dentro de su rango de operación.

La segunda pregunta más común sobre los metamateriales es: ¿para qué sirven? Tal y como hemos dicho antes, el rango de operación de los metamateriales es muy amplio y sus límites están por averiguar a día de hoy. Si nos movemos dentro del rango de las microondas, que será nuestro caso en este trabajo, los metamateriales son usados para la miniaturización y mejora de prestaciones de diferentes dispositivos de telecomunicaciones, como filtros [16–19], divisores de potencia [20–22] y antenas [23–26], entre otros. Debido a la posibilidad de obtener estructuras con índice de refracción negativo es posible la creación de *superlentes* [27–30], que pueden ser utilizadas para mejorar de una forma notoria la calidad de imágenes que pueden ser usadas en diagnósticos médicos. Y cómo no, el tema más común cuando se habla de metamateriales es la invisibilidad. En este tema cabe comentar que se ha conseguido hacer invisible un objeto mediante un “escudo metamaterial”. Esto se ha demostrado experimentalmente a frecuencias de microondas y para un estrecho rango de frecuencial muy específico [31], lo que ha supuesto un paso en la investigación sobre la invisibilidad.

1.2 Medios Efectivos

Un medio efectivo es aquel en el que la longitud de onda de la frecuencia de trabajo es mucho mayor que las partículas o celdas base que forman el medio, de forma que la radiación incidente sobre el medio percibe a éste como un medio homogéneo, en términos de sus propiedades electromagnéticas, sin diferenciar las partículas que forman parte del medio. De esta forma, cabe la posibilidad de que sus propiedades como medio sean diferentes a las propiedades de sus componentes, es decir, las características del medio pueden ser distintas a las características de los elementos que lo forman. Como ejemplo para entender este concepto, es algo parecido a lo que ocurre en un televisor, donde solamente

CAPÍTULO 1. INTRODUCCIÓN

existen tres colores diferentes pero, sin embargo, nosotros percibimos muchos más que son la combinación de estos tres.

De esta forma, para este tipo de medios, se definen las magnitudes electromagnéticas efectivas: la permeabilidad magnética efectiva μ_{eff} y la permitividad dieléctrica efectiva ε_{eff} . Atendiendo al signo de estas dos magnitudes, que caracterizan la propagación de las ondas electromagnéticas en el medio, se pueden diferenciar 4 tipos de medios que se representan en la figura 1.1, sin tener en cuenta las pérdidas. En el primer cuadrante de la figura 1.1 correspondiente a μ_{eff} y ε_{eff} positivas, se encuentran los llamados *right-handed materials* (RHM), *double-positive* (DPS), o medios diestros, entre los que se encuentran los dieléctricos convencionales, aún cuando éstas no sean estructuras artificiales. Existen otros dos grupos de medios en los cuales sólo uno de los parámetros es negativo. Son los denominados *single negative* (SNG) *media* (pudiendo ser *epsilon-negative* (ENG) o *mu-negative* (MNG), que se encuentra en el segundo y cuarto cuadrante, respectivamente) y en los que no es posible la propagación. Existe otro grupo en el que ambos parámetros son negativos simultáneamente. En este caso la propagación sí es posible, como también lo es en los DPS ya que en estos medios, al contrario que en los SNG, la constante de propagación es real; dichos medios son los denominados *double negative* (DNG) *media*, *left-handed materials* (LHM) o medios zurdos [15].

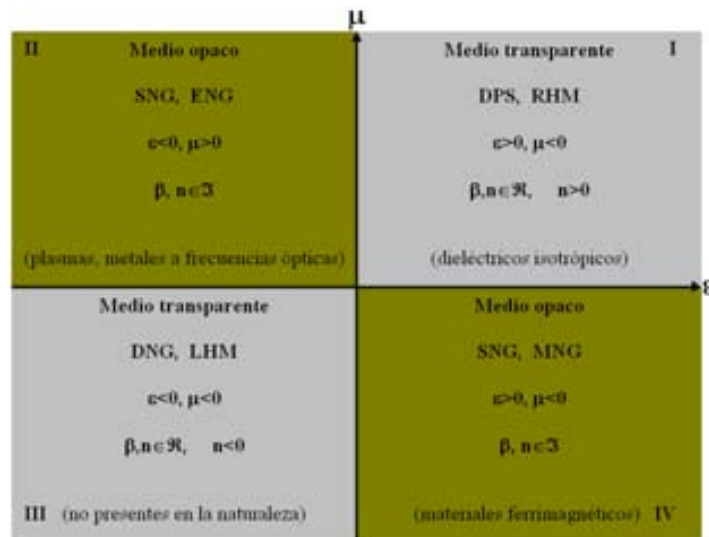


Figura 1.1: Diagrama $\varepsilon - \mu$ para la clasificación de medios en función del signo de los parámetros que regulan la propagación de ondas electromagnéticas.

1.2. MEDIOS EFECTIVOS

Como hemos dicho antes los medios del tercer cuadrante no han existido hasta hace pocos años [9], pero sus propiedades se conocen desde hace tiempo. En 1968, Veselago [15] publicaba un artículo en el cual se hacía un estudio meramente teórico de las características de propagación de un hipotético medio que presentara simultáneamente parámetros ε y μ negativos. Analizando la ecuación de dispersión para un medio isótropo:

$$k^2 = \frac{\omega^2}{c^2} n^2 \quad (1.1)$$

donde

$$n^2 = \varepsilon\mu \quad (1.2)$$

y k^1 es la constante de propagación (o constante de fase), ω es la frecuencia angular, c es la velocidad de la luz en el vacío y n es el índice de refracción del medio, se observa que si ε y μ toman valores negativos simultáneamente, no se impide la propagación, ya que se sigue proporcionando una constante de propagación real.

En las ecuaciones 1.1 y 1.2, un cambio simultáneo en el signo de ε y μ no produce cambios. Para poder analizar las consecuencias del cambio de signo de dichos parámetros, hay que analizar aquellas relaciones donde aparecen estos términos por separado, como hizo Veselago [15]. Éstas son las ecuaciones de Maxwell y las ecuaciones constitutivas:

$$\vec{\nabla} \times \vec{E} = -\frac{\partial \vec{B}}{\partial t} \quad (1.3)$$

$$\vec{\nabla} \times \vec{H} = \frac{\partial \vec{D}}{\partial t} \quad (1.4)$$

$$\vec{B} = \mu \vec{H} \quad (1.5)$$

$$\vec{D} = \varepsilon \vec{E} \quad (1.6)$$

¹Aunque la constante de propagación se denote aquí como k , que es la forma convencional de designar la propagación en medios continuos, más adelante, en este trabajo se citará como β , siendo ésta su forma habitual para líneas de transmisión.

donde \vec{E} es el campo eléctrico, \vec{D} el desplazamiento eléctrico, \vec{H} el campo magnético y \vec{B} la inducción magnética, comúnmente llamado campo magnético B. Considerando la propagación de una onda plana monocromática, las expresiones anteriores se reducen a:

$$\vec{k} \times \vec{E} = \omega \mu \vec{H} \quad (1.7)$$

$$\vec{k} \times \vec{H} = -\omega \epsilon \vec{E} \quad (1.8)$$

De estas dos ecuaciones (1.7 y 1.8) se puede deducir que, para un medio en el que $\epsilon > 0$ y $\mu > 0$, los vectores de los campos eléctrico y magnético (\vec{E} y \vec{H}) y el vector de propagación (\vec{k}) forman un triplete diestro (figura 1.2(a)), como es el caso de los medios convencionales. Mientras que en medios con permitividad y permeabilidad efectiva negativa ($\epsilon < 0$ y $\mu < 0$), \vec{E} , \vec{H} y \vec{k} forman un triplete zurdo (figura 1.2(b)). Por esta razón se les denominan medios zurdos (LHM, del inglés *left handed media*).

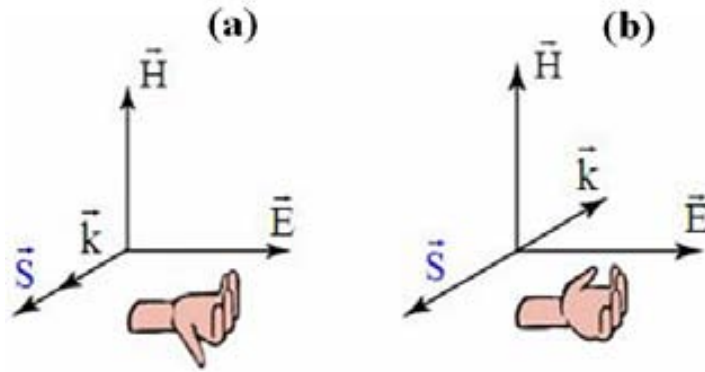


Figura 1.2: Relación entre los vectores \vec{E} , \vec{H} y \vec{k} para un medio convencional diestro (RHM) (a) y un medio zurdo (LHM) (b).

El vector de Poynting, \vec{S} , de una onda que se propaga en un medio, tiene la dirección y sentido de la propagación de la energía y, por tanto, de la velocidad de grupo. Dicho vector viene dado por la ecuación siguiente:

$$\vec{S} = \vec{E} \times \vec{H} \quad (1.9)$$

se puede observar que \vec{S} siempre forma un triplete diestro con los vectores de campo \vec{E} y \vec{H} (figura 1.2), sea cual sea el signo de los parámetros ε y μ . Por lo tanto, en el caso de un medio diestro, \vec{S} y \vec{k} tendrán el mismo sentido, mientras que en el caso de un medio zurdo sus sentidos serán opuestos. Así pues, en un medio zurdo la velocidad de fase, paralela a \vec{k} , y la velocidad de grupo, paralela a \vec{S} , tendrán sentidos opuestos [15].

Como ya sabemos, una característica singular de este tipo de medios es que su índice de refracción se hace negativo. En la ecuación 1.2, el índice de refracción n presenta una ambigüedad en el signo de la raíz cuadrada que se resuelve mediante un análisis adecuado [32]. Por ejemplo, si es expresado en coordenadas polares ($\varepsilon = \mu = -1 = \cos(\pi) + i \operatorname{sen}(\pi) = e^{i\pi}$), la expresión 1.2 del índice de refracción, puede expresarse como:

$$n = \sqrt{\varepsilon\mu} = \sqrt{\varepsilon}\sqrt{\mu} = e^{i\pi/2}e^{i\pi/2} = e^{i\pi} = -1 \quad (1.10)$$

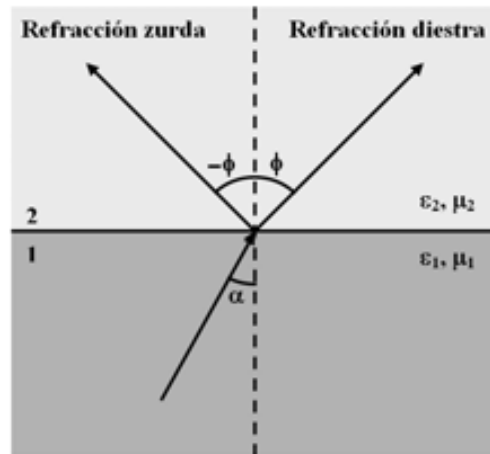


Figura 1.3: Refracción entre dos medios convencionales (refracción diestra) y entre un medio diestro y uno zurdo (refracción zurda). En la refracción negativa, los rayos incidente y refractado caen al mismo lado de la normal.

Al realizar este análisis es importante tener en cuenta que la raíz cuadrada tanto de ε como de μ ha de tener parte imaginaria positiva, necesario para un

material pasivo. Este cambio en el signo del índice de refracción influye directamente en la ley de refracción de Snell. Si se considera un rayo que incide desde un medio 1, diestro, hacia un medio 2 de características zurdas (ver figura 1.3), como consecuencia de esta inversión en la refracción, si se emplearan metamateriales para la fabricación de lentes, las convergentes se transformarían en divergentes y viceversa. Una propiedad de este tipo de lentes con índice de refracción negativo, es la posibilidad de focalizar una fuente puntual con una resolución menor que la longitud de onda [28], es decir, focalizar de manera perfecta. Aunque, como se comenta en [33], algunos investigadores sugieren que al introducir las pérdidas en el material desaparece la focalización sub-lambda. Otros fenómenos tales como el efecto Doppler y la radiación Cherenkov se invierten [15, 34].

1.3 Resonadores eléctricamente pequeños

A finales de los años 90, John Pendry introduce el resonador de anillos abiertos (*Split Ring Resonators* o SRR) [35]. Su aparición supuso el comienzo de la investigación en el campo de los metamateriales, ya que resultó ser la primera estructura resonante no magnética capaz de proporcionar valores negativos de la permeabilidad magnética efectiva cuando se disponen en una estructura periódica. En la figura 1.4(a) se muestra su estructura. Está formado por dos anillos metálicos concéntricos en los cuales se han practicado unas aberturas en posiciones opuestas una respecto de la otra. Si sobre esta estructura se aplica un campo magnético variable en la dirección del eje de los anillos (eje z en la figura 1.4(a)), se inducen corrientes en los anillos. Estas corrientes son capaces de pasar de un anillo a otro gracias a la capacidad distribuida que se forma entre ellos, de forma que se cierra el circuito [36]. El SRR se puede modelar como un tanque resonante LC , donde L modela el comportamiento inductivo introducido por las tiras conductoras y C modela la capacidad distribuida entre los anillos (ver figura 1.4). Las propiedades electromagnéticas de los SRRs han sido estudiadas en diversos trabajos [37–39]. Dichos análisis muestran que los SRRs se comportan como un resonador que puede ser excitado mediante un campo magnético variable axial, paralelo al eje z en la figura 1.4 (a), o por un campo eléctrico variable polarizado en el sentido del eje y [36].

En la figura 1.4(b) se muestra el circuito equivalente para el SRR. Su frecuencia de resonancia puede ser calculada mediante los elementos del tanque $L_S C_S$ que modelan el SRR. La inductancia L_S puede ser aproximada por la que

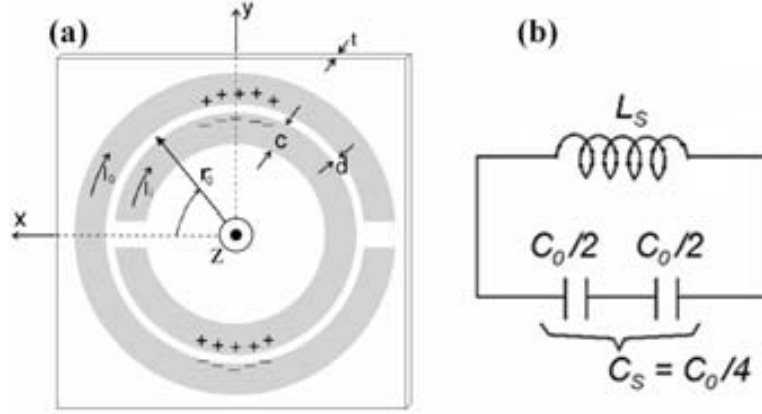


Figura 1.4: Esquema de las dimensiones más relevantes del SRR (a), modelo eléctrico (b). (a) c ancho de las tiras metálicas, d separación entre tiras, r_0 radio medio de ambas tiras metálicas, t grosor de sustrato sobre el que se implementa el resonador, I_o e I_i indican el sentido de las corrientes de ambas tiras de metal y los signos positivos y negativos indican la distribución de cargas en la primera resonancia del anillo. (b) L_S representa la inductancia del modelo eléctrico del SRR, C_0 representa la capacidad total entre tiras y C_S la conexión serie de las capacidades de las mitades superior e inferior del SRR.

tendría un anillo con un radio r_0 de valor medio de los dos anillos que forman el SRR y cuya anchura, c , fuera la de dichos anillos [40]. Cuando se desprecia la capacidad que aparece entre los extremos de las aperturas de los anillos, C_0 representa la capacidad total entre anillos y C_S corresponde a la conexión serie de las capacidades de las mitades superior e inferior del SRR (figura 1.4). La capacidad aproximada (C_0) está calculada teniendo en cuenta su distribución a lo largo del perímetro de los anillos, a partir de la capacidad estática por unidad de longitud C_{pul} entre dos tiras metálicas paralelas de ancho c y distancia d [37]. Este análisis para el SRR es aproximado; en [40] se presenta un método numérico con mayor precisión, que puede ser implementado mediante un programa informático para calcular L_S , C_S y la frecuencia de resonancia del SRR, f_0 , a partir de sus dimensiones físicas [41]. En las siguientes ecuaciones están representadas las capacidades y la frecuencia de resonancia del SRR.

$$C_0 = 2\pi r C_{pul} \quad (1.11)$$

$$C_S = \frac{\pi r C_{pul}}{2} \quad (1.12)$$

$$f_0 = \frac{1}{2\pi\sqrt{L_S C_S}} \quad (1.13)$$

Como ya se ha dicho, el SRR puede ser excitado mediante un campo magnético variable axial. Hemos de destacar que el SRR presenta una longitud eléctrica menor que la longitud de onda de la radiación incidente a su primera frecuencia de resonancia. Si la distancia entre tiras (d) es suficientemente pequeña, la capacidad (C_S) tomará valores elevados que proporcionarán una frecuencia de resonancia baja (ecuación 1.13). Esta frecuencia está por debajo de las frecuencias de resonancia de las partículas (anillos) que forman el resonador, reduciendo la longitud eléctrica del resonador, en términos de longitud de onda.

Introduciendo modificaciones en la topología de la figura 1.4(a) se pueden obtener nuevos resonadores como, por ejemplo, el *nonbianisotropic split ring resonator* (NB-SRR) y el *spiral resonator* (SR) [40, 42], representados junto a sus circuitos equivalentes en la figura 1.5. En el capítulo 2 analizaremos detenidamente diferentes topologías de resonadores eléctricamente pequeños.

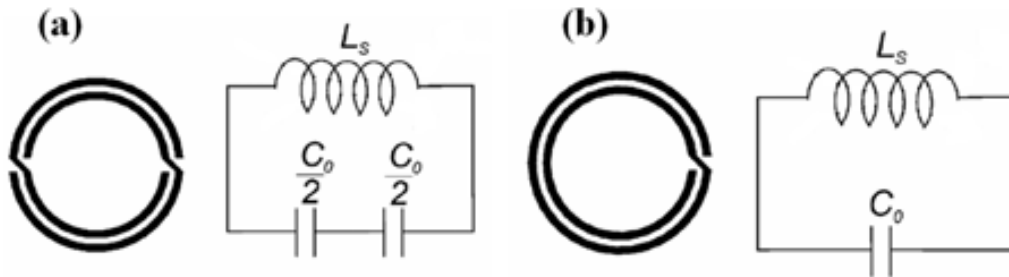


Figura 1.5: Esquemas y circuitos equivalentes correspondientes al NB-SRR (a) y al SR (b). L_S representa la inductancia del modelo eléctrico y C_0 representa la capacidad total entre tiras.

1.4 Medios zurdos

Debido a que en la naturaleza no existen medios conocidos con permitividad (ε_{eff}) y permeabilidad (μ_{eff}) efectiva negativa simultáneamente (medios zurdos), para poder obtener un medio con estas características hay que crearlo combinando dos medios que tengan estas propiedades en un mismo rango frecuencial por separado.

1.4. MEDIOS ZURDOS

Debido al carácter resonante de los SRRs, si se construye una estructura periódica y se ilumina mediante una señal incidente polarizada de forma que el campo magnético sea perpendicular a la superficie de los SRR (paralelo al eje z de los SRRs, figura 1.4), se observa que, para un cierto rango frecuencial, se inhibe la propagación de la señal. Dicho rango frecuencial es aquel en el que los SRRs dan lugar a valores negativos de μ_{eff} [35] (lo cual ocurre en un estrecho rango frecuencial por encima de la resonancia). Si por otro lado se construye una estructura periódica de hilos metálicos delgados, y se ilumina con una señal cuyo campo eléctrico tiene la misma dirección que los hilos, se obtiene una zona a bajas frecuencias donde no se permite la propagación y que corresponde a una ϵ_{eff} negativa [43]. En estos dos medios que se acaban de nombrar se puede controlar fácilmente, mediante las diferentes variables físicas del medio, las frecuencias donde se obtiene ϵ_{eff} y μ_{eff} negativos, de forma que si se diseñan ambos medios para que esos rangos frecuenciales coincidan y combinándolos como hizo Smith, se obtiene un medio con ϵ_{eff} y μ_{eff} negativos simultáneamente [9]. En la figura 1.6 se muestra la fotografía de este primer medio con características zurdas junto a su respuesta frecuencial.

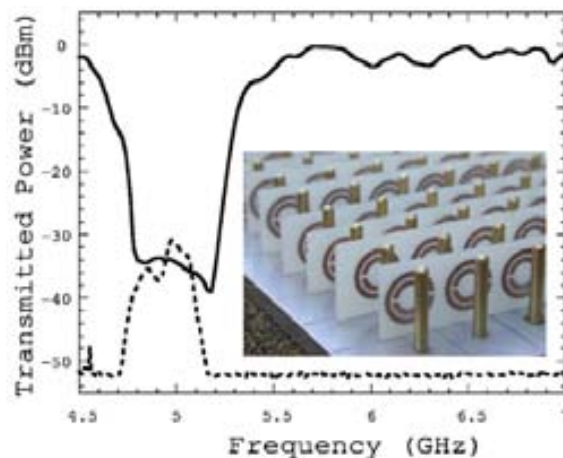


Figura 1.6: Fotografía del primer medio compuesto con características zurdas junto a la transmisión medida de este medio (línea punteada) y la del medio compuesto por solamente SRRs (línea continua). Gráfica extraída de [9].

Desde que Smith construyera este primer medio zurdo, ha habido muy diversas configuraciones para obtener medios con estas características, como el de la figura 1.7(a) [44]. Otra forma de obtener propagación zurda es la que propuso Marqués en [45], donde una guía de ondas se comporta como un medio

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con ε_{eff} a bajas frecuencias (guía de corte), combinada con SRRs con μ_{eff} negativa, obteniendo una banda de paso gracias a la propagación zurda, como se muestra en la figura 1.8. En la figura 1.7(b) se muestra el primer medio metamaterial que se ha construido con el fin de hacer invisible (en el rango frecuencial de las microondas) un objeto que esté situado en el centro del dispositivo [31]. Recientemente se ha conseguido un medio con índice de refracción negativo a frecuencias ópticas [46]. Estos medios con estructuras tridimensionales están teniendo un gran impacto dentro de la comunidad científica debido a sus posibilidades, como por ejemplo en medicina donde se está utilizando una estructura de este tipo, a modo de lente, para mejorar las imágenes de resonancia magnética [27].

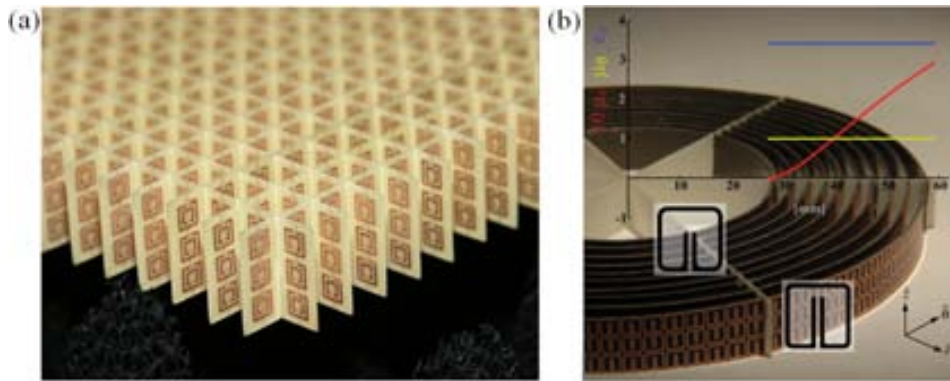


Figura 1.7: Medio bidimensional con características zurdas [44] (a) y primer escudo metamaterial para hacer invisible (en el rango frecuencial de las microondas) un objeto situado en su interior (b) [31].

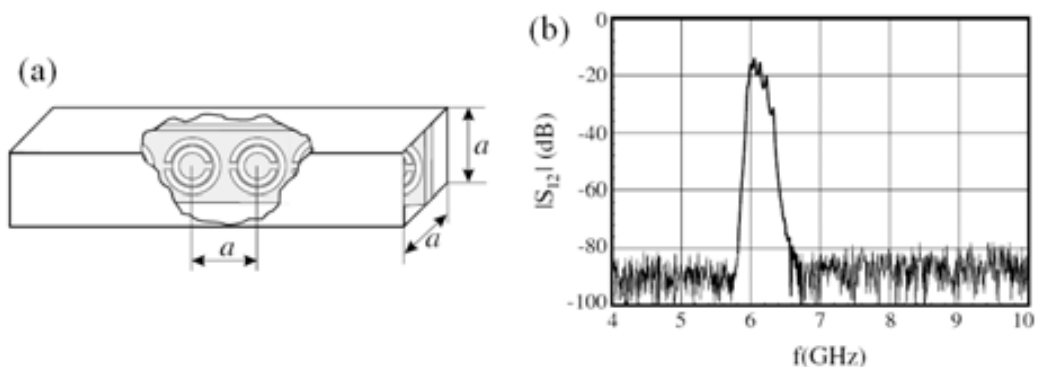


Figura 1.8: Estructura montada para el experimento (a) y coeficiente de transmisión medido (d). Figuras extraídas de [45].

1.5 Medios zurdos en tecnología plana. Líneas de transmisión metamaterial

La posibilidad de utilizar un medio zurdo en tecnología planar abre infinitud de posibilidades, en cuanto a la miniaturización y mejora de prestaciones de componentes y circuitos de microondas basados en líneas de transmisión y stubs (parámetros distribuidos) debido a las pequeñas dimensiones de los elementos que constituyen y por sus singulares propiedades de transmisión. Los medios zurdos vistos en el apartado 1.4 son poco útiles para su utilización en la ingeniería de microondas. En el año 2002 aparecieron diversos trabajos con diferentes estructuras basadas en líneas de transmisión duales, que se comportaban como medios zurdos en un cierto rango frecuencial. El método que se proponía en estos trabajos consiste en cargar líneas de transmisión convencionales, mediante capacidades en serie e inductancias en derivación [47–51]. Al emplear un modelo de línea de transmisión para describir la propagación en un medio, se puede hacer una equivalencia entre la permeabilidad magnética efectiva (μ_{eff}) y la permitividad dieléctrica efectiva (ε_{eff}) con la impedancia serie (Z_S) y la admitancia paralelo (Y_P) distribuidas por unidad de longitud, respectivamente [47]:

$$Z_S = j\omega\mu_{eff} \quad (1.14)$$

$$Y_P = j\omega\varepsilon_{eff} \quad (1.15)$$

Una línea de transmisión convencional sin pérdidas, cuyo modelo se representa en la figura 1.9(a), puede cargarse con capacidades en serie e inductancias en derivación, de forma que a aquellas frecuencias a las que estos elementos de carga dominen sobre los elementos propios de la línea, ésta se comportará como una *línea dual* (1.9(b)). En las líneas duales puede producirse transmisión zurda, ya que se obtienen simultáneamente valores negativos de la reactancia serie, χ_S , y la susceptancia en derivación, B_P , (y por tanto de μ_{eff} y ε_{eff}).

En 2002 se presentan líneas de transmisión planares con características zurdas. Grbic *et al.* [51] proponen una estructura en tecnología coplanar como la que se muestra en la figura 1.10(a) y al mismo tiempo Caloz *et al.* [50] proponen una estructura en tecnología microstrip como la de la figura 1.10(b). En ambos casos las líneas convencionales se cargan con capacidades en serie e inductancias en derivación.

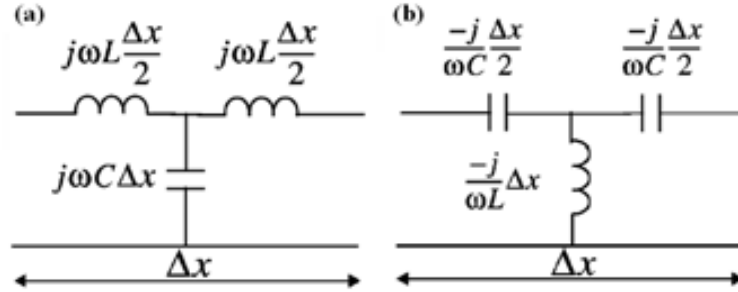


Figura 1.9: Modelos de líneas de transmisión convencional (a) y dual (zurda) (b).

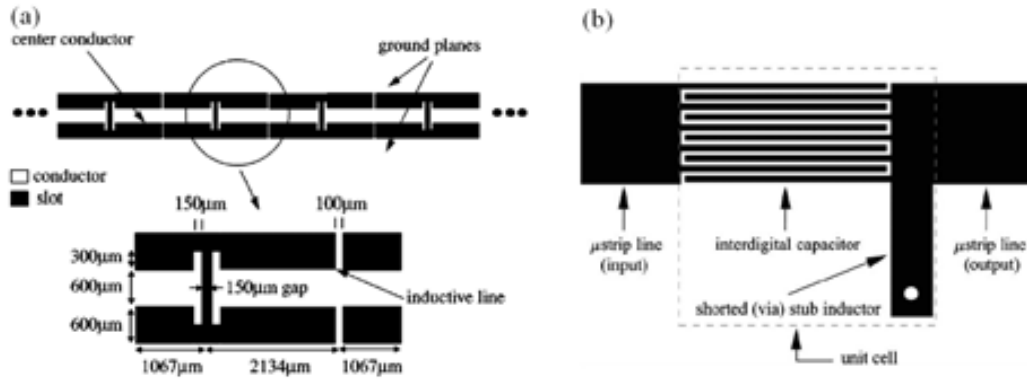


Figura 1.10: Implementación de una línea de transmisión dual en tecnología coplanar (a) y microstrip (b). Figuras extraídas de [51] y [50] respectivamente.

Otro modelo que permite conseguir transmisión zurda ($\epsilon < 0$ y $\mu < 0$) es el llamado modelo resonante. Este modelo consiste en combinar alguno de los elementos que forman la línea dual (capacidades en serie o inductancias en derivación) con resonadores (SRRs o sus versiones complementarias CSRRs) acoplados a una línea de transmisión. Este tipo de líneas fue introducida por Martín *et al.* en 2003 [17]. En dicho trabajo se empleaba una línea coplanar a la que se le acoplaban SRRs e inductancias en derivación tal y como podemos ver en el layout de la figura 1.11 junto al circuito equivalente propuesto en [17]. Este circuito equivalente es el equivalente a media celda unidad, ya que por simetría se puede aplicar la condición de contorno de pared magnética.

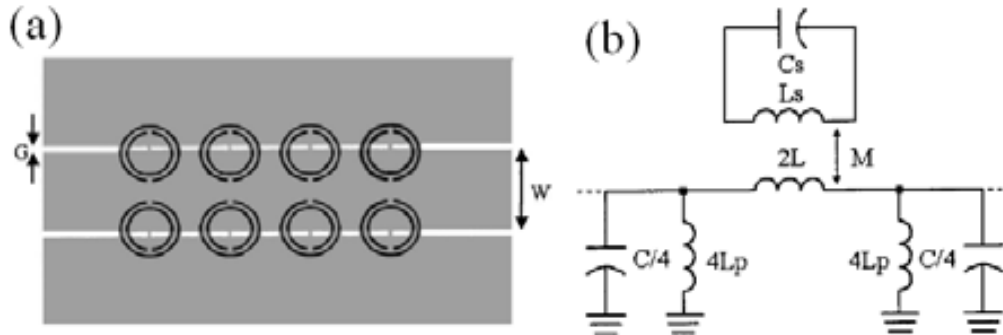


Figura 1.11: Layout de una línea zurda en tecnología coplanar (a) y su circuito equivalente para media celda unidad (b). En gris se representa el metal del plano inferior y en negro el metal en el plano superior del sustrato. L y C son la inductancia y la capacidad de la línea, L_S y C_S la inductancia y la capacidad del anillo, M es la inductancia mutua entre línea y anillo y L_P la inductancia en derivación. Figura extraída de [17].

Otro ejemplo de celda zurda es el que se presenta en la figura 1.12. En él, resonadores de anillos abiertos complementarios (CSRRs) [18] son grabados en el plano de masa de una línea microstrip, de forma que pueden ser excitados mediante el campo eléctrico que se dirige desde la línea de transmisión al plano de masa. Cuando la señal se transmite por la línea, la combinación de CSRRs y gaps capacitivos practicados en la línea y alineados con el centro de los anillos da lugar a una estructura con comportamiento zurdo. En la figura 1.12(a) se observa una imagen en 3D de una de estas celdas unidad, junto a su modelo circuital equivalente. Esta estructura fue propuesta en 2004 por Falcone *et al.* [52].

Con estas estructuras basadas en resonadores, se consiguen características zurdas en un cierto rango de frecuencias donde la impedancia serie es predominantemente capacitiva y la impedancia paralelo es predominantemente inductiva. Los modelos circuitales de estas líneas de transmisión son válidos a aquellas frecuencias tales que la celda unidad presente unas dimensiones sustancialmente menores que la longitud de onda de la señal guiada (λ_g) en su frecuencia de operación, ya que no tiene en cuenta efectos que aparecen a frecuencias superiores, tales como elementos parásitos o resonancias superiores. La revisión del modelo circuital de estas dos estructuras basadas en el modelo resonante de línea de transmisión es uno de los fines de esta tesis y serán tratados con detalle en el capítulo 3.

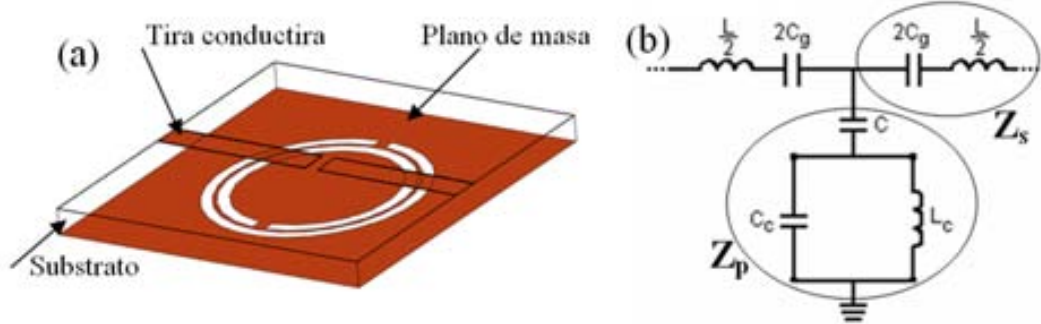


Figura 1.12: Representación tridimensional de la celda base empleada para la síntesis de líneas zurdas basadas en CSRRs (a) y su circuito equivalente (b) en el que se resalta su impedancia serie (Z_S) y paralelo (Z_P). L inductancia de la línea, C_g capacidad serie del gap capacitivo, L_C y C_C inductancia y capacidad del CSRR y C acoplamiento eléctrico entre el anillo y línea.

Para determinar las frecuencias en las que una celda es zurda (en el caso de que lo sea) hay que estudiar su fase (βl). La fase βl representa el desfase introducido por la celda completa cuando la señal se propaga por ella, en condiciones de transmisión total, y su valor puede ser calculado a partir de la fase y la magnitud del coeficiente de transmisión (S_{21}), mediante la siguiente ecuación [16]:

$$\cos(\beta l) = \frac{\cos(\phi_{S_{21}})}{|S_{21}|} \quad (1.16)$$

Por otro lado, la fase puede expresarse en función de las impedancias del modelo circuital mediante la expresión:

$$\cos(\beta l) = 1 + \frac{Z_S}{Z_P} \quad (1.17)$$

donde Z_S y Z_P son la impedancia serie y paralelo del circuito equivalente (en T o π) de la celda, respectivamente. La zona zurda corresponde al rango de frecuencias donde βl tiene un valor real que varía entre -180° y 0° , es decir, donde la fase toma valores negativos, siendo su velocidad de grupo ($d\omega/dk$) positiva.

Otra peculiaridad de los medios zurdos que permite identificar cuándo se obtiene propagación zurda es que, al contrario de lo que sucede en el caso de

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propagación convencional o diestra, un aumento de la longitud en una línea de transmisión zurda se traduce en un aumento del desfase de la señal que se transmite por esa línea. De esta forma, si tenemos una línea de transmisión zurda formada por n celdas, a medida que añadamos nuevas celdas a la línea el desfase de la señal que se transmite por dicha línea aumenta, al contrario de lo que ocurre en una línea convencional. Así, si se consideran dos líneas iguales formadas por distinto número de celdas, n_a y n_b con $n_a > n_b$, cada una de ellas tendrá una fase Φ (Φ_a, Φ_b) y una longitud l ($l_a > l_b$). A partir de la ecuación 1.18, puede deducirse que si las líneas son zurdas y, por tanto, la constante y la velocidad de fase son negativas, es decir, $\beta < 0$ y $v_\Phi < 0$, la diferencia de fase entre ellas será positiva, $\Delta\Phi > 0$, y si las líneas son diestras se obtiene $\Delta\Phi < 0$.

$$\Delta\Phi = \Phi_a - \Phi_b = -\beta(l_a - l_b) = -\frac{\omega(l_a - l_b)}{v_\Phi} \quad (1.18)$$

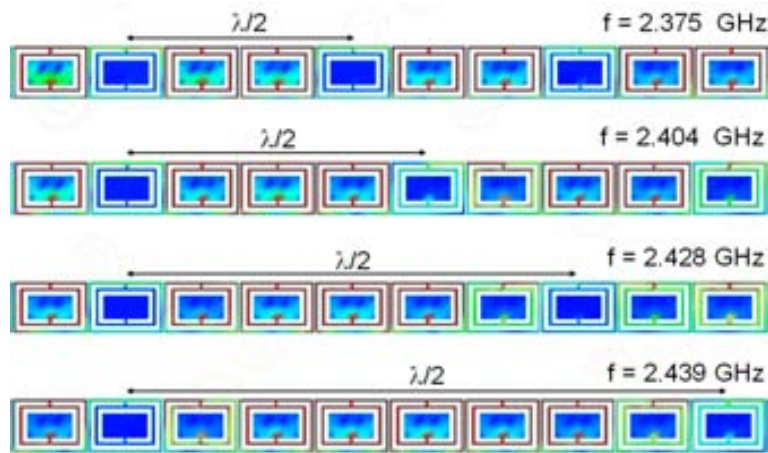


Figura 1.13: Diagramas de corrientes de una línea de transmisión zurda a diferentes frecuencias dentro de una banda de transmisión zurda. A medida que la frecuencia aumenta, la longitud de onda (λ) también aumenta.

También puede identificarse una banda de transmisión zurda si se comprueba que, dentro de dicha banda, un aumento en frecuencia supone un aumento en la longitud de onda de la señal que se transmite por la línea de transmisión. Esta comprobación puede hacerse, por ejemplo, estudiando el diagrama de densidad de corriente de la estructura que forma la línea de transmisión y

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localizando la posición de dos máximos o mínimos consecutivos. Un ejemplo de esto se muestra en la figura 1.13, donde se muestra una línea de transmisión zurda, basada en CSRs donde éstos han sido grabados sobre la propia línea microstrip y combinados con gaps capacitivos [53].

1.5. MEDIOS ZURDOS EN TECNOLOGÍA PLANA

Capítulo 2

Resonadores eléctricamente pequeños

En este capítulo se presentarán resonadores eléctricamente pequeños que se utilizan habitualmente en la síntesis de metamateriales. Se presentarán nuevos resonadores cuyo tamaño ha sido reducido considerablemente, respecto a los resonadores habituales como el SRR. Se mostrarán técnicas para poder reducir el tamaño de un resonador. Se caracterizarán los diferentes resonadores propuestos y se demostrará que son válidos para la síntesis de dispositivos de microondas basados en líneas de transmisión zurdas.

2.1 Resonadores planares

En esta sección estudiaremos los resonadores que se pueden modelar como un tanque resonante LC en paralelo y que pueden ser excitados mediante un campo eléctrico o magnético que esté debidamente polarizado. Entre estos resonadores se encuentra el *split ring resonator* (SRR), que fue la primera estructura resonante que supuso un punto de inflexión para el comienzo de la investigación sobre los metamateriales. Se hablará del *complementary split ring resonator* (CSRR), que es la estructura complementaria al SRR. También se mostrarán resonadores que son modificaciones sobre la topología de una de estas dos estructuras, de forma que cada resonador tenga unas características que lo haga idóneo para un determinado dispositivo.

2.1.1 *Split Ring Resonator* (SRR)

Como se ha comentado en la introducción, fue John Pendry quien a finales de los años 90 introdujo el resonador de anillos abiertos (*Split Ring Resonator*, SRR) [35]. El SRR fue la primera estructura no magnética capaz de dar lugar a valores negativos de permeabilidad magnética efectiva (μ_{eff}) y supuso el comienzo de la investigación en el campo de los metamateriales. En la figura 2.1(a) se muestra su estructura. Está formada por dos anillos metálicos concéntricos en los cuales se han practicado unas aberturas en posiciones opuestas una respecto de la otra. Si sobre esta estructura se aplica un campo magnético variable perpendicular a la superficie de los anillos (paralelo al eje z en la figura 2.1(a)) se inducen corrientes en los anillos. Estas corrientes son capaces de pasar de un anillo a otro gracias a la capacidad distribuida que se forma entre ellos, de forma que se cierra el circuito [36]. El SRR se puede modelar como un tanque resonante LC , donde L modela el comportamiento inductivo introducido por las tiras conductoras y C modela la capacidad distribuida entre los anillos. Las propiedades electromagnéticas de los SRRs han sido estudiadas en diversos trabajos [37–39]. Dichos análisis muestran que los SRRs se comportan como un resonador que puede ser excitado mediante un campo magnético variable axial, paralelo al eje z en la figura 2.1 (a), o por un campo eléctrico variable polarizado en el sentido del eje y [40].

En la figura 2.1(b) se muestra el circuito equivalente para el SRR. Su frecuencia de resonancia puede ser calculada mediante los elementos del tanque $L_S C_S$ que modelan el SRR. La inductancia L_S puede ser aproximada por la que tendría un anillo con un radio r_0 de valor medio entre los dos anillos que forman el SRR y cuya anchura, c , fuera la de dichos anillos [40]. Cuando se desprecia la capacidad que aparece entre los extremos de las aperturas de los anillos, C_0 representa la capacidad total entre anillos y C_S corresponde a la conexión serie de las capacidades de las mitades superior e inferior del SRR (figura 2.1). La capacidad aproximada (C_0) está calculada teniendo en cuenta su distribución a lo largo del perímetro de los anillos, a partir de la capacidad estática por unidad de longitud C_{pul} entre dos tiras metálicas paralelas de ancho c y distancia d [37]. Este análisis para el SRR es aproximado; en [40] se presenta un método numérico con mayor precisión, que puede ser implementado mediante un programa informático para calcular L_S , C_S y la frecuencia de resonancia del SRR, f_0 , a partir de sus dimensiones físicas [41] (ver expresiones 2.1, 2.2 y 2.3). En las siguientes ecuaciones están representadas las capacidades y la frecuencia de resonancia del SRR.

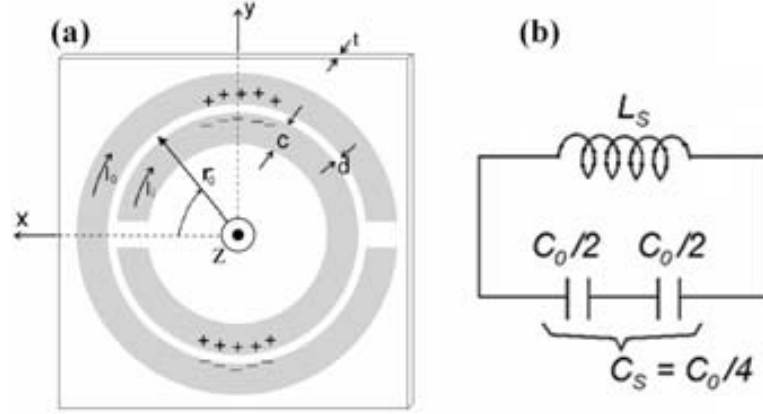


Figura 2.1: Esquema de las dimensiones más relevantes del SRR (a), modelo eléctrico (b). En (a), c es el ancho de las tiras metálicas, d separación entre tiras, r_0 radio medio de ambas tiras metálicas, t grosor de sustrato sobre el que se implementa el resonador, I_o e I_i indican el sentido de las corrientes de ambas tiras de metal y los signos positivos y negativos indican la distribución de cargas en la primera resonancia del anillo. En (b), L_S representa la inductancia del modelo eléctrico del SRR, C_0 representa la capacidad total entre tiras y C_S la conexión serie de las capacidades de las mitades superior e inferior del SRR.

$$C_0 = 2\pi r C_{pul} \quad (2.1)$$

$$C_S = \frac{\pi r C_{pul}}{2} \quad (2.2)$$

$$f_0 = \frac{1}{2\pi\sqrt{L_S C_S}} \quad (2.3)$$

Cuando el SRR es excitado presenta varias frecuencias de resonancia [38]. En este trabajo, nos centraremos en la primera de estas frecuencias, para la que el resonador es eléctricamente más pequeño. La longitud de onda guiada (λ_g) a esta primera frecuencia de resonancia es mayor que el diámetro del resonador. Esta cualidad hace que el SRR sea un resonador eléctricamente pequeño e idóneo para emplearlo en el diseño de medios efectivos. El hecho de que el SRR presente una longitud por debajo de λ_g en su primera frecuencia de resonancia, es debido a que si la distancia entre tiras (d) es suficientemente

2.1. RESONADORES PLANARES

pequeña, la capacidad (C_S) tomará valores elevados que proporcionarán una frecuencia de resonancia baja (ecuación 2.3). Esta frecuencia está por debajo de las frecuencias dinámicas de las partículas que forman el resonador, de forma que, aumentando esa capacidad (reducir la distancia d), se reduce el tamaño eléctrico del resonador [38]. En la figura 2.2 podemos ver el *layout* de una estructura típica utilizada para la excitación de un SRR, donde se ha utilizado una línea *microstrip* para excitarlo (la topología es cuadrada para facilitar el acoplamiento entre la línea y el resonador). En la gráfica de la figura 2.2 podemos observar como la frecuencia de resonancia del SRR (f_0) está por debajo de las resonancias de los anillos individuales que forman el resonador (f_{0_E} y f_{0_I}). Esta frecuencia de resonancia, f_0 , inferior a la resonancia del anillo externo, f_{0_E} , es debida al acoplamiento que hay entre ambos anillos. Debido a este acoplamiento, la segunda frecuencia de resonancia del SRR, f_1 , está por encima de la resonancia del anillo interno, f_{0_I} [38]. A un mayor acoplamiento, le corresponde una mayor separación entre resonancias y, por tanto, una menor longitud eléctrica del resonador a la primera resonancia. Las posibilidades de reducción de la longitud eléctrica del SRR vienen limitadas principalmente por la tecnología utilizada en la fabricación, y la longitud eléctrica del SRR, determinada experimentalmente, es del orden de $\lambda/10$.

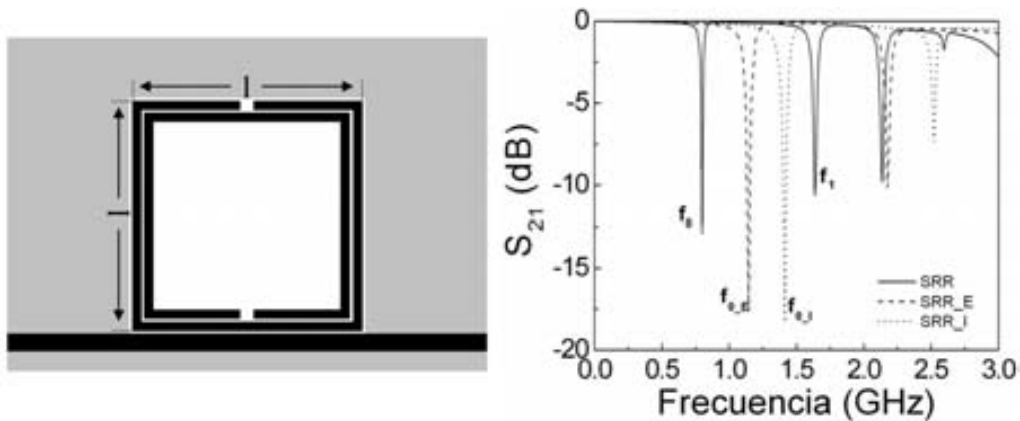


Figura 2.2: *Layout de un SRR excitado mediante una línea microstrip. El color gris corresponde al plano de masa y el negro a las tiras de metal. En la gráfica se muestra la respuesta en frecuencia del coeficientes de transmisión de la estructura (línea continua). Las líneas discontinuas corresponden a las frecuencias dinámicas de las partículas, simulando cada una de las tiras de metal por separado. f_0 corresponde a la primera frecuencia de resonancia del SRR, f_{0-E} corresponde a la primera frecuencia de resonancia del anillo exterior y f_{0-I} corresponde a la primera frecuencia de resonancia del anillo interior.*

En el artículo A se ha caracterizado un SRR, además de otros resonadores, acoplados a una línea coplanar. El método para la caracterización de resonadores acoplados a una línea coplanar se explicará en el apartado 3.2.3. En la tabla 1 del artículo B podemos ver los resultados obtenidos en la caracterización del SRR (con tecnología microstrip), donde ha sido analizado con diferentes estructuras y diferentes grosores de substrato para determinar su frecuencia de resonancia, longitud eléctrica a su frecuencia de resonancia, el factor de reducción al acoplar los dos anillos, el coeficiente de acoplamiento que existen entre los anillos y el factor de calidad descargado. En la referencia [54] y en el artículo B se detalla cada uno de los parámetros que han sido analizados para la caracterización de estos resonadores, donde además se presentan otros resonadores que veremos en la sección 2.2.

2.1.2 Complementary Split Ring Resonator (CSRR)

El *complementary split ring resonator* (CSRR) fue propuesto por Falcone *et al.* en el año 2004 [18]. El resonador consiste en dos anillos abiertos y concéntricos, igual que en el SRR, pero en este caso los anillos no son metálicos, sino que las tiras se retiran de un plano de metal, quedando libre de metal la zona del resonador. En base al principio de Babinet, puede decirse que al excitar el CSRR mediante un campo eléctrico axial al eje z o por un campo magnético paralelo al eje y (ver figura 2.3), este resonador tiene un comportamiento “complementario” al del SRR [52] (figura 2.1). En el CSRR, campos y corrientes eléctricas se intercambian por campos y corrientes magnéticas, respecto al SRR, presentando valores negativos de la permitividad dieléctrica efectiva (ε_{eff}). En la figura 2.3 se representa un CSRR en un plano de metal junto a su circuito equivalente. Podemos ver que el circuito equivalente es el dual circuital del SRR [55]. Al igual que se ha hecho en el SRR, se puede hacer una interpretación de los valores de la inductancia y capacidad del tanque resonante. En el CSRR, la capacidad C_C correspondería a un disco de radio $r_0 - c/2$ rodeado por un plano metálico a una distancia c de su borde, y la inductancia L_S sería la combinación paralela de las dos inductancias que conectan el disco interior y la región exterior al resonador [40].

Idealmente, el SRR y el CSRR son estructuras duales de forma que sus circuitos equivalentes están relacionados por las ecuaciones $C_C = 4(\varepsilon_0/\mu_0)L_S$ y $C_0 = 4(\varepsilon_0/\mu_0)L_0$ y sus frecuencias de resonancia coinciden [40]. En condiciones reales, donde el substrato y el metal no son ideales y, además son finitos, esta dualidad no es exacta, pero es una muy buena aproximación. De esta forma

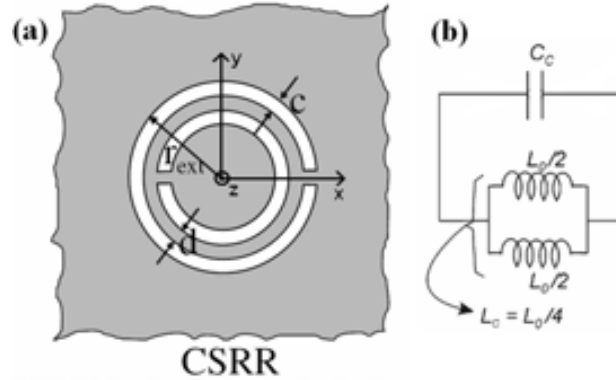


Figura 2.3: Layout de un CSRR (a) y su circuito equivalente (b). El color gris representa el metal y el blanco representa las zonas donde el metal ha sido retirado.

podemos decir que un SRR y un CSRR con las mismas dimensiones físicas darán aproximadamente la misma frecuencia de resonancia.

2.1.3 Resonadores con distintas topologías

A partir del SRR o del CSRR, modificando su topología, se pueden obtener otros resonadores que presenten características distintas, dependiendo de la utilidad y uso que se le quiera dar. En esta sección hablaremos solamente de resonadores metálicos, como el SRR, pero para todo resonador que se expone en este apartado existe su resonador complementario (dual), que se obtendría de la misma forma que el CSRR se obtiene respecto al SRR.

Como hemos dicho, introduciendo modificaciones en la topología de la figura 2.1(a) se pueden obtener nuevos resonadores como, por ejemplo, el *non-bianisotropic split ring resonator* (NB-SRR) que fue introducido por primera vez en el año 2003 por Baena *et al.* [56]. Su topología puede verse en la figura 2.4 junto al coeficiente de transmisión (S_{21}) obtenido al excitar el resonador mediante una línea microstrip [54]. En la gráfica puede verse el S_{21} para el resonador (línea continua) y para las partículas que forman el resonador (línea discontinua), simuladas por separado. En este resonador las partículas que se acoplan son iguales, al contrario que en el SRR, por este motivo sólo aparece una resonancia para las partículas, debido a que son iguales, y su frecuencia de resonancia también lo es. Al acoplar estas dos partículas obtenemos la reducción en frecuencia de la primera resonancia [38].

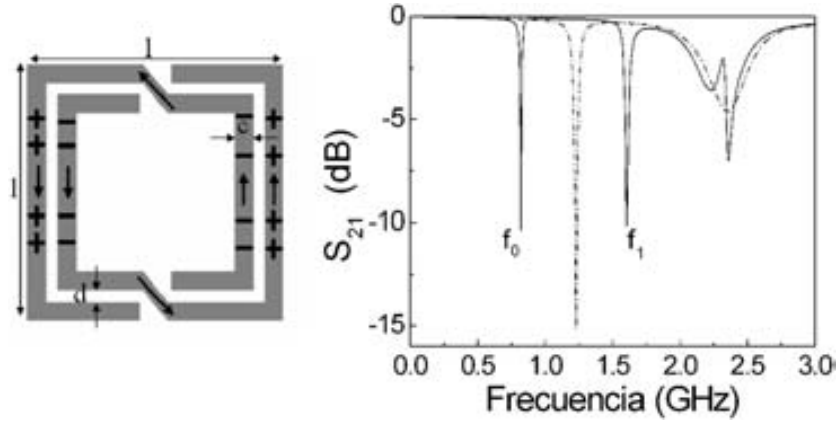


Figura 2.4: Layout de un NB-SRR. Las flechas indican el sentido de la corriente, y los signos la distribución de cargas en la primera frecuencia de resonancia. En la gráfica se muestra la respuesta en frecuencia del coeficiente de transmisión del resonador al ser excitado mediante una línea microstrip (línea continua). Las líneas discontinuas corresponden a la respuesta frecuencial de las partículas, simulando cada una de las tiras de metal por separado. f_0 y f_1 corresponden a la primera y segunda frecuencia de resonancia del NB-SRR.

Lo que caracteriza a este resonador es la distribución de cargas que se produce en su primera frecuencia de resonancia (ver figura 2.4). Debido a esta distribución de cargas se evita la bianisotropía del SRR, es decir, evita los efectos de polarización cruzada estudiados en [40]. Otra característica de este resonador es que muestra una simetría de rotación de 180° . La frecuencia de resonancia de un NB-SRR está muy cercana a la que tendría un SRR de iguales dimensiones. Su circuito equivalente es igual al del SRR y valores muy cercanos (iguales idealmente). Por tanto, el interés principal de este resonador (NB-SRR) reside en la eliminación de la bianisotropía propia del SRR [40].

Otra topología muy utilizada en circuitos de microondas planares es la espiral (*spiral resonator*, SR). Podemos ver su estructura en la figura 2.5(a) junto a su circuito equivalente [40]. En el SR, cabe destacar una reducción de la frecuencia de resonancia en un factor 2 respecto a la del SRR. Esta disminución en frecuencia es debido al aumento en un factor 4 en su capacidad C_S (ecuación 2.3) ya que sería igual a C_0 [42]. Otra topología utilizada para la síntesis de medios metamateriales es el *double split ring resonator* (DSRR) [56]. Podemos ver la topología del DSRR en la figura 2.5(b), junto a su circuito

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equivalente. En el caso del DSRR la frecuencia de resonancia se ve aumentada en un factor 2 respecto al SRR, ya que si nos fijamos en la ecuación 2.3 y en el circuito equivalente, al disminuir la capacidad en un factor 4 respecto al SRR, su frecuencia se ve aumentada en un factor 2.

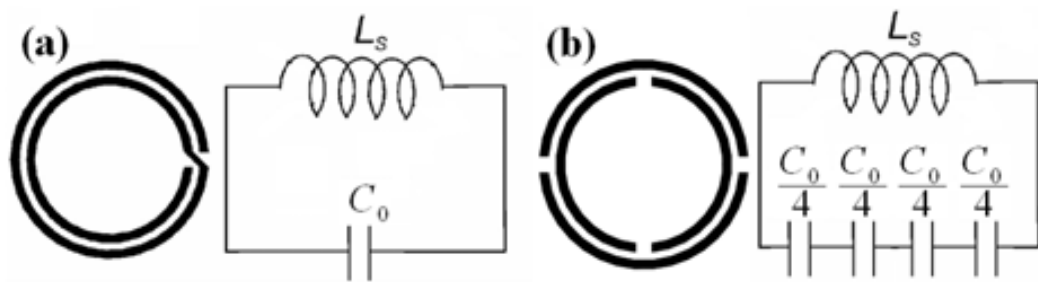


Figura 2.5: Layout de un SR (a) y un DSRR (b) junto a sus circuitos equivalentes.

Otras estructuras que se utilizan como resonadores utilizados en metamateriales es el *archimedean spiral resonator* (ASR), cuyo layout podemos ver en la figura 2.6(a) y que puede ser acoplado a otro ASR, como se puede observar en la figura 2.6(b) [57, 58].

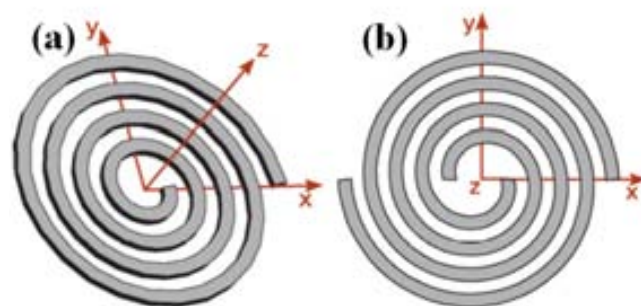


Figura 2.6: Layout para un *archimedean spiral resonator* simple (a) y doble (b). Figura extraída de [57].

2.1.4 Reducción de la longitud en resonadores planares

En el **artículo A** se proponen diferentes técnicas para la reducción del tamaño eléctrico de resonadores planares de una única capa de metal, como los expuestos en esta sección. Esta técnica consiste en incrementar el número de elementos que forman el resonador, es decir, para el caso del SRR consistiría en añadir anillos más pequeños en su interior, como puede verse en la figura 1(d) del artículo A. En el caso del SR, consistiría en aumentar el número de vueltas de la espiral, como puede verse en la figura 1(e) del artículo A. Esta técnica ha sido estudiada en [59] y [60], y un ejemplo de aplicación puede verse en [61]. Para comentar esta técnica utilizaremos como ejemplo al SRR de la figura 1(a) del artículo A. Como se puede observar, este SRR tiene una longitud eléctrica de $\lambda/13$, y al acoplarle tres anillos más en su interior, se obtiene una longitud eléctrica de $\lambda/15$. Los anillos que se añaden tienen que estar en la parte interior del resonador, para que su diámetro no aumente. El acoplar estos anillos implica reducir el área interior del resonador, lo que implica reducir la facilidad de excitación del mismo. Por otro lado, llega un momento en el que se mantiene la frecuencia de resonancia, aunque se sigan acoplando mas anillos [59], por este motivo esta técnica para reducir el tamaño eléctrico de un resonador está muy limitada. Otro ejemplo que se muestra en el artículo A, es el de la espiral de la figura 1(b), con una longitud eléctrica que pasa de $\lambda/26$ a $\lambda/55$ al pasar de 2 a 5 vueltas, resultando la espiral de la figura 1(e) del artículo A.

2.2 Resonadores planares bimetálicos

En la sección 2.1 se han presentado resonadores planares para los cuales se necesita una única capa de metal para implementar el resonador. En esta sección se presentarán resonadores compuestos por dos capas de metal. Esto hace posible que se pueda producir un acoplamiento a través del sustrato, al que llamaremos acoplamiento *broadside*. Además se proponen técnicas para la disminución de la longitud eléctrica de estos resonadores bimetálicos, cuya principal estrategia está en la introducción de vías a través del sustrato donde se imprime el resonador y que conectan los dos niveles de metal.

2.2.1 *Broadside Coupled Split Ring Resonator* (BC-SRR)

El *Broadside Coupled Split Ring Resonator* (BC-SRR) fue propuesto en [39] y su estructura puede verse en la figura 2.7. Este resonador, al igual que el SRR, puede ser excitado mediante un campo magnético perpendicular a la superficie del BC-SRR. Tanto el SRR como el BC-SRR presentan características muy similares, en términos de circuito equivalente. Para este caso, en lugar de poner un anillo en el interior del otro y en el mismo nivel de metal, en el BC-SRR las partículas se acoplan de forma *broadside* al situarlas una a cada lado del sustrato donde se imprime el resonador. Respecto al SRR, la C_S y la L_S del circuito equivalente del BC-SRR serán mayores que las del SRR con las mismas dimensiones [37], ya que la capacidad será debida al ancho de la tira metálica (c) y no al grosor del metal (t). La inductancia, al ser la correspondiente a una tira metálica de radio igual al de los anillos (r_0), también se ve aumentada ligeramente. El cálculo más preciso de L_S y C_S se hace mediante un método numérico similar al del SRR, incluido en el programa informático ya mencionado [41].

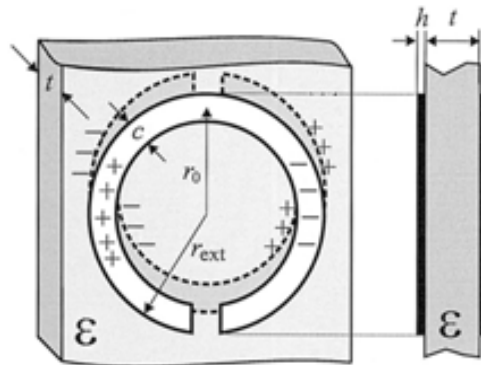


Figura 2.7: Esquema y dimensiones del BC-SRR. El bloque de color gris representa el sustrato sobre el que se imprime el resonador. c , r_0 , r_{ext} y h son el ancho, el radio medio, radio exterior y el grosor de las tiras metálicas, t es el grosor de sustrato con una permitividad dieléctrica ϵ sobre el que se implementa el resonador. Los signos positivos y negativos indican la distribución de cargas en la resonancia del anillo.

2.2.2 Reducción de la longitud eléctrica en resonadores planares bimetálicos

En el **artículo A**, además de proponer un método para la reducción de la longitud eléctrica de resonadores planares, expuesta en el apartado 2.1.4, se presenta otra técnica para la reducción del tamaño eléctrico, combinando el uso de dos niveles de metal con vías que los conecten. El introducir vías en el diseño de resonadores bimetálicos sirve para aumentar la longitud de las partículas que forman el resonador, sin necesidad de aumentar su área. Además, esto nos permite una gran flexibilidad a la hora de diseñar el resonador y favorecer el acoplamiento lateral o el acoplamiento *broadside*, a conveniencia de las necesidades en cada caso. En la figura 1(c) del artículo A se presenta un BC-SRR con una longitud eléctrica de $\lambda/22$. Al conectar ambas capas de metal con una vía, se puede obtener un *broadside spiral resonator* de dos vueltas (BC-SR2), como el de la figura 1(f) del artículo A, con un tamaño de $\lambda/43$. Este BS-SR2 presenta una reducción de un factor 2 en su frecuencia de resonancia respecto del BC-SRR, al igual que ocurre entre el SRR y el SR. Si además del uso de vías aplicamos la técnica de añadir más elementos (apartado 2.1.4) se pueden obtener resonadores como por ejemplo el *broadside spiral resonator* de cuatro vueltas (BC-SR4) de la figura 1(g) del artículo A, llegando a obtener un tamaño eléctrico de $\lambda/110$, siendo este un nivel de miniaturización muy alto y que aún puede ser incrementado. Para la determinación de la frecuencia de resonancia y la longitud eléctrica de estos resonadores se ha analizado una línea de transmisión CPW a la que se le acoplan los diferentes resonadores, como se muestra en la figura 2 del artículo A, y, analizando el coeficiente de transmisión utilizado el software comercial *Agilent Momentum*, donde el cero de transmisión coincide con la frecuencia de resonancia de los resonadores.

En el **artículo B** se presentan y caracterizan una selección de diferentes resonadores bimetálicos diseñados utilizando las técnicas propuestas. En la tabla 1 del artículo B se muestran las imágenes en 3D de los diferentes resonadores diseñados, todos ellos están constituidos por dos capas de metal unidas mediante vías, excepto el SRR y el BC-SRR que se han utilizado como resonadores de referencia. Para un análisis más profundo, se han caracterizado los resonadores para dos sustratos de grosores diferentes ($h = 1.27mm$ y $h = 0.254mm$) y permitividad de $\epsilon = 10.2$. Para la caracterización, cuyo resultado se incluye en la misma tabla, se ha utilizado tecnología microstrip. Todos los resonadores han sido diseñados con forma cuadrada, para facilitar el acoplamiento con la línea microstrip. Las dimensiones para todos los resonadores son las mismas, siendo su lado $l = 15.28mm$, las tiras de metal tienen un

2.2. RESONADORES PLANARES BIMETÁLICOS

grosor $c = 0.6mm$ y la separación entre tiras $d = 0.2mm$. El hecho de que los resonadores hayan sido diseñados con las mismas dimensiones, y que se hayan caracterizado para diferentes grosores de substrato, es para facilitar un análisis comparativo entre resonadores. Las flechas sobre los dibujos de los resonadores corresponden al lugar donde se encontrarían los máximos de corriente en la primera resonancia. Estos diagramas de corrientes pueden verse en la referencia [54] obtenida con el software *Agilent Momentum*. Para la caracterización se ha obtenido la frecuencia de resonancia acoplando el resonador a una línea de transmisión microstrip (ver figura 2.2), y tras el análisis del coeficiente de transmisión, obtenido mediante simulación electromagnética con el software *Agilent Momentum*, se obtiene la primera frecuencia de resonancia, f_0 , que coincide con el primer cero de transmisión. El siguiente punto analizado es el tamaño eléctrico del resonador, que es la relación que existe entre el lado del resonador, l , y la longitud de onda guiada a la frecuencia de resonancia del resonador, λ_g , obtenida mediante la siguiente ecuación

$$\lambda_g = \frac{c}{f_0 \sqrt{\varepsilon_{eff}}} \quad (2.4)$$

donde c es la velocidad de la luz, ε_{eff} la permitividad dieléctrica efectiva y f_0 la primera frecuencia de resonancia. El factor de reducción, R , que corresponde al porcentaje de reducción de la frecuencia de resonancia obtenido al acoplar las dos partículas que forman el resonador, respecto a la resonancia de una sola partícula. Para determinar de forma cuantitativa el grado de acoplamiento que hay entre las partículas que forman el resonador existen diferentes ecuaciones, dependiendo del tipo de acoplamiento que existe entre las partículas [62]. El tipo de acoplamiento que existe entre las partículas del resonador puede determinarse mediante un análisis de la distribución de corrientes [38, 63]. Independientemente del tipo de acoplamiento, y suponiendo que las frecuencias de resonancia de las partículas independientes son iguales o lo suficientemente cercanas, podemos reducirla a la ecuación (1) del artículo B, que corresponde al coeficiente de acoplamiento, M , entre las partículas. Para el cálculo de M solamente necesitamos conocer las dos primeras frecuencias de resonancia del resonador completo. Estos valores de R y M no pueden darse para los resonadores espirales, porque estos resonadores no están compuestos por varias partículas acopladas, sino por una sola acoplada consigo misma. Esto ocurre para los resonadores 3 y 4 de la tabla 1 del artículo B. Para terminar con la caracterización de los resonadores, se ha calculado el factor de calidad descargado, Q_U , que es una estimación de las pérdidas electromagnéticas de los elementos que forman el resonador. Q_U para un resonador se puede definir como la energía media almacenada entre la pérdida en el resonador por ciclo

[62]. Podemos obtener Q_U a partir de la ecuación 3 del artículo B, para la cual necesitamos la frecuencia de resonancia, el ancho de banda a $-3dB$ y el módulo del coeficiente de transmisión, para un circuito en configuración paso de banda, como el que se muestra en la figura 2.8.

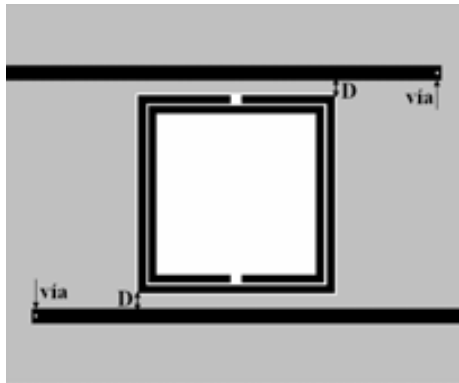


Figura 2.8: Layout del circuito en modo de transmisión, usado para calcular el factor de calidad de los resonadores.

Observando la tabla 1 del artículo B, podemos comprobar que en todos los resonadores donde existe acoplamiento *broadside* se observa una gran reducción de la longitud eléctrica al disminuir el grosor del sustrato. Los resonadores 3 y 4 están basados en el SR, pero aprovechando el acoplamiento *broadside*. Los resonadores 5 y 6 han sido diseñados para favorecer el acoplamiento *broadside*. Esto se puede observar en el valor de M , ya que aumenta al disminuir el grosor del sustrato. Esto mismo no ocurre con los resonadores 7 y 8, ya que, aunque disminuya la frecuencia de resonancia al disminuir el grosor del sustrato, se puede observar que el valor de M disminuye. Esto se debe a que en este caso el acoplamiento entre las partículas es lateral, pero en la frecuencia de resonancia influye de manera considerable el auto-acoplamiento *broadside* de cada una de las partículas que forman el resonador, como ocurre en los resonadores 3 y 4. Respecto al factor de calidad, vemos que en resonadores bimetálicos se ve reducido respecto al SRR, pero su disminución no es tan importante como lo es la disminución de su longitud eléctrica.

En el artículo B también se pone de manifiesto la viabilidad de la utilización de estos resonadores para la implementación de circuitos de microondas y metamateriales. Como ejemplo se usan algunos de los resonadores en

dos configuraciones distintas de líneas de transmisión metamaterial utilizando tecnología microstrip y CPW. Estas estructuras se estudiarán más adelante en el capítulo 3.

Para concluir, en este apartado se ha demostrado que es posible la reducción o miniaturización de resonadores, utilizando dos niveles de metal conectados mediante vías e incrementando el número de partículas acopladas. También ha quedado demostrada la posibilidad de utilizar estos resonadores para la implementación de dispositivos de microondas y metamateriales, como queda de manifiesto en los artículos A y B.

2.3 Resonadores abiertos

En esta sección hablaremos de los resonadores abiertos. Estos resonadores necesitan de dos terminales que estén directamente conectados al resonador para permitir su excitación mediante una fuente de tensión o de corriente. Se les llama abiertos porque el *Open Split Ring Resonator* (OSRR), que fue el primer resonador de este tipo, propuesto en 2004, es la versión *abierta* del SRR. Veremos como es este resonador y se presentará y analizará con detalle su versión complementaria, el *Open Complementary Split Ring Resonator* (OC-SRR), que se caracteriza por primera vez en esta tesis.

2.3.1 *Open Split Ring Resonator* (OSRR)

En 2004 se propone el *Open Split Ring Resonator* (OSRR) [64]. Este resonador es la versión abierta del SRR. Su topología, conectado a una línea microstrip, puede verse en la figura 2.9 junto a su circuito equivalente. Este resonador puede ser modelado por un tanque resonante LC serie. Su topología se puede describir como dos ganchos concéntricos situados uno en el interior del otro. La inductancia del tanque resonante, L_S , es igual a la que tendría un SRR de iguales dimensiones, y la capacidad, C_0 , viene dada por la capacidad distribuida entre los dos ganchos. Por tanto, podemos decir que la frecuencia de resonancia de un OSRR será la mitad que en un SRR de iguales dimensiones, como ocurre con el SR. Sin embargo, en este caso el tanque resonante está “abierto”, de ahí su nombre.

La facilidad de acoplar este resonador a una línea de transmisión y su re-

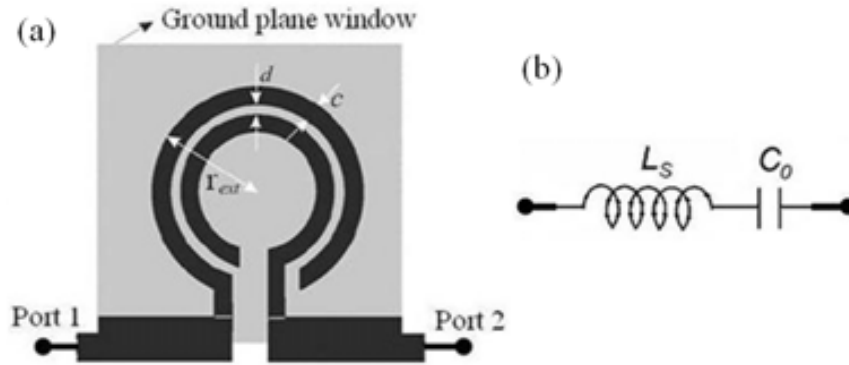


Figura 2.9: Open Split Ring Resonator (OSRR) excitado mediante una línea de transmisión microstrip (a). Circuito equivalente para el OCSR (b).

ducido tamaño, lo hace idóneo para la síntesis de dispositivos de microondas, como se hace en [64] y [65].

2.3.2 Open Complementary Split Ring Resonator (OCSR)

En esta sección hablaremos del *Open Complementary Split Ring Resonator* (OCSR) que, aunque ha sido utilizado en la síntesis de circuitos de microondas metamaterial en diferentes trabajos [66, 67], se caracteriza y analiza por primera vez en el **artículo C** de esta tesis. El OCSR viene a ser el resonador complementario del OSRR, al igual que el CSRR lo es del SRR, y cuya topología y circuito equivalente puede verse en la figura 2 del artículo C. Este resonador, al igual que para el OSRR, es necesario excitarlo de forma directa mediante dos terminales, indicados en la figura 2 del artículo C. Como se puede observar en la figura, los terminales necesarios para la excitación de la partícula pueden conectarse de dos formas posibles: una es a través de la tira de metal que hay entre las ranuras que forman el resonador y que correspondería a la inductancia L_0 de su circuito equivalente y la otra es a través de la capacidad de las ranuras modelada por C_C del tanque resonante que modela al resonador. Por lo tanto, es de esperar que por dualidad, la frecuencia de resonancia sea la misma que la del OSRR, que sería la mitad que la correspondiente al CSRR, como ocurre con el OSRR respecto al SRR. Ha de tenerse en cuenta que, al

igual que con el resto de resonadores presentados en este trabajo, este modelo circuital es válido solamente en un cierto rango frecuencial alrededor de su frecuencia de resonancia.

Para validar que la frecuencia de resonancia del OCSR es la mitad de la del CSR y que su circuito equivalente es un tanque resonante LC paralelo, hemos utilizado una estructura de test como la que se muestra en la figura 3(a) del artículo C. Esta estructura consiste en una línea CPW, con dos OCSR simétricos a cada lado de la línea central grabados en el plano de masa. Se han utilizado vías a través del substrato y tiras metálicas en la parte inferior, para conectar el plano de masa y evitar modos de propagación indeseados. En este caso, el circuito equivalente de la estructura es el que se presenta en la figura 3(b) del artículo C. Los dos tanques resonantes en derivación corresponden a los dos OCSSRs y las secciones de línea modelan la línea coplanar con una impedancia característica $Z_0 = 50\Omega$, que es la misma de los puertos utilizados. Con esta configuración se puede determinar la frecuencia de resonancia a partir de la respuesta frecuencial, donde se obtiene un cero de reflexión o transmisión total. La transmisión total se obtiene a la frecuencia de resonancia de los OCSSRs, es decir, la impedancia de los resonadores es infinita, de esta forma solo tendremos la línea de transmisión CPW adaptada a los puertos ($Z_0 = 50\Omega$). En la figura 4(a) del artículo C, se representa la respuesta frecuencial (coeficiente de transmisión, S_{21} , y reflexión, S_{11}), para la estructura de la figura 3(a) del artículo C, simulada con *Agilent Momentum* y medida con el analizador de redes vectorial *Agilent 8720ET*. También se muestra en esa figura la simulación circuital obtenida de la extracción de parámetros para el tanque resonante LC que modela los resonadores. Como puede verse, se obtiene un ajuste muy bueno en el rango de frecuencias donde se encuentra la resonancia de los OCSSRs. Para comparar, y ver que efectivamente la frecuencia de resonancia corresponde con la mitad de la de un CSR de iguales dimensiones, se ha simulado y medido un CSR de iguales dimensiones, grabado en el plano de masa debajo de una línea de transmisión microstrip, cuya respuesta frecuencial puede verse en la figura 4(b) del artículo C. En el caso del CSR, su frecuencia de resonancia ha sido obtenida mediante el método propuesto en la referencia [68]. Con estos resultados se demuestra que, efectivamente, el OCSSR tiene una frecuencia de resonancia que es la mitad de la del CSR con iguales dimensiones.

Una vez comprobado que la frecuencia de resonancia del OCSR es la mitad de la de un CSR. Hay que comprobar que al igual que ocurre entre el SRR y el OSRR, la reducción en frecuencia es debida al aumento de la ca-

pacidad, que es cuatro veces mayor en el OSRR que en el SRR, mientras que la inductancia es prácticamente la misma. En sus versiones complementarias, hay que comprobar que el OCSR y CSRR tiene una inductancia cuatro veces mayor e igual capacidad que el CSRR. Para hacer esta comprobación, se ha realizado un estudio con diferentes dimensiones para el OCSR y CSRR, se han extraído sus parámetros que podemos ver reflejados en las tablas 1 y 2 del artículo C, quedando demostrado y validado el modelo circuital propuesto para el OCSR.

Para un estudio más amplio del resonador se ha analizado la respuesta frecuencial de la estructura para frecuencias de hasta cinco veces la primera frecuencia de resonancia del OCSR. Como se puede observar en la figura 5 del artículo C, después de la primera frecuencia de resonancia (f_0), hay otras dos frecuencias donde se observa un estrecho pico (f_1) en el coeficiente de transmisión seguido de un cero de transmisión (f_Z). Como se demuestra en el artículo C, el pico que se obtiene en el S_{21} a la frecuencia f_1 , se produce por un modo de propagación indeseado en la línea de acceso al resonador, que puede ser eliminado mediante tiras de metal y vías que conectan el plano de masa de ambos lados de la línea de acceso que conecta el resonador con la línea de transmisión CPW. La supresión de este pico puede verse en las figuras 10 y 11 del artículo C, donde, a mayor ancho de la tira de metal, mayor eliminación del pico que aparece a la frecuencia f_1 . El cero de transmisión a la frecuencia f_Z viene determinado por la inductancia que introduce el trozo de línea que accede al OCSR, como se demuestra en el análisis de las respuestas frecuenciales obtenidas tras la simulación de varias estructuras en las que se han mantenido las dimensiones de los resonadores y variando las de la línea de acceso. Como se puede ver en la figura 7 del artículo C, al variar las dimensiones del trozo de línea sólo modificamos la frecuencia del cero de transmisión f_Z . Por este motivo, el circuito equivalente del resonador, teniendo en cuenta la línea de acceso, puede modelarse como el de la figura 6(a) del artículo C, cuya respuesta puede verse en la figura 5 del artículo C. El circuito equivalente de la figura 6(b) del artículo C modela hasta la segunda banda de transmisión, como puede verse en la figura 5 del artículo C.

Otro paso en la caracterización del OCSR, ha sido el estudio del diagrama de corrientes magnéticas, obtenido mediante *Agilent Momentum*, que puede verse en la figura 8 del artículo C. Se observa que, para la primera frecuencia de resonancia, las corrientes magnéticas son codireccionales, lo que corrobora la dualidad con el OSRR [64], como ocurre entre el SRR y el CSRR [52]. El máximo de corriente magnética, para la primera resonancia se produce en la región que hay entre los anillos y la línea central de la CPW, lo que

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coincide con el máximo de corriente magnético de un *complementary spiral resonator* (CSR), que correspondería a la zona donde se unen el anillo interior y exterior.

Este nuevo resonador se ha caracterizado por primera vez en esta tesis, y como se verá en el capítulo 4 es idóneo para su uso en dispositivos de microondas. Además se verán diferentes configuraciones posibles para el uso de este resonador en el diseño de dispositivos, tanto en tecnología microstrip como en tecnología CPW.

Capítulo 3

Líneas de transmisión metamaterial

En la introducción de esta tesis se ha hablado de los medios efectivos con características zurdas, a los que se les llamó medios zurdos. Estos medios aparecieron por primera vez en el año 2000 y fue en el año 2002 cuando se empezaron a aplicar estos conceptos para la síntesis de líneas de transmisión artificiales, pasando de medios con estructuras en tres dimensiones a tecnología plana. Lo que permitió pasar de medios en tres dimensiones a tecnología plana fue que los conceptos de operación son los mismos tanto en medios como en líneas de transmisión. En este capítulo se mostrarán las líneas de transmisión basadas en el modelo de línea dual y en el modelo resonante. Nos centraremos especialmente en el modelo resonante. Dicho modelo consiste en acoplar resonadores a las líneas de transmisión convencionales; será analizado en detalle y prestando especial interés sus modelos circuitales, teniendo en cuenta que los modelos circuitales de las líneas de transmisión metamaterial son válidos solo en el rango de frecuencias donde los elementos son eléctricamente pequeños.

3.1 Líneas de transmisión artificiales basadas en el modelo dual

El concepto de línea de transmisión artificial metamaterial fue introducido por primera vez en el año 2002 [48, 49, 69]. Como se comenta en el apartado 1.5, estas estructuras consisten en cargar líneas de transmisión convencionales con elementos reactivos, obteniéndose una ε_{eff} y μ_{eff} negativas en un cierto rango frecuencial. A estas líneas se denomina líneas de transmisión metamaterial, debido a que pueden controlarse sus propiedades eléctricas de forma

que puedan presentar un rango frecuencial con propagación zurda, lo cual no se puede obtener mediante líneas convencionales. Aunque estas líneas no son parte fundamental en el estudio de esta tesis, sí que han sido mencionadas en el **artículo G**, donde se hace un resumen de las líneas de transmisión artificiales más utilizadas. En la figura 1 del artículo G se muestran varios ejemplos de estas líneas artificiales en diferentes tecnologías, que consisten en cargar las líneas de transmisión convencionales con capacidades en serie e inductancias en derivación. Estos elementos de carga pueden ser elementos circuitales discretos (inductancias y capacidades SMD) o componentes semi-discretos (componentes planares eléctricamente pequeños) [51, 70]. El circuito equivalente general, sin pérdidas, para la celda elemental de este tipo de líneas está representado en la figura 2 del artículo G. En estas líneas podemos, dentro de unos límites, controlar el rango frecuencial donde se obtiene tanto transmisión zurda como diestra. De esta forma forma, cuando los elementos con los que se carga la línea, capacidades en serie (C_L en la figura 2 del artículo G) e inductancias en derivación (L_L en la figura 2 del artículo G), predominan sobre los elementos propios de la línea, inductancia serie (L_R en la figura 2 del artículo G) y capacidad en derivación (C_R en la figura 2 del artículo G), obtendremos propagación zurda, y cuando los elementos propios de la línea predominan sobre los de carga, se obtiene propagación diestra. De esta forma, se puede diseñar estas líneas. El hecho de presentar dos rangos frecuenciales de propagación, uno zurdo y otro diestro, hace que a este tipo de estructuras se las denomine líneas de transmisión compuestas zurdo-diestra (CRLH TL, del inglés *Composite Right/Left-Handed Transmission Line*) [71–73]. El nombre de modelo dual es debido a que el circuito equivalente de los elementos que dominan en la zona de transmisión zurda es el dual del circuito equivalente de una línea de transmisión convencional.

El diagrama de dispersión típico de estas estructuras CRLH está representado en la figura 3(a) del artículo G. En este diagrama de dispersión se pueden observar dos rangos frecuenciales donde se produce la propagación. Existe un rango frecuencial donde la fase es negativa ($-\pi < \beta l < 0$) que corresponde a las frecuencias donde $\varepsilon_{eff} < 0$ y $\mu_{eff} < 0$, es decir propagación zurda. Otro rango frecuencial, donde la fase es positiva ($0 < \beta l < \pi$) y que corresponde a $\varepsilon_{eff} > 0$ y $\mu_{eff} > 0$, obteniéndose propagación diestra. Se observa también que existe un rango frecuencial situado entre ambas bandas de transmisión, donde no es posible la propagación debido a que los valores de ε_{eff} y μ_{eff} son de signo contrario. En la figura 4(a) del artículo G se muestra el diagrama de dispersión para un caso particular, donde los dos límites de las bandas de transmisión coinciden, obteniéndose un rango continuo en el que es posible la

transmisión. Este caso se denomina caso balanceado, y da lugar a una sola banda de transmisión continua [71]. Este tipo de líneas balanceadas tienen un especial interés en aplicaciones de banda ancha [74, 75].

Además de poderse controlar los rangos frecuenciales de propagación, pueden controlarse también sus características eléctricas, es decir, la fase de la línea (βl), y la impedancia característica (Z_B) más allá de lo que se puede controlar en líneas convencionales. El comportamiento típico de Z_B para este tipo de líneas está representado en la figura 3(b) del artículo G para el caso no balanceado y en la figura 4(b) del artículo G para el caso balanceado. La posibilidad de poderse controlar estas características, con mayor flexibilidad, hace de este tipo de líneas muy interesantes en aplicaciones donde se requieran valores concretos de estos parámetros (ingeniería de dispersión y de impedancias). Esta controlabilidad que ofrecen las líneas artificiales ha llevado a su empleo en el diseño de dispositivos de microondas. Ejemplos de estos dispositivos basados en el modelo dual son acopladores [76–79], divisores de potencia [21, 22, 80], filtros [81] o antenas [82], etcetera.

3.2 Líneas de transmisión artificiales basadas en SRRs (modelo resonante)

Las líneas de transmisión artificiales basadas en SRRs consisten en acoplar dichos resonadores a una línea de transmisión convencional. En nuestro caso nos centraremos en tecnología microstrip y CPW. Al cargar una línea con estos resonadores obtenemos líneas que presentan permeabilidad magnética efectiva negativa ($\mu_{eff} < 0$) en un cierto rango frecuencial alrededor de la frecuencia de resonancia de los SRRs. En el caso de la tecnología CPW, los resonadores deben situarse dentro de las ranuras que separan la línea de los planos de masa (ver figura 3.1a) utilizando una única capa de metal, o bien situarlos al lado contrario del substrato donde se implementa la línea, alineados con la ranura de la línea CPW. En ambos casos el campo magnético que se genera en la línea, excita adecuadamente a los resonadores [83]. Ejemplos de estas dos estructuras podemos verlos en la figura 3.1. En tecnología microstrip se pueden obtener líneas con $\mu_{eff} < 0$ acoplando los SRRs a cada lado de la línea [84], como puede observarse en la figura 3.2.

Si estas líneas con $\mu_{eff} < 0$, además de ser cargadas con SRRs, se cargan con tiras metálicas en derivación en el caso de líneas CPW [17], y vías conec-

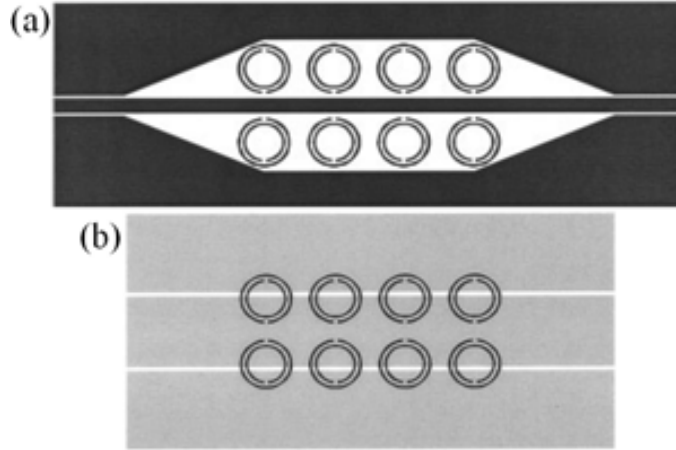


Figura 3.1: Líneas de transmisión CPW con $\mu_{eff} < 0$ en una sola capa de metal (a) y en dos capas de metal (b). Figuras extraídas de [83].

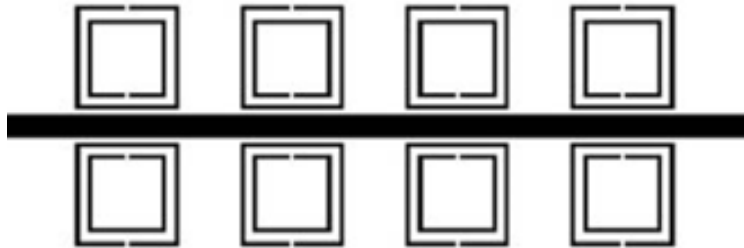


Figura 3.2: Líneas de transmisión microstrip cargada con SRRs para obtener $\mu_{eff} < 0$.

tadas a masa en las líneas microstrip [85] (ver figura 3.3) entonces podemos obtener transmisión zurda. En ambos casos los resonadores proporcionan la $\mu_{eff} < 0$, y la $\varepsilon_{eff} < 0$ es debida al efecto de las conexiones metálicas a masa. Debido a los grados de libertad que se tienen para el diseño de celdas de este tipo, se pueden obtener diferentes posibilidades en las características de propagación. Teniendo en cuenta la celda unidad, una condición necesaria, aunque no suficiente, para que la propagación sea posible es que se tiene que cumplir simultáneamente que la reactancia serie sea positiva ($\chi_S > 0$) y la reactancia en derivación sea negativa ($\chi_P < 0$), obteniéndose propagación diestra (*right handed*, RH). Para que la propagación sea zurda (*left handed*, LH) se tiene que cumplir que $\chi_S < 0$ y $\chi_P > 0$ simultáneamente. Otra condición es que la impedancia característica (Z_B) sea real, y la fase (βl) sea real y que tome

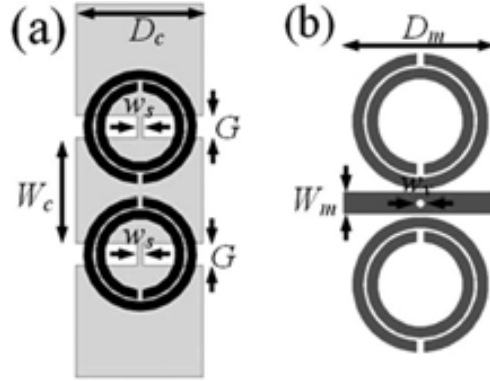


Figura 3.3: Líneas de transmisión artificiales con características de propagación zurdas en CPW (a) y microstrip (b).

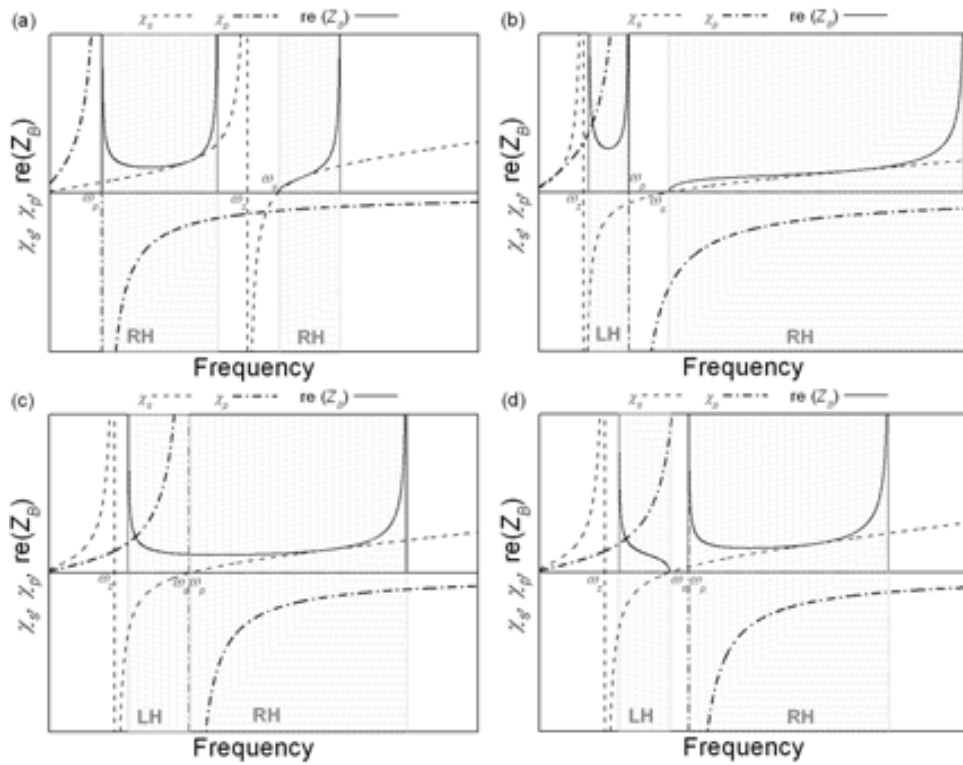


Figura 3.4: Representación de la reactancia en serie (χ_S) y en derivación (χ_P) y la parte real de la impedancia característica ($re(Z_B)$) para una celda unidad como las representadas en la figura 3.3. Se han considerado los casos donde $\omega_P < \omega_Z < \omega_S$ (a), $\omega_Z < \omega_P < \omega_S$ (b), $\omega_Z < \omega_P = \omega_S$ (c) (celda balanceada) y $\omega_Z < \omega_S < \omega_P$ (d).

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valores positivos para la transmisión RH ($0^\circ < \beta l < 180^\circ$) y negativos para la transmisión LH ($-180^\circ < \beta l < 0^\circ$). Teniendo en cuenta los diferentes valores de las frecuencias donde $\chi_S \rightarrow \infty$ (ω_Z), $\chi_S = 0$ (ω_S) y $\chi_P \rightarrow \infty$ (ω_P) se obtienen diferentes combinaciones de bandas de propagación. Estas combinaciones se muestran en la figura 3.4, donde se han representado χ_S , χ_P y la parte real de la impedancia característica ($re(Z_B)$) para las combinaciones donde $\omega_P < \omega_Z < \omega_S$, $\omega_Z < \omega_P < \omega_S$, $\omega_Z < \omega_P = \omega_S$ (celda balanceada [2, 5]) y $\omega_Z < \omega_S < \omega_P$. En la figura 3.5 se ha representado la fase (βl) para cada uno de estos casos. Las gráficas de las figuras 3.4 y 3.5 se han obtenido a partir de la simulación con el software comercial *Advanced Design System* de *Agilent Technologies*.

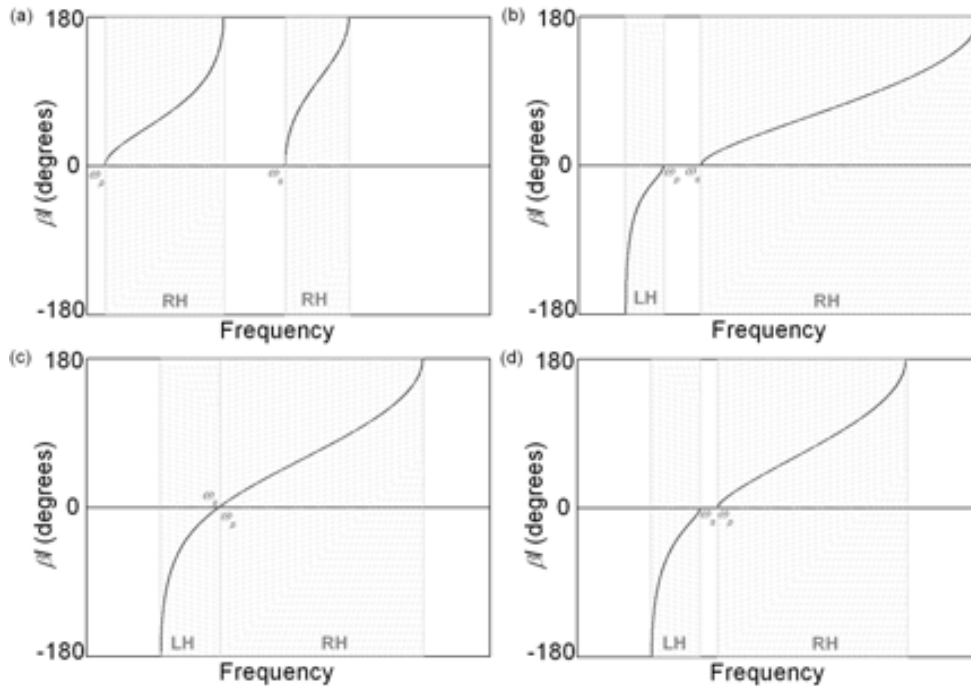


Figura 3.5: Representación de la fase (βl) para los diferentes casos representados en la figura 3.4.

En las representaciones gráficas de la figura 3.4 es fácil identificar los rangos frecuenciales donde es posible la propagación, (donde Z_B toma valores reales). Para poder identificar si dicho rango corresponde a una banda de transmisión zurda o diestra, es necesario atender al signo de χ_S y χ_P . En aquellas bandas donde χ_S sea negativa y χ_P positiva, la propagación será zurda, y viceversa.

Asimismo, pueden identificarse fácilmente los rangos de propagación zurda o diestra mediante los diagramas de dispersión representados en la figura 3.5. En aquellas bandas donde la fase (βl) toma valores negativos, la propagación es zurda, mientras que donde ésta es diestra la fase toma valores positivos. Puede verse cómo las diferentes combinaciones en las posiciones de las distintas frecuencias notables dan lugar a diversas combinaciones de bandas zurdas y/o diestras en las que la fase y la impedancia característica evolucionan de diferente forma.

En CPW además se pueden utilizar ambos lados de la línea donde se pueden acoplar resonadores de diferentes dimensiones sin aumentar el área de la estructura y obtener de esta manera varias bandas de transmisión zurda [86]. Que este tipo de líneas esté basado en SRRs, no implica que sea el único resonador que se pueda utilizar con esta configuración; otros resonadores, como los presentados en esta tesis y que pueden ser utilizados para la reducción del tamaño de la línea, pueden ser empleados como se demuestra con los resultados presentados y recogidos en esta tesis. Estas estructuras pueden ser utilizadas para diversas aplicaciones como filtros [17, 87–89], elementos radiantes como, por ejemplo, antenas [90] y filtros sintonizables [91], entre otros.

3.2.1 Modelo circuital

Cuando Martín *et al.* en 2003 [17], proponen la línea artificial CPW basada en SRRs, también proponen su modelo circuital. Este modelo circuital puede verse en varios de los artículos que forman parte de esta tesis (figura 1 del artículo E y la figura 6 del artículo G). Debido a la simetría de la línea, se utiliza el concepto de pared magnética, de forma que el modelo presentado corresponde solamente a la mitad de la celda unidad. En la figura 1(a) y 1(b) del artículo E tenemos este modelo circuital con y sin inductancia en derivación. En dicho modelo, C y L corresponden a la capacidad e inductancia de la línea CPW, C_S y L_S representan al tanque resonante que modela un SRR, el acoplamiento entre la línea y el SRR es modelado por M y la inductancia en derivación viene modelada por L_P , que corresponde al paralelo de las inductancias de las tiras de metal en derivación. Este modelo circuital es igualmente válido para la línea artificial microstrip basada en SRRs [85]. Al igual que en el caso CPW, C y L representan a la capacidad e inductancia de la línea microstrip, C_S y L_S corresponde al tanque resonante que modela un SRR, M modela el acoplamiento entre la línea y el SRR y L_P modela la inductancia en derivación dada por la vía conectada a masa. Este circuito puede

transformarse en el circuito de la figura 1(c) y 1(d) del artículo E para el caso con y sin inductancia en derivación, respectivamente. Esta transformación circuital viene determinada por las ecuaciones que aparecen en la misma figura 1 del artículo E, y que se propone por la comodidad de trabajar con un circuito más sencillo. Tanto para la línea en tecnología CPW como en microstrip, el modelo circuital es validado en los trabajos [17] y [85], respectivamente, donde se compara la respuesta circuital con la respuesta electromagnética.

3.2.2 Revisión del modelo circuital

En el apartado anterior se ha estudiado el modelo circuital de una línea artificial basada en SRRs. Este modelo ha sido validado en los trabajos donde se propone [17, 85]. En este apartado haremos una revisión del modelo circuital y se propondrá un nuevo modelo más preciso que el anterior. Empezaremos con un análisis del modelo circuital simplificado, donde el cero de transmisión que aparece en su respuesta frecuencial viene dado a la frecuencia de resonancia de los SRRs. Dicha frecuencia coincide con la frecuencia en la que la impedancia de la rama serie se hace infinita ($Z_S(\omega_Z) = \infty$). Como se puede observar en la figura 1 del **artículo E**, esta frecuencia coincide para ambas estructuras, con y sin inductancia en derivación (L_P). Sin embargo, al hacer un estudio sobre una sola celda unidad y al cambiar L_P o eliminar el *strip* conductor a masa, se observa un desplazamiento en el cero de transmisión (f_Z). Este desplazamiento puede verse en la figura 2(c) del artículo E, donde está representado el coeficiente de transmisión (S_{21}) y la fase (βl) para una misma estructura, con y sin la inductancia en derivación. Este desplazamiento en f_Z hace pensar que el modelo propuesto en [17] no es muy preciso, y que el desplazamiento no fue identificado debido a que se ve enmascarado al utilizar una línea de 4 celdas, donde este desplazamiento no es fácil de identificar. Por este motivo, en el artículo E se propone el modelo circuital que está representado en la figura 3(a) del mismo artículo. Dicho modelo se diferencia del anterior en que no divide la inductancia en derivación (L_P) colocándola a ambos lados de la inductancia de la línea (L), en este nuevo modelo es L la inductancia dividida y L_P queda en el centro del circuito. En este modelo no se aplica el concepto de pared magnética, de forma que el circuito modela la celda completa, además de asemejarse mas a la física real de la estructura.

El nuevo modelo también puede transformarse en un circuito en π , como muestra la figura 3(b) del artículo E, que es igual al modelo anterior transformado propuesto en [17], aunque con ecuaciones de transformación diferentes.

Esto confirma que, aunque el modelo anterior no explicara el comportamiento al cambiar L_P , el modelo era igualmente válido. Las ecuaciones de esta transformación son la 6 y 7 del artículo E. A estas ecuaciones se ha llegado tras analizar el circuito de la figura 3(a) del artículo E, reduciéndolo al modelo en π de un circuito bipuerta, que al ser simétrico su matriz de admitancias satisface que $Y_{12} = Y_{21}$ y $Y_{11} = Y_{22}$, donde $Z_S(\omega) = -(Y_{21})^{-1}$ y $Z_P(\omega) = (Y_{11} + Y_{21})^{-1}$ [92]. La impedancia serie ($Z_S(\omega)$) y paralelo ($Z_P(\omega)$), en función de sus elementos, corresponden a las ecuaciones 4a y 4b del artículo E. Como puede observarse en la figura 2(c) del artículo E, se aprecia un buen ajuste entre la simulación electromagnética y circuital, tanto para S_{21} como βl . Además, en las ecuaciones obtenidas también se observa los fenómenos físicos apreciados en las simulaciones, explicando perfectamente tanto el desplazamiento de f_Z al tener o no tener L_P . También se observa que a la frecuencia ω_S (expresión 8 del artículo E) donde $\beta l = 0^\circ$ y $Z_S(\omega_S) = 0$, es independiente de L_P , como se puede apreciar también en la figura 2 del artículo E.

En conclusión, el nuevo modelo para una línea metamaterial basada en SRRs, propuesto en esta tesis, es mucho más preciso que el anterior y explica toda la fenomenología física que ocurre en la estructura. En el artículo F se muestra como este nuevo modelo circuital también es válido para una línea metamaterial basada en SRRs en tecnología microstrip.

3.2.3 Extracción de parámetros

En el artículo E se propone el nuevo modelo circuital para la celda unidad de una línea de transmisión metamaterial basada en SRRs. Por otro lado, en el **artículo F** se propone un método de extracción de los parámetros de este modelo, a partir de la respuesta frecuencial obtenida tras la simulación electromagnética de la celda. Para ello, extraeremos los parámetros del circuito transformado (figura 2 del artículo F) que consta de cinco elementos, por lo que necesitamos cinco condiciones que nos permitan obtener los valores de dichos elementos. Haciendo la suposición de que la capacidad de la línea es prácticamente la misma con y sin la inductancia en derivación, a partir del coeficiente de reflexión (S_{11}), obtenemos una condición de la celda unidad sin inductancia en derivación. Dicha condición es la frecuencia donde S_{11} corta con el círculo de conductancia unidad en la carta de Smith, a la cual podemos obtener el valor de la susceptancia de la rama en derivación, que coincide con el valor de C (normalizado a los puertos) debido a que a esa frecuencia $Z_S(\omega_S) = 0$. Las otras cuatro condiciones las obtendremos a partir de la simu-

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lación con la celda completa, es decir, incluida la inductancia en derivación. Al igual que en la condición anterior, la frecuencia donde $Z_S(\omega_S) = 0$ es la frecuencia donde el S_{11} corta con el círculo de conductancia unidad, obteniendo el valor de la susceptancia de la rama en derivación, de donde podemos obtener el valor de L_P mediante la ecuación 10 del artículo F, ya que conocemos C de la condición anterior ($Z_S(\omega_S) = 0$), cuya expresión es la 9 del artículo F. Otra condición la obtenemos a la frecuencia de resonancia de la rama serie, donde $Z_S(\omega_Z) = \infty$ y viene dada por la expresión 11 del artículo F, fácilmente identificable en el coeficiente de transmisión (S_{21}), coincidiendo con la frecuencia del cero de transmisión. La última condición la podemos obtener a la frecuencia donde la fase es $\beta l(f_{\pi/2}) = -90^\circ$. A dicha frecuencia se cumple que la impedancia de la rama serie es igual a la impedancia de la rama en derivación cambiada de signo, como se expresa en la ecuación 13 del artículo F a partir de la expresión 12 del artículo F. En resumen, con las ecuaciones 9, 10, 11 y 13 del artículo F, junto a la condición obtenida de la simulación de la celda sin inductancia, obtenemos los cinco elementos del circuito. Un ejemplo de identificación de las frecuencias nombradas puede verse en la figura 4 del artículo F. Este método de extracción también es válido para la estructura sin inductancia en derivación, es decir, eliminando el elemento L_P del circuito de la figura 2 del artículo F, que consiste en eliminar la vía a masa de la estructura microstrip (figura 1(c) del artículo F) o las tiras metálicas en derivación de la celda en CPW (figura 1(b) del artículo F).

Con el fin de validar este método de extracción se han simulado las celdas de la figura 1 del artículo F, con y sin las conexiones a masa. Los valores obtenidos tras la extracción están representados en la tabla I del artículo F. Las respuestas frecuenciales, electromagnéticas y circuitales están representadas en las siguientes figuras del artículo F: en la figura 4(b) y 6(a) están representadas las respuestas para la estructura microstrip con y sin la vía a masa, y en la figura 5 y 6(b) están las respuestas para la estructura en tecnología CPW con y sin tiras conductoras en derivación. En todos los casos se observa un ajuste muy bueno, tanto en respuesta frecuencial como en el diagrama de dispersión, entre la simulación electromagnética y la circuital obtenida con los parámetros extraídos, lo que demuestra la validez del método propuesto.

3.2.4 Caracterización

Una vez propuesto y comprobado el método de extracción de parámetros, donde se obtienen los valores de los cinco elementos que forman parte del

circuito equivalente simplificado (figura 2 del artículo F) surge la posibilidad de obtener los parámetros físicos reales del circuito sin transformar (figura 3 del artículo F). Pasar de un circuito a otro es trivial invirtiendo las expresiones de la 5 a la 8 del artículo F. Para el caso en el que no se tiene la inductancia en derivación, L_P , se utilizan las mismas ecuaciones, de la 5 a la 8 del artículo F, haciendo $L_P \rightarrow \infty$, obteniendo las ecuaciones de la 1 a la 3 del artículo F. La solución a la inversión de las ecuaciones, teniendo en cuenta L_P , son:

$$L = \frac{2L'_P(L'_S + L')}{2L'_P + L'_S + L'} \quad (3.1)$$

$$L_P = \frac{(L'_P)^2}{2L'_P + L'_S + L'} \quad (3.2)$$

$$L_S = \frac{M^2(4(L'_P)^2 + 2L'_PL'_S + 4L'_PL' + L'_SL' + (L'_P)^2)}{2L'_SL'_P} \quad (3.3)$$

$$C_S = \frac{2C'_S(L'_P)^2(L'_S)^2}{M^2(2L'_P + L'_S + L')^2} \quad (3.4)$$

y para el circuito sin L_P ($L_P \rightarrow \infty$) las ecuaciones de transformación son:

$$L = L' + L'_S \quad (3.5)$$

$$L_S = \frac{2M^2}{L'_S} \quad (3.6)$$

$$C_S = \frac{C'_S(L'_S)^2}{2M^2} \quad (3.7)$$

Como se puede observar, para esta transformación necesitamos un elemento que no se obtiene tras la extracción de parámetros. Este elemento es el acoplamiento entre la línea y el resonador M . El valor de este parámetro es directamente proporcional al flujo de líneas de campo magnético generadas en la línea de transmisión y que atraviesan el interior del resonador. En [17] se hace una estimación de este parámetro, que viene dado por la siguiente expresión:

$$M = f \cdot L \quad (3.8)$$

donde L es la inductancia de la línea y f es el área fraccional de los slots de la

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línea CPW, es decir, el área del slot que hay sobre el resonador dividida por el área total del slot. El valor de este parámetro es necesario para obtener el valor de los elementos que modelan al resonador con la presencia de la línea CPW (C_S y L_S). Si obtenemos un valor exacto de M , podemos obtener los valores reales de los resonadores acoplados a la línea o incluso el valor del resonador aislado, sin la presencia de la línea CPW (C_I y L_I). Este punto se discute en el apartado III del artículo F. La aproximación que se hace en la ecuación 3.8 para el valor de M es válida bajo las condiciones siguientes: cuando el resonador es eléctricamente pequeño y todas las líneas de campo magnético atraviesan el resonador. Se ha hecho un estudio con diferentes resonadores acoplados a una línea CPW con inductancias en derivación (figura 7 del artículo F), variando los grosores de substrato y aplicando la extracción de parámetros propuesta. Para poder tener un punto de referencia se ha utilizado el método propuesto en [40], obteniendo los parámetros de los resonadores aislados (C_I y L_I) mediante el programa informático [41]. Una vez conocidos los valores C_I y L_I , podemos asumir que el valor de la capacidad del resonador (C_S) no se ve influenciada por la línea CPW, siempre y cuando el grosor de substrato sea lo suficientemente grande. Para la inductancia del resonador (L_S), sí que se ve afectada por la presencia de la línea, de forma que la inductancia del resonador aislado (L_I) la podemos relacionar con L_S , mediante la expresión 14 del artículo F, donde M' es el acoplamiento mutuo entre el resonador y su imagen. El valor de M' se puede obtener a partir de la formula de Neumann, mediante el método de las imágenes, aproximando la línea CPW a un plano de metal, considerando nulo el efecto de las ranuras de la línea, y considerando el resonador como un anillo delgado de metal cuyo radio es el radio medio de los anillos que forman el resonador. En la figura 8 del artículo F, se han representado los resultados del estudio donde se representan los valores para el área fraccional normalizada en función de diferentes grosores de substrato y varios resonadores. En todos los casos, se ha representado el valor de M_t , que corresponde a la inductancia mutua para la que $C_S = C_I$, entre el valor de M obtenido a partir de la ecuación 3.8, es decir, se ha representado $M_t/M = f_t/f$ en función del grosor del substrato. Analizando los resultados obtenidos en el estudio (figura 8 del artículo F) se tiene que al aumentar el grosor de substrato, como es de esperar, disminuye el flujo magnético que atraviesa el resonador. Como vemos en la gráfica, donde $M_t/M = f_t/f \rightarrow 1$, es para substratos delgados y no podemos decir que se cumple $C_I = C_S$, ya que la capacidad del resonador se ve influenciada por el plano de metal de la línea CPW. La conclusión de este resultado es que no tenemos un grosor de substrato donde M_t coincida con M simultáneamente y $C_I = C_S$. Por este motivo, no podemos calcular con precisión los valores de C_I y L_I de forma directa. Sin embargo, utilizando este método con un grosor de

substrato donde se cumpla que $C_I \approx C_S$, y obteniendo el valor de M_t como se ha hecho para la figura 8 del artículo F, podemos obtener los valores de C_S y L_S , que junto a la ecuación 14 del artículo F, teniendo en cuenta que $C_I \approx C_S$ podemos obtener los valores del resonador aislado. Para confirmar la validez de este método, en las tablas III y IV del artículo F están representados los valores obtenidos aplicando este método para diferentes casos. En la tabla III del artículo F, se representan los parámetros para líneas microstrip y CPW, con y sin inductancias en derivación, cuyo resultados han sido obtenidos a partir de los valores de la tabla I del mismo artículo. En la tabla IV del artículo F se representan los valores del circuito para una línea CPW, sin inductancias en derivación, para diferentes resonadores acoplados y dos grosores de substrato distintos.

En conclusión, en este apartado se ha corroborado que el método de extracción de parámetros es correcto y los valores obtenidos tras la extracción son coherentes con los valores obtenidos analíticamente. Con esto, estamos en disposición de decir que este método es válido para la caracterización de líneas artificiales basadas en SRRs. Este método de extracción es de especial interés para la caracterización de resonadores cuya topología es compleja y no es fácil su modelización, como los resonadores presentados en el artículo B. En el artículo A se utiliza la extracción de parámetros para la caracterización de una línea coplanar a la que se le acoplan diferentes resonadores. En el artículo F también se ha aplicado el método de extracción para una línea microstrip cargada con SRRs con y sin vía conectada a masa, y también se ha aplicado este mismo método para una línea CPW cargada con SRRs, con y sin tiras conductoras en derivación. También en el artículo F se utilizan varios resonadores de los presentados en esta tesis acoplados a una línea CPW con tiras metálicas en derivación.

3.2.5 Relación entre el ancho de banda y las dimensiones del resonador

A lo largo de esta tesis se ha observado que a menor longitud eléctrica, menor es el ancho de banda de un resonador. En el **artículo H** se ha hecho un estudio a partir del circuito equivalente de una línea artificial basada en SRRs, que relaciona el tamaño eléctrico del resonador y el ancho de banda. Para este estudio se ha utilizado el circuito simplificado (figura 2(b) del artículo H) sin la inductancia en derivación L_P . De esta forma, para aumentar el ancho de banda, hay que disminuir la pendiente de la susceptancia ($b = \omega'_0 C'_S$)

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del tanque resonante paralelo $L'_S C'_S$, por lo que hay que aumentar el valor de L'_S y disminuir el valor de C'_S . Esto se puede conseguir aumentando el acoplamiento entre la línea y el resonador M , aumentando la capacidad del resonador C_S y disminuyendo la inductancia del resonador L_S , como se puede ver a partir de las ecuaciones 1 y 2 del artículo H. Como se puede observar en dichas ecuaciones, el elemento más relevante es el valor de M , debido a que la dependencia de C'_S y L'_S respecto a M es cuadrática. Como el valor de M depende del área del resonador, a menor longitud eléctrica en el resonador, menor valor de M y, por tanto, menor ancho de banda a su resonancia. En el artículo H se intenta contrarrestar el valor de M incrementando el valor de C_S y disminuyendo L_S . Para este estudio, se ha considerado un SRR, un SR y un BC-SRR, de forma que se ha incrementado C_S y disminuido L_S en el SR y el BC-SRR con el fin de conseguir un ancho de banda mayor que en el SRR. Los resultados obtenidos se representan en la figura 4 del artículo H, tanto para la estructura con y sin inductancias en derivación, donde se observa que, aun teniendo valores extremos de C_S y L_S , no se consigue obtener un ancho de banda mayor. De este hecho puede concluirse que, aunque sobre el modelo circuital sea posible obtener respuestas más anchas, sobre layout no es posible implementar resonadores que sean eléctricamente más pequeños y a su vez obtengamos un ancho de banda mayor en líneas de transmisión metamaterial en cualquiera de sus configuraciones (con y sin inductancias en derivación).

3.3 Líneas de transmisión artificiales basadas en CSRRs (modelo resonante)

Una de las posibilidades de utilizar los CSRRs para la síntesis de líneas de transmisión artificiales es grabándolos en el plano de masa que está bajo una línea de transmisión microstrip, como se muestra en la figura 3.6, donde podemos ver la foto de la primera línea artificial implementada con CSRRs. Esta configuración fue propuesta por primera vez en 2004 por Falcone *et al.* [18], quienes, poco más tarde, en ese mismo año, propusieron una línea artificial basada en CSRRs con características de transmisión zurdas [52] (figura 3.7). En la primera estructura, sin cortes capacitivos sobre la línea microstrip, se obtiene una configuración de rechazo de banda, ya que $\varepsilon_{eff} < 0$ debido a la presencia de los CSRRs. En la estructura con cortes capacitivos se obtiene una configuración paso de banda en el rango frecuencial donde $\varepsilon_{eff} < 0$, por los CSRRs, y $\mu_{eff} < 0$ debido a los cortes en la línea. Las respuestas de ambas estructuras pueden verse en las figuras 3.6 y 3.7 respectivamente.

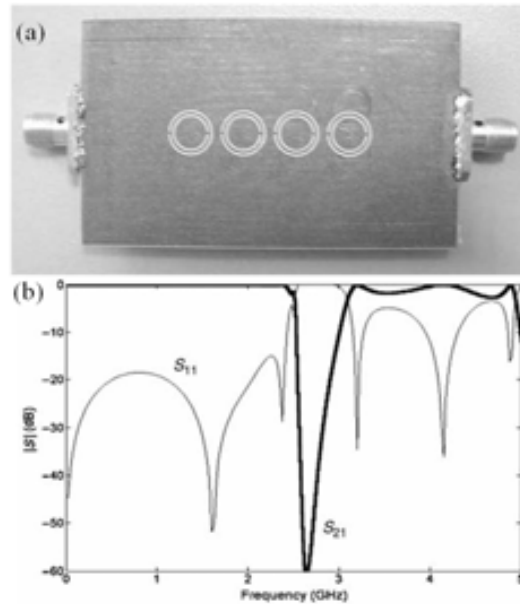


Figura 3.6: Plano de masa de una línea microstrip donde se han grabado CSRRs (a) para obtener una respuesta rechazo de banda (b). Figura extraída de [18].

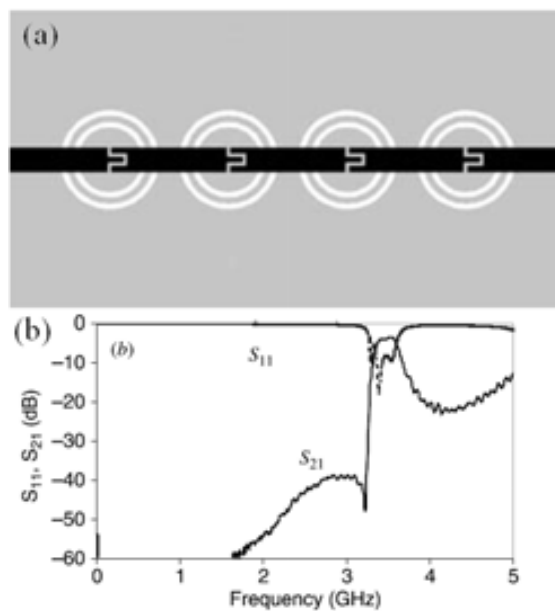


Figura 3.7: Layout de una línea de transmisión artificial basada en CSRRs (a) con características zurdas junto a su respuesta frecuencial (b). Figura extraída de [52].

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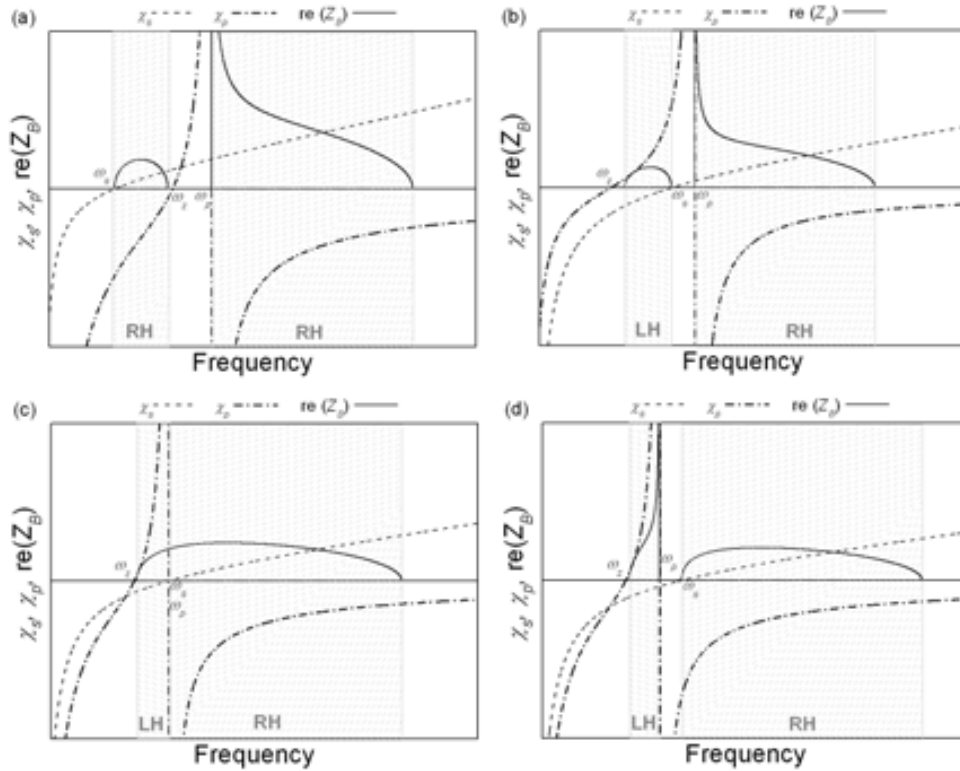


Figura 3.8: Representación de la reactancia en serie (χ_S) y en derivación (χ_P) y la parte real de la impedancia característica ($re(Z_B)$) para una celda unidad de la estructura representadas en la figura 3.7. Se han considerado los casos donde $\omega_S < \omega_Z < \omega_P$ (a), $\omega_Z < \omega_S < \omega_P$ (b), $\omega_Z < \omega_S = \omega_P$ (c) (celda balanceada) y $\omega_Z < \omega_P < \omega_S$ (d).

Este tipo de líneas ha sido analizado en detalle en diferentes trabajos [20, 93, 94]. Al igual que se ha hecho en la sección 3.2 para líneas basadas en SRRs, en la figura 3.8 se pueden observar las representaciones gráficas de las reactancias serie (χ_S) y derivación (χ_P) y la parte real de la impedancia característica ($re(Z_B)$), y en la figura 3.9 la fase (βl), para diferentes celdas unidad de líneas metamaterial basadas en CSRRs como la presentada en la figura 3.7. En este caso también se ha hecho para diferentes valores de las frecuencias: $\omega_S < \omega_Z < \omega_P$, $\omega_Z < \omega_S < \omega_P$, $\omega_Z < \omega_S = \omega_P$ (celda balanceada) y $\omega_Z < \omega_P < \omega_S$, siendo ω_S la frecuencia donde $\chi_S = 0$, ω_Z la frecuencia donde $\chi_P = 0$ y ω_P la frecuencia donde $\chi_P \rightarrow \infty$. En las representaciones gráficas de la figura 3.8 es fácil identificar los rangos frecuenciales donde es posible la propagación, (donde Z_B toma valores reales). Para saber si dicho

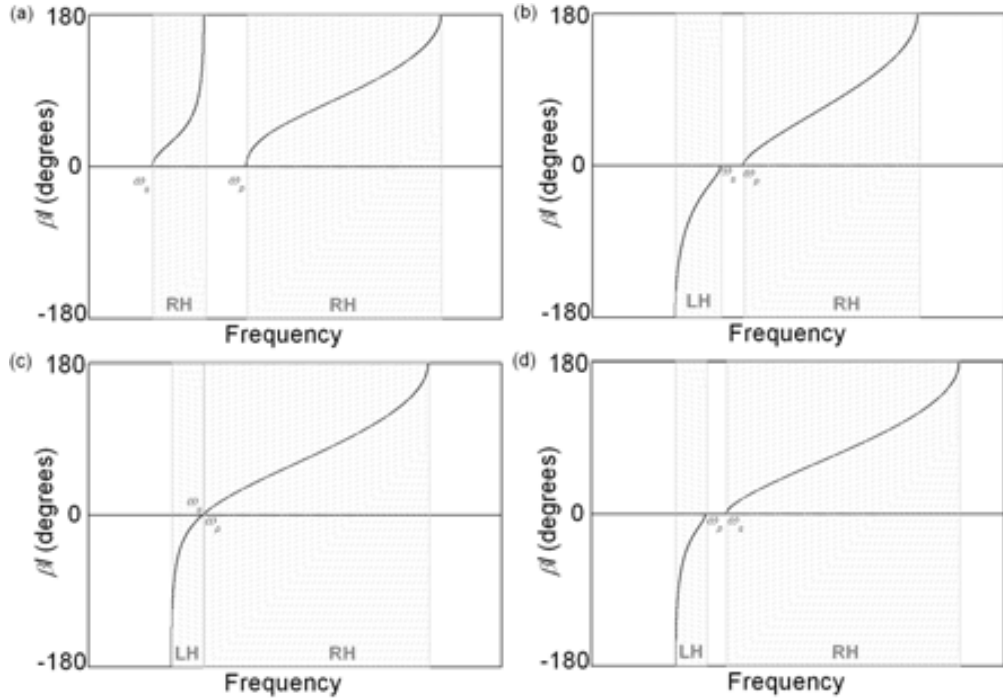


Figura 3.9: Representación de la fase (βl) para los diferentes casos representados en la figura 3.8.

rango corresponde a una banda de transmisión zurda o diestra, podemos fijarnos en el signo de χ_S y χ_P . En aquellos rangos frecuenciales donde $\chi_P < 0$ y $\chi_S > 0$, la propagación será diestra, y viceversa. Además, pueden identificarse fácilmente las bandas de transmisión zurda o diestra mediante los diagramas de dispersión representados en la figura 3.9. En aquellas zonas donde la fase (βl) toma valores positivos, la propagación es diestra, mientras que donde la fase es negativa, la propagación es zurda. Puede verse cómo las diferentes combinaciones en las posiciones de las distintas frecuencias notables dan lugar a diversas combinaciones de bandas zurdas y/o diestras en las que la fase y la impedancia característica evolucionan de diferente forma.

Al comparar los diagramas de dispersión de una celda basada en CSRRs (figura 3.9), con las obtenidas para una celda basada en SRRs (figura 3.8), se observa que se obtienen las mismas posibles combinaciones en lo que a bandas de transmisión se refiere. Sin embargo, al comparar las gráficas para la impedancia característica (figuras 3.4 y 3.8) puede observarse una evolución

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diferente en dicha impedancia para las bandas de transmisión que es debida al distinto comportamiento de las reactancias χ_S y χ_P en los dos tipos de celdas. Dado que χ_S y χ_P divergen y tienden a cero a distintas frecuencias, la impedancia característica tiene comportamientos diferentes en los límites de las bandas para una y otra celda.

En las secciones siguientes se estudia el circuito equivalente de este tipo de líneas de transmisión artificiales basadas en CSRRs. No obstante, aunque no forman parte del estudio de esta tesis, sí que nombraremos otras configuraciones de líneas basadas en CSRRs. Por ejemplo, los CSRRs pueden ser grabados en la propia línea, ya sea en tecnología microstrip [53, 95] o CPW [96]. También puede ser utilizado en estructuras más complejas, como en el modelo híbrido [97]. Este tipo de líneas pueden ser utilizadas para el diseño de diferentes dispositivos de microondas como filtros [97–102], acopladores [103], con celdas balanceadas para aplicaciones en banda ancha [74, 75], divisores [104], diplexores [20], componentes de banda dual [105] y de banda ancha [106–108], elementos radiantes [90], transductores magnetoinductivos [109], filtros sintonizables [91, 110, 111] y antenas [112, 113], entre otros.

3.3.1 Modelo circuital

El modelo circuital para una línea metamaterial basada en CSRRs ha sido propuesto y estudiado en diferentes trabajos [18, 40, 52, 68]. Este modelo circuital en T para una celda unidad de este tipo de línea puede verse en la figura 8 del artículo G. Con referencia al layout de la celda (figura 7 del artículo G) los elementos L_C y C_C corresponden al tanque resonante que modela el CSRR, L corresponde a la inductancia de la línea microstrip, C modela el acoplamiento eléctrico que hay entre la línea y el resonador y C_g modela la capacidad del corte realizado en la línea. Para el caso de línea sin el corte capacitivo, el modelo es el mismo pero eliminando C_g , es decir, cortocircuitando esa capacidad. Este modelo circuital es capaz de describir, dentro de un cierto rango frecuencial, el comportamiento electromagnético de la estructura, como puede verse en la figura 9(a) y 9(b) del artículo G, donde se observa un buen ajuste entre la simulación electromagnética y circuital.

3.3.2 Revisión del modelo circuital

En el apartado anterior se ha presentado, descrito y comprobado que el circuito propuesto para modelar la celda unidad de una línea de transmisión artificial metamaterial basada en CSRRs, es preciso y se recupera la respuesta obtenida por simulación o medida electromagnética. Sin embargo, recientemente se ha propuesto un nuevo modelo circuital para esta celda [114]. El nuevo modelo circuital puede verse en la figura 3 del artículo D. La razón que ha llevado a la modificación del modelo puede verse en la tabla I del artículo D, donde están los valores de los elementos del circuito inicial (figura 2(b) del artículo D) obtenidos para una misma celda en la que solamente se ha variado la distancia del corte capacitivo realizado en la línea. En los valores representados en la tabla, se observa que todos ellos varían al variar el valor de la distancia del corte, pero resulta llamativa la enorme diferencia que se observa en el valor de C , que hasta ahora modelaba el acoplamiento entre la línea y el resonador. Esta variación en C sorprende, ya que el hecho de practicar un corte en la línea, aunque puede aumentar la capacidad a masa, debido a las capacidades de borde, no lo hace en un grado tan grande. Sin embargo vemos que pasa, como puede apreciarse en la tabla I del artículo D. El nuevo modelo tiene en cuenta el modelo en π de la capacidad del corte en la línea, donde se tiene en cuenta las capacidades de borde. De esta forma, en el nuevo modelo (figura 3 del artículo D) L_C y C_C siguen modelando el tanque resonante del resonador, L sigue siendo la inductancia de la línea, y en el nuevo modelo, C_S modela la capacidad del corte en la línea, C_L corresponde a la capacidad de la línea y C_f a la capacidad de borde que aparece al practicar el corte en la línea. Este nuevo modelo puede ser transformado en el anterior mediante las expresiones 1 y 2 del artículo D. Teniendo en cuenta estas ecuaciones vemos que tanto C_g como C (modelo previo) dependen de C_S , C_L y C_f . Analizando la ecuación 2 del artículo D, tenemos que al aumentar la distancia del corte en la línea estamos disminuyendo el valor de C_S , lo que implica un aumento en el valor de C , que es exactamente lo que podemos observar en la tabla I del artículo D. En conclusión, el modelo propuesto originalmente es correcto, pero el nuevo modelo proporciona una mejor descripción física de la estructura y explica todos los efectos que se observan al variar cualquier elemento sobre el layout de la estructura [114].

3.3.3 Extracción de parámetros

En este apartado se recoge un método de extracción de parámetros (propuesto en [68]), para obtener los valores de los elementos del modelo circuital simplificado (figura 8 del artículo G), a partir de la respuesta frecuencial obtenida tras la simulación o la medida electromagnética de la celda unidad de una línea artificial basada en CSRRs [68]. Este método consiste en la identificación de diferentes frecuencias, dentro del rango de interés, necesitando cinco condiciones para determinar los cinco elementos. Las condiciones que usaremos para la extracción de parámetros será la frecuencia del cero de transmisión (f_Z) que viene dada por la expresión 10 del artículo G, y que puede obtenerse a partir del coeficiente de transmisión (S_{21}), la frecuencia de resonancia del CSRR cuya expresión es

$$f_0 = \frac{1}{2\pi\sqrt{C_C L_C}} \quad (3.9)$$

que es a la frecuencia en la que la rama en derivación se abre ($Z_P \rightarrow \infty$) y que puede ser identificada ya que es a la frecuencia en la que el coeficiente de reflexión (S_{11}) corta con el círculo de resistencia unidad. Otra condición la obtenemos a la frecuencia donde la fase es $\beta l = -90^\circ$ donde $Z_S = -Z_P$. En f_0 podemos obtener otra condición, ya que a esta frecuencia la admitancia en derivación es cero ($Y_P = 0$) y la impedancia que se ve desde los puertos coincide con la impedancia serie y la impedancia del puerto de salida, de forma que si la impedancia de los puertos es 50Ω tenemos que a f_0

$$Z_{in}(j\omega_0) = 50 + j\chi_S(\omega_0) \quad (3.10)$$

De esta expresión obtenemos dos condiciones, una tras la simulación de la celda sin el corte en la línea (donde la impedancia está formada por L) y otra con la celda con el corte (donde la impedancia está formada por L y C_g). Con estas cinco condiciones podemos obtener los valores de los cinco elementos del circuito. En la figura 9(a) y 9 (b) del artículo G se puede observar el buen ajuste en la respuesta frecuencial, donde se representa el S_{11} y S_{21} obtenidos tras la simulación electromagnética y circuital para una celda con y sin corte en la línea. Este método de extracción puede aplicarse, junto con alguna variante, a estructuras cuyo circuito equivalente sea similar a este [53, 95].

3.4 Dualidad entre líneas de transmisión artificiales basadas en SRR y CSRR

En la sección IV del artículo D se hace un pequeño análisis entre las similitudes de comportamiento que existen entre las líneas artificiales basadas en SRR y CSRRs, y en el **artículo I** se analiza con más detalle. En la figura 1 del artículo I se vuelven a presentar los layouts bien conocidos de las celdas unidad para líneas de transmisión artificiales basadas en SRRs y CSRRs. Para empezar viendo la dualidad de estas dos estructuras nos fijaremos en los layouts, donde para ver que una estructura es prácticamente la dual de la otra, empezaremos por la celda basada en SRRs. Si intercambiamos las zonas de metal y las zonas con ausencia de este, obtenemos dos celdas de línea basada en CSRRs, por lo que los dos layouts son duales entre sí. Aunque los layouts de las líneas son duales, no son estructuras electromagnéticas duales, sin embargo, los circuitos de ambas estructuras (figura 2(b) y 3(b)) son duales circuitales [55]. Esto quiere decir que la impedancia serie o derivación de uno de los circuitos, puede ser expresada como la admitancia en derivación o serie proporcional del otro circuito, es decir, una L en serie correspondería con una C en derivación. No sólo los layouts y los modelos circuitales son duales, sino que, como se puede observar en la figura 4 del artículo I, se obtiene el mismo comportamiento de ambas estructuras al variar el ancho de las tiras metálicas en derivación para el caso de la estructura en CPW, o variar el ancho del corte en la línea microstrip para la estructura basada en CSRRs. En ambos casos, al variar estas dimensiones, se obtiene un desplazamiento en el cero de transmisión. También podemos observar que con este parámetro podemos controlar el ancho de banda de la zona de propagación con características zurdas ($-180^\circ < \beta l < 0^\circ$). Además en ambos casos obtenemos el mismo comportamiento al aumentar este parámetro, es decir, al aumentar el ancho de las tiras metálicas en derivación o del gap en la línea microstrip el cero de transmisión se desplaza a frecuencias más bajas, al mismo tiempo que el ancho de banda de la transmisión zurda disminuye. También se cumple que la frecuencia a la que la fase de la línea es cero, $\beta l = 0^\circ$, es independiente del valor de este parámetro.

En conclusión, las líneas de transmisión artificiales basadas en SRRs y CSRRs son estructuras duales, tal y como se ha corroborado al estudiar sus layouts, sus modelos circuitales y sus respuestas frecuenciales.

3.5 Comparación entre el modelo resonante y el modelo dual de líneas artificiales

En el apartado anterior hemos visto la dualidad que existe entre las líneas de transmisión artificiales basadas en SRRs y en CSRRs. En este apartado compararemos el modelo resonante con el modelo dual para una línea de transmisión artificial metamaterial. En concreto nos fijaremos en una línea cargada con capacidades e inductancias y una línea artificial basada en CSRRs. Vamos a comparar estas dos estructuras debido a sus similitudes en el modelo circuital de ambas líneas, que podemos ver en las figuras 2 y 8 del artículo G. Como se puede ver, la única diferencia entre ambos circuitos es la capacidad de acoplamiento C que aparece en el modelo de la línea basada en CSRRs. Esta capacidad de acoplamiento hace que la línea presente un cero de transmisión a una determinada frecuencia, mientras que en el modelo dual, este cero de transmisión, está en el origen. El cero de transmisión, que se obtiene en el modelo resonante, permite obtener un corte abrupto por debajo de la banda zurda, que es interesante para el diseño de filtros. Prácticamente, este cero de transmisión es la única diferencia entre ambos modelos, ya que en ambos casos obtenemos un comportamiento zurdo-diestro [75]. Para el caso de las líneas de transmisión basadas en SRRs, si comparamos el circuito en T de la figura 2 del artículo G, para el modelo dual, con el circuito en π de la figura 12(b) del artículo G, para el modelo resonante basado en SRRs, el cero de transmisión que aparece antes de la banda de transmisión, con el modelo resonante, se debe a la inductancia que introduce el resonador (L'_S) y que no está presente en el circuito del modelo dual. Así pues, las líneas de transmisión basadas en SRRs presentan también un comportamiento parecido a las líneas CRLH basadas en líneas cargadas con inductores y capacidades.

Capítulo 4

Aplicaciones

En los capítulos anteriores se han propuesto y caracterizado resonadores que son eléctricamente pequeños y se ha demostrado que éstos resonadores son válidos para la síntesis de circuitos de microondas basados en metamateriales. También se han caracterizado diferentes líneas de transmisión que utilizan resonadores, de forma que podemos controlar sus propiedades eléctricas, que son la impedancia característica (Z_B) y la fase (βl), cuyo control es indispensable en el diseño de circuitos de microondas. En este capítulo veremos algunos ejemplos concretos en los que se pueden apreciar las posibilidades de estos resonadores y líneas de transmisión metamaterial, donde se han conseguido buenas prestaciones a la vez que una disminución del tamaño respecto a un dispositivo convencional.

4.1 Filtros *notch* (rechazo de banda)

Hemos visto que los resonadores como el SRR y, por tanto, todos los demás presentados en esta tesis, son capaces de inhibir la propagación de una señal en un determinado rango frecuencial. Esta capacidad se puede utilizar acoplando los resonadores a una línea de transmisión convencional, de forma que se puede utilizar para la eliminación de bandas espurias en filtros de microondas. Dicha eliminación se consigue sintonizando los resonadores, de forma que su frecuencia de resonancia esté alrededor de la frecuencia donde aparecen los espurios, con lo que éstos son eliminados. Esta técnica de eliminación de espurios se ha estudiado en diversos trabajos publicados [96, 115]. En la figura 4.1 podemos ver el layout de un filtro de líneas acopladas al que al añadirle diferentes SRRs, sintonizados a diferentes frecuencias, se ha conseguido mejorar la respuesta del filtro fuera de la banda de transmisión.

4.1. FILTROS NOTCH (RECHAZO DE BANDA)

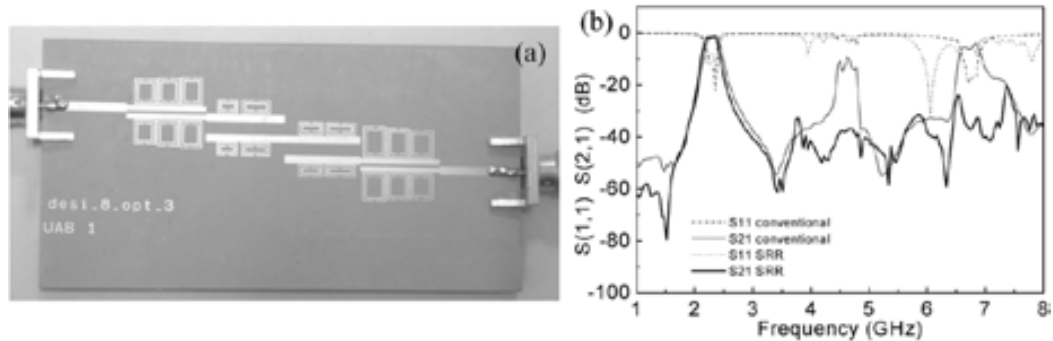


Figura 4.1: Layout (a) y respuesta frecuencial medida con y sin SRRs (b) de un filtro de líneas acopladas en tecnología microstrip. Figuras extraídas de [96].

Esta técnica puede emplearse también en tecnología coplanar. Un ejemplo de dicha aplicación puede verse en la figura 4.2 donde se han grabado CSRRs en las líneas acopladas de un filtro paso banda en tecnología CPW.

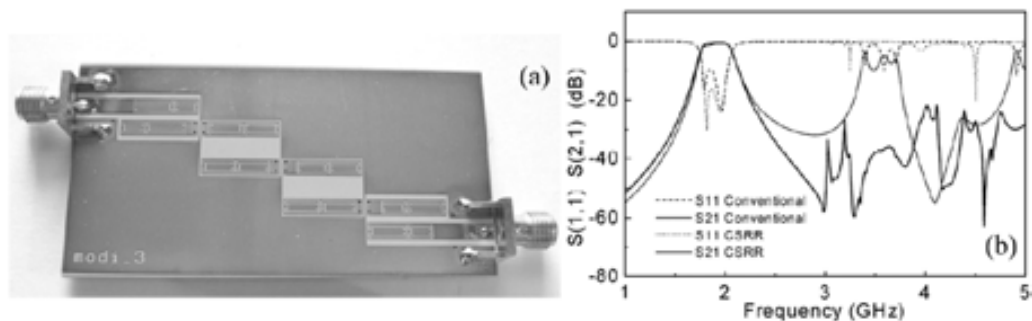


Figura 4.2: Layout (a) y respuesta frecuencial simulada con y sin CSRRs (b) de un filtro de líneas acopladas en tecnología CPW. Figuras extraídas de [96].

Tanto en el caso del filtro en tecnología microstrip (figura 4.1) como en el de tecnología CPW (figura 4.2), se ha conseguido eliminar las dos primeras bandas espurias sin necesidad de añadir otras etapas en cascada, que sería el proceso habitual y, por tanto, sin aumentar el área del dispositivo. Estas técnicas también pueden ser utilizadas con otros filtros como, por ejemplo, los de paso bajo de impedancia escalonada [98].

En el trabajo [96], además de utilizar los resonadores para la eliminación de bandas espurias de filtros paso bajo o paso banda, también se presentan filtros elimina banda. Este tipo de filtros son sencillos de implementar utilizando resonadores. Se trata de acoplar los resonadores a una línea de transmisión convencional y, de esta forma, a la frecuencia de resonancia donde se sintonizan los resonadores obtenemos rechazo en un cierto rango frecuencial que puede ser ensanchado con la adición de diferentes resonadores sintonizados a frecuencias ligeramente distintas. Ejemplos de estos tipos de filtros pueden verse en la figura 4.3, donde se implementan tanto en tecnología microstrip como en CPW.

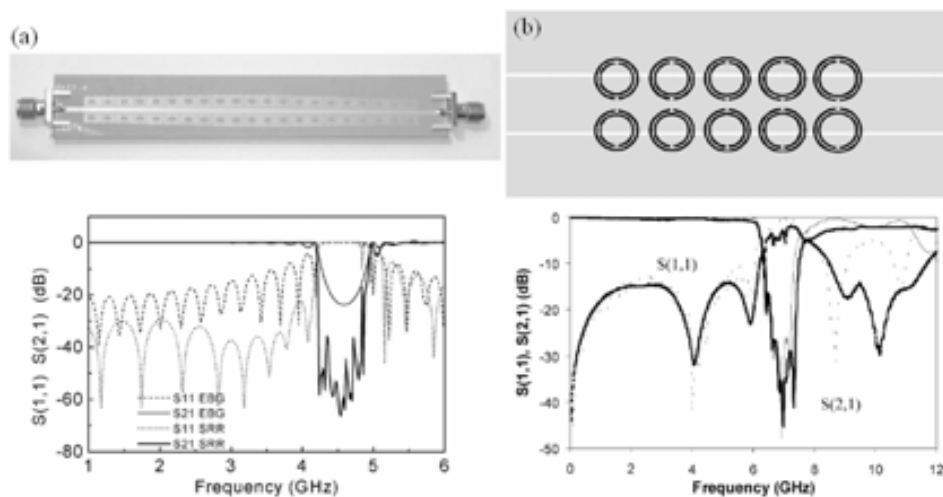


Figura 4.3: Layout de un filtro paso banda basado en SRRs en tecnología microstrip (a) y en tecnología CPW (b), junto a sus respuestas frecuenciales. Figuras extraídas de [96] (a) y [87] (b).

Como puede observarse en los ejemplos de filtros de rechazo de banda basados en resonadores acoplados a una línea de transmisión convencional, el tamaño de los resonadores determina el tamaño final de dicho filtro. Por este motivo, los resonadores propuestos en esta tesis, cuyo tamaño eléctrico es muy pequeño, se pueden utilizar para el diseño de filtros compactos, de forma que al disminuir el tamaño de los resonadores, se reduce el tamaño final del dispositivo. En el **artículo J** se han utilizado distintos resonadores presentados en esta tesis (resonadores 2, 3 y 4 de la tabla 1 del artículo B) para demostrar que estos resonadores son idóneos para conseguir disminuir el tamaño de este tipo de filtros. En la figura 1 del artículo J se presentan los layouts de los tres resonadores que se han utilizado para el diseño de este tipo de filtro. Los

4.1. FILTROS NOTCH (RECHAZO DE BANDA)

layouts de los filtros diseñados se presentan en la figura 2 del artículo J. Como puede observarse en el artículo J, se han diseñado tres filtros notch (rechazo de banda) cuyo tamaño es el mismo para los tres casos, pero utilizando diferentes resonadores. Los filtros están compuestos por tres etapas iguales. Su respuesta frecuencial se muestra en la figura 3 del artículo J, donde se observa que a pesar de ser de igual tamaño, al utilizar resonadores de menor longitud eléctrica, la frecuencia de rechazo del filtro es menor. De esta forma, a menor frecuencia de trabajo, con iguales dimensiones físicas del filtro, menor tamaño del filtro.

También se han fabricado estos filtros de forma que la frecuencia de operación de los resonadores sea aproximadamente la misma en los tres casos. Esto hace que los resonadores sean de distinto tamaño en cada caso y por tanto también los filtros diseñados, como se puede ver en la figura 4.4. Al igual que en el caso presentado en el artículo J, se ha utilizado un substrato de grosor $h = 0.254 \text{ mm}$ y permitividad dieléctrica $\varepsilon = 10.2$, de forma que para una frecuencia de trabajo alrededor de $f_0 = 0.3 \text{ GHz}$, la longitud lateral de los resonadores se ve reducida ($l = 23.02 \text{ mm}$ para el resonador 2, $l = 12.82 \text{ mm}$ para el resonador 3 y $l = 6.92 \text{ mm}$ para el resonador 4 de la tabla 1 del artículo B), como puede verse en la figura 4.4. Las gráficas de la figura 4.4 muestran los coeficientes de transmisión simulados (figura 4.4d) y medidos (figura 4.4e) correspondiente a los tres dispositivos. La línea continua corresponde a la línea con BC-SRRs (resonador 2 de la tabla 1 del artículo B), la discontinua corresponde al resonador 3 de la tabla 1 del artículo B y la línea punteada corresponde al resonador 4 de la tabla 1 del artículo B. En este caso el factor de reducción que se obtiene en los resonadores es el mismo que se obtiene en los filtros.

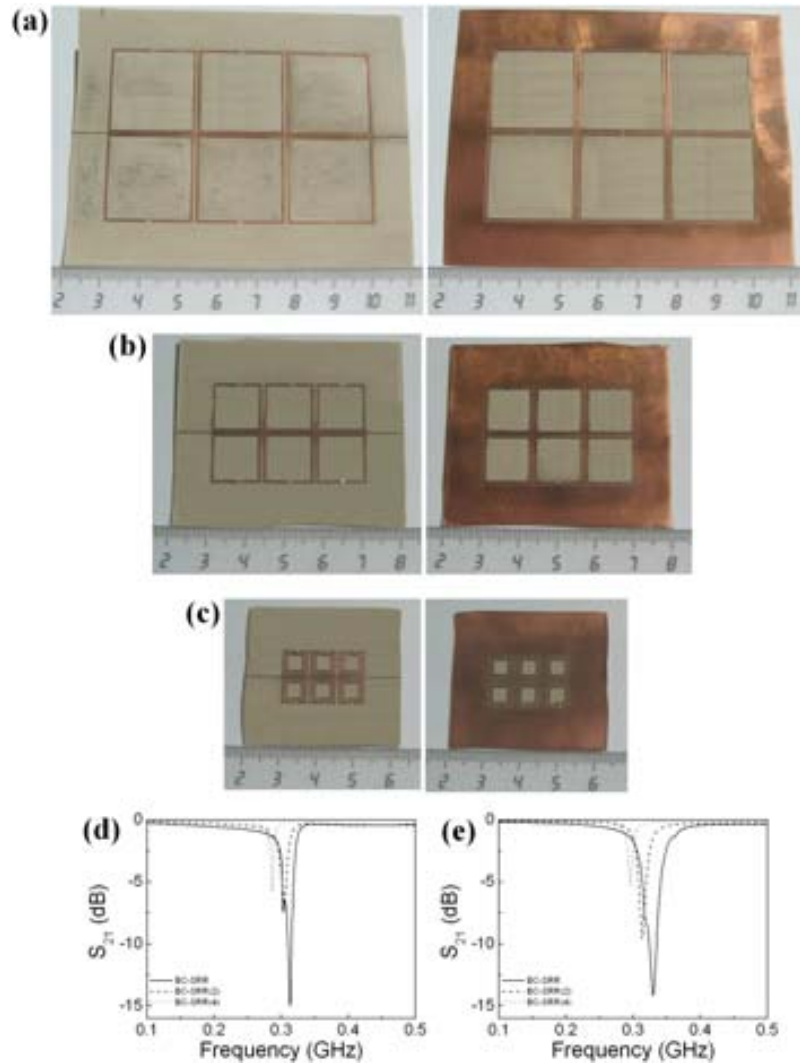


Figura 4.4: Fotografías de las diferentes líneas microstrip con resonadores acoplados que se han diseñado, fabricado y caracterizado. Su frecuencia de operación es cercana a los 0.3 GHz ; (a) con BC-SRRs de área $l \times l = (23.02 \times 23.02) \text{ mm}^2$, (b) con BC-SRs(2) de área $l \times l = (12.82 \times 12.82) \text{ mm}^2$ y (c) con BC-SRs(4) de área $l \times l = (6.92 \times 6.92) \text{ mm}^2$. Resultado de la simulación (d) y medida (e) del coeficiente de transmisión (S_{21}) de las tres líneas microstrip con resonadores acoplados que han sido diseñadas y fabricadas con una misma frecuencia de trabajo alrededor de 0.3 GHz , con BC-SRRs (línea continua), BC-SRs(2) (línea discontinua) y BC-SRs(4) (línea punteada).

4.2 Divisores de potencia

A lo largo de esta tesis se ha mencionado que una de las cualidades de los metamateriales es la posibilidad que tenemos de controlar las propiedades electromagnéticas de dicho material. Cuando hablamos de líneas de transmisión metamaterial, estas propiedades son la impedancia característica, Z_B , y la longitud eléctrica, βl . Además de poder controlar las características electromagnéticas, podemos obtener una reducción en tamaño e incluso obtener valores que con líneas convencionales es muy difícil obtener [116]. En [20] se demuestra la utilidad de líneas de transmisión metamaterial basadas en CSRRs en el diseño de dispositivos de microondas compactos. Esto se hace aprovechando la zona de transmisión con propiedades zurdas que aparece a bajas frecuencias [20]. En este apartado se propone la utilización de líneas de transmisión metamaterial basadas en SRRs para la miniaturización de divisores de potencia, presentados en el **artículo K**. Una de las topologías posibles para implementar estos dispositivos es la que se presenta en la figura 4 del artículo K. Para este dispositivo se necesita un inversor de impedancias, es decir, una línea cuya fase sea $\beta l = \pm 90^\circ$. La impedancia del inversor debe ser tal que la impedancia de entrada del dispositivo (Z_{in}), vista desde el puerto de entrada, sea igual a la de los puertos, que normalmente es $Z_0 = 50\Omega$. Para el esquema propuesto (figura 4 del artículo K) se necesita que el inversor de impedancias utilizado a la entrada del divisor de potencia presente una impedancia característica de 35.35Ω , siendo sus dos salidas líneas de 50Ω . La celda que se ha utilizado para la entrada es la representada en la figura 1 del artículo K. La respuesta frecuencial de esta celda y las características eléctricas se representan en la figura 3 del artículo K. Para este caso concreto, la frecuencia de operación ha sido fijada a $1.5GHz$, donde se obtiene $\beta l = -90^\circ$ y $Z_B = 35.35\Omega$. Una vez obtenida la celda que cumple las características necesarias se pasa al diseño final del dispositivo, que puede verse en la figura 5 del artículo K. Dicho dispositivo puede ser comparado con las dimensiones de un dispositivo convencional con las mismas características, obteniéndose un 40% de reducción en el tamaño de la celda diseñada. La respuesta frecuencial medida para el dispositivo diseñado y fabricado se presenta en la figura 6 del artículo K. Como puede observarse, se ha obtenido una buena respuesta en cuanto a las pérdidas de inserción, estando cercanas al valor ideal ($-3dB$). En la respuesta frecuencial medida se observa un pequeño desplazamiento en frecuencia, que se ha confirmado que es debido al margen de error en la permitividad dieléctrica del substrato. Para probar que utilizando resonadores eléctricamente pequeños se reduce el tamaño de la celda y, por tanto, del dispositivo donde se utilice, se ha diseñado otra celda con las mismas características que la anterior, pero, en

esta ocasión, utilizando SRs que son eléctricamente dos veces más pequeños que los SRRs. El layout final del dispositivo diseñado con SRs se muestra en la figura 7 del artículo K. En la figura 8 del artículo K se ha representado la respuesta frecuencial medida para dicho dispositivo. Aunque en este dispositivo presenta un ligero aumento en las pérdidas de inserción (alrededor de $-5dB$), se ha conseguido una reducción de tamaño del 60% respecto del dispositivo convencional.

En conclusión, se ha demostrado que líneas artificiales basadas en metamateriales pueden ser utilizadas para la síntesis de circuitos de microondas. En concreto se han utilizado celdas en tecnología CPW basadas en SRRs. Se han diseñado, fabricado y medido dos divisores de potencia, llegando a obtener una reducción de un 60% en tamaño, respecto a un dispositivo convencional, y que puede ser aumentada utilizando algunos de los resonadores propuestos en esta tesis. Debido a que el uso de estos dispositivos queda limitado para su uso en aplicaciones de banda estrecha, son de especial interés en aquellas ocasiones donde sea necesario un filtrado de la señal de salida.

4.3 Filtros Paso Bajo

En el **artículo L** se presenta un filtro paso bajo en tecnología microstrip obtenido mediante la utilización de varios OCSRRs en cascada. El dispositivo consiste en grabar de forma periódica varios OCSRRs en una línea de transmisión microstrip, como se muestra en la figura 1 del artículo L. De esta forma, a la frecuencia de resonancia de los OCSRRs se produce un cero de transmisión. Los OCSRRs, al ser eléctricamente pequeños, el dispositivo diseñado también lo es. En este tipo de filtro aparece una banda espuria cercana a la frecuencia de corte, que puede ser eliminada mediante la inclusión de nuevos resonadores sintonizados a dicha frecuencia. En concreto, se pretende diseñar un filtro con las especificaciones que se muestran en la figura 2 del artículo L. El filtro debe presentar un rizado menor a $1dB$ en la banda de paso, frecuencia de corte a $1.33GHz$ y un rechazo de al menos $40dB$ a partir de $1.45GHz$ y que se mantenga al menos hasta $2GHz$. Para conseguir la transición abrupta con el menor número de etapas posibles, se han diseñado los resonadores de forma que presenten una baja inductancia L_C y alta capacidad C_C , de forma que su resonancia sea lo mas estrecha posible y en consecuencia la pendiente de la banda de transición se maximice [117]. En la figura 3 del artículo L se representa el coeficiente de transmisión para el dispositivo con diferentes etapas,

donde se puede observar que, a mayor número de etapas, mayor pendiente presenta la transición, al mismo tiempo que se aumenta el ancho de la banda de rechazo. El filtro resultante, que cumple las especificaciones requeridas, es el que se muestra en la figura 4(a) del artículo L, y su respuesta frecuencial se puede ver en la figura 5 del artículo L. Aunque el filtro ya cumple las especificaciones requeridas, éste presenta una banda espuria relativamente cerca de la frecuencia de corte. Esta banda espuria puede ser eliminada añadiendo otros resonadores sintonizados en dicha banda, que corresponden a los cuatro resonadores de menor tamaño que se han añadido al filtro anterior. El layout del filtro en el que se elimina la banda espuria se muestra en la figura 4(b) del artículo L. En la figura 5 del artículo L se muestra la respuesta del filtro con y sin la banda espuria, junto a la respuesta ideal de un filtro paso bajo elíptico de orden 7, que es el necesario para obtener la transición requerida. Para mostrar cómo actúan los resonadores a diferentes frecuencias, en la figura 14 del artículo C se muestra el diagrama de corrientes, obtenido mediante el software *Agilent Momentum*. Las frecuencias que se han representado en la figura 14 del artículo C están situadas en la banda de paso, $0.7GHz$, donde se puede observar que la señal pasa de un puerto al otro; a una frecuencia de la banda de rechazo principal, $1.6GHz$, donde los resonadores más grandes impiden la transmisión de la señal, resultando “invisibles” para la radiación los resonadores de menor tamaño; y otra frecuencia en la banda espuria, $2.2GHz$, donde se observa que son los resonadores de menor tamaño los que inhiben la transmisión de la señal [118].

4.4 Filtros Elípticos

Un filtro paso bajo elíptico tiene interés para aquellas aplicaciones de microondas en las que es importante la eliminación de armónicos y espurios no deseados, especialmente en aquellos dispositivos que además requieran una transición muy abrupta. La respuesta para un filtro paso bajo elíptico puede verse en la figura 4.5. La principal cualidad de este tipo de filtros, además de la abrupta transición, es la existencia de ceros en la banda de rechazo, causados por la presencia de resonadores.

Las redes circuitales más comunes que se utilizan para implementar este tipo de filtros son las que se muestran en la figura 4.6. Los valores de los elementos para un filtro paso bajo elíptico normalizado con un rizado en la frecuencia de paso de $L_{Ar} = 0.1dB$ están representados en la tabla 4.1.

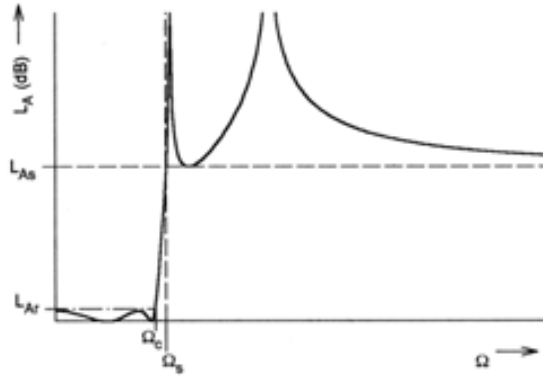


Figura 4.5: Respuesta paso bajo normalizada de un filtro elíptico. Ω es la frecuencia normalizada, Ω_C es la frecuencia de corte, L_{Ar} las pérdidas de inserción en la banda de paso, Ω_S frecuencia de comienzo de la banda de rechazo y L_{As} las pérdidas de inserción en la banda de rechazo.

La gran mayoría de los filtros elípticos diseñados en tecnología planar que se pueden encontrar en la bibliografía han sido diseñados en tecnología microstrip [119–126], aunque también se encuentran en tecnología CPW [127]. En el **artículo M** se propone diseñar un filtro paso bajo elíptico con la configuración del circuito de la figura 4.6(a). Para los resonadores en serie se han utilizados los OCSRrs, presentados en esta tesis. Debido al pequeño tamaño de estos resonadores, es de esperar que el tamaño del filtro también lo sea. Para las capacidades en derivación se han utilizado parches metálicos en la parte posterior del sustrato, conectados a masa mediante vías, de forma que se pueden obtener altos valores de capacidad con secciones cortas de línea. Se ha diseñado un filtro de orden 5, con un rizado $L_{Ar} = 0.1dB$, una frecuencia de corte $f_C = 1GHz$ y un rizado en la banda de rechazo de $L_{As} = 39.59dB$ que comienza a $1.4085GHz$ (en la tabla 4.1 podemos ver los elementos de dicho filtro). Los valores necesarios para el filtro diseñado, cuyo esquema circuital se muestra en la figura 1(a) del artículo M, son $C_1 = 3.20pF$, $C_2 = 0.58pF$, $L_2 = 9.44nH$, $C_3 = 5.02pF$, $C_4 = 1.73pF$, $L_4 = 6.87nH$ y $C_5 = 2.41pF$. En la figura 6(a) del artículo M, se representa el layout del filtro diseñado. Consiste en dos OCSRrs grabados en la tira de metal central de la línea de transmisión CPW (proporcionando los resonadores en serie del circuito de la figura 1(a) del artículo M), intercalados por los parches metálicos en el otro lado del sustrato y conectados a masa mediante vías (proporcionando las capacidades en derivación del circuito de la figura 1(a) del artículo M). Dichas capacidades se

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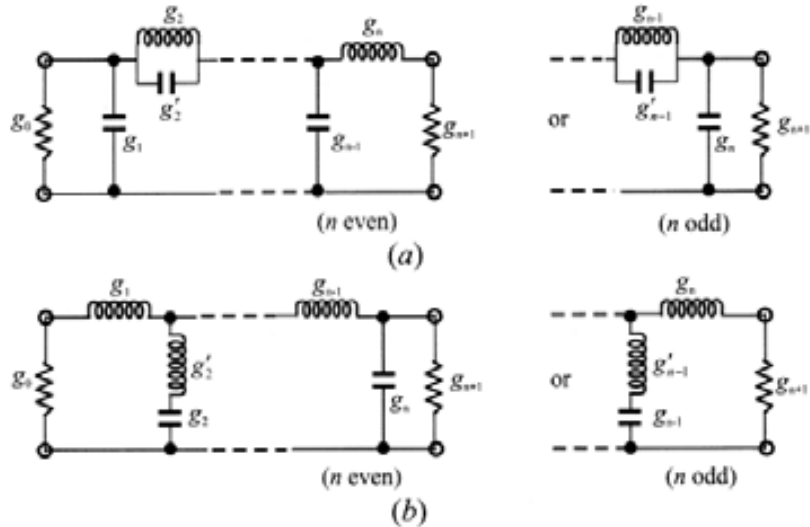


Figura 4.6: Prototipos de filtros paso bajo elípticos con resonadores paralelo en serie (a) y con resonadores serie en derivación (b). Ambos circuitos son versiones duales.

n	Ω_c	L_{-40} dB	g_1	g_2	g'_2	g_3	g_4	g'_4	g_5
3	1.4493	13.5698	0.7427	0.7096	0.5412	0.7427			
	1.6949	18.8571	0.8333	0.8439	0.3252	0.8333			
	2.0000	24.0012	0.8949	0.9375	0.2070	0.8949			
	2.5000	30.5161	0.9471	1.0173	0.1205	0.9471			
4	1.2000	12.0856	0.3714	0.5664	1.0929	1.1194	0.9244		
	1.2425	14.1259	0.4282	0.6437	0.8902	1.1445	0.9289		
	1.2977	16.5343	0.4877	0.7284	0.7155	1.1728	0.9322		
	1.3962	20.3012	0.5675	0.8467	0.5261	1.2138	0.9345		
	1.5000	23.7378	0.6282	0.9401	0.4073	1.2471	0.9352		
	1.7090	29.5343	0.7094	1.0688	0.2730	1.2943	0.9348		
	2.0000	36.0438	0.7755	1.1765	0.1796	1.3347	0.9352		
5	1.0500	13.8785	0.7081	0.7663	0.7357	1.1276	0.2014	4.3812	0.0499
	1.1000	20.0291	0.8130	0.9242	0.4934	1.2245	0.3719	2.1350	0.2913
	1.1494	24.5451	0.8726	1.0084	0.3845	1.3097	0.4991	1.4450	0.4302
	1.2000	28.3031	0.9144	1.0652	0.3163	1.3820	0.6013	1.0933	0.5297
	1.2500	31.4911	0.9448	1.1060	0.2694	1.4415	0.6829	0.8827	0.6040
	1.2987	34.2484	0.9681	1.1366	0.2352	1.4904	0.7489	0.7426	0.6615
	1.4085	39.5947	1.0058	1.1862	0.1816	1.5771	0.8638	0.5436	0.7578
	1.6129	47.5698	1.0481	1.2416	0.1244	1.6843	1.0031	0.3540	0.8692
	1.8182	54.0215	1.0730	1.2741	0.0919	1.7522	1.0903	0.2550	0.9367
	2.000	58.9117	1.0876	1.2932	0.0732	1.7939	1.1433	0.2004	0.9772

Tabla 4.1: Valores de los elementos para filtros prototipos paso bajo elíptico normalizado ($g_0 = g_{n+1} = 0.1$, $\Omega_C = 1$, $L_{Ar} = 0.1dB$)

han diseñado de esta forma para disminuir el efecto inductivo que se obtendría al utilizar secciones de línea, además de para disminuir considerablemente el tamaño del dispositivo. Para determinar las dimensiones de estos parches que dan los valores de las capacidades deseados (C_1 , C_3 y C_5), se ha simulado una sección de línea con diferentes dimensiones de parches y tras un análisis de los parámetros-S, donde se ajusta la respuesta a un circuito en π de una línea de transmisión [128], se obtiene una relación entre las dimensiones del parche y la capacidad en derivación. En la figura 3 del artículo M se representan los valores de capacidades obtenidos para diferentes dimensiones del parche. De esta forma, quedan determinadas las dimensiones de los parches que cumplan los valores deseados de capacidad en derivación. Las dimensiones de los OCSRrs vienen determinadas por las frecuencias de resonancia requeridas y cuya pendiente de susceptancia se ajusta a la del tanque resonante LC ideal, necesario para cada uno de los resonadores del filtro elíptico [62, 129]. No es suficiente con ajustar la frecuencia de resonancia de los OCSRrs, sino que es necesario obtener valores concretos de inductancia y capacidad, lo que hace que sea necesario modificar sustancialmente las dimensiones de los dos OCSRrs, de forma que se obtengan resonadores que ajusten sus pendientes de susceptancia como se muestra en la figura 4 del artículo M. La obtención de los valores de L y C de los OCSRrs, se ha utilizado la ecuación de la susceptancia para un resonador que tiene susceptancia cero a la frecuencia de resonancia. La ecuación para la pendiente de susceptancia es la siguiente [62]:

$$b = \frac{\omega_0}{2} \left. \frac{dB(\omega)}{d\omega} \right|_{\omega=\omega_0} \quad (4.1)$$

donde $B(\omega_0)$ es la susceptancia del resonador no ideal y ω_0 la frecuencia de resonancia del resonador. A partir de esta ecuación se puede obtener la pendiente para el tanque LC , que es igual a $\omega_0 C$, y junto a la ecuación que determina la frecuencia de resonancia ($\omega_0 = 1/\sqrt{CL}$) podemos obtener los valores de L y C del tanque resonante que modela el resonador. Una vez obtenidos los resonadores y los parches metálicos que cumplen las especificaciones deseadas, se unen debidamente para formar el filtro. Una vez montado el filtro y tras un proceso de optimización, se obtiene el filtro deseado, como se puede ver en la figura 6 del artículo M, donde se representa el layout final, la fotografía del dispositivo fabricado y la respuesta simulada y medida comparada con la respuesta ideal. En la respuesta del filtro diseñado se observa que está en concordancia con la del filtro ideal hasta dos veces la frecuencia de corte. Otro punto a destacar sobre el dispositivo fabricado es que, en la banda de rechazo,

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el coeficiente de transmisión se mantiene por debajo de $25dB$ hasta 4.5 veces la frecuencia de corte. Como se puede observar, en uno de los resonadores se ha utilizado un parche metálico en el lado opuesto del sustrato, conectado a dos partes del resonador mediante vías. Este parche se ha añadido para eliminar resonancias parásitas debidas a modos de propagación indeseados que interferían en la respuesta deseada del dispositivo enmascarando uno de los polos del filtro. En el otro resonador no se ha añadido un parche como el mencionado debido a que no es necesaria la eliminación de este modo ya que no influye negativamente en las especificaciones del filtro diseñado. Como punto final, cabe destacar que las dimensiones del filtro son muy pequeñas, resultando un filtro de $2.5cm$ o, lo que es lo mismo, 0.13λ , donde λ es la longitud de onda en la frecuencia de corte del filtro.

En conclusión, se ha propuesto una metodología para el diseño de filtros paso bajo elípticos en tecnología CPW utilizando OCSRRs. Los OCSRRs proporcionan los ceros en la banda de rechazo, y las capacidades en derivación se han realizado mediante parches en la parte posterior del sustrato que permiten desprestigiar la componente inductiva del trozo de línea donde se implementan. Se ha diseñado, fabricado y medido un filtro paso bajo elíptico de quinto orden, obteniendo una buena respuesta que está en concordancia con la función ideal hasta el doble de la frecuencia de corte. Con esto queda demostrado que los OCSRRs son útiles para el diseño de filtros elípticos paso bajo de pequeñas dimensiones.

Capítulo 5

Conclusiones y líneas futuras de investigación

En esta tesis se han propuesto y caracterizado diferentes resonadores cuyas dimensiones son mucho más pequeñas que las de los resonadores utilizados hasta el momento para la síntesis de estructuras basadas en metamateriales, a la vez que se han presentado diferentes técnicas para reducir el tamaño eléctrico de estos resonadores. Para ver que estos nuevos resonadores pueden ser utilizados en metamateriales y conseguir reducir el tamaño de dispositivos de microondas, se han utilizado dichos resonadores en el diseño de diferentes dispositivos basados en líneas de transmisión artificiales basadas en el modelo resonante. En especial, se han estudiado con detalle celdas unitarias de líneas metamaterial basadas en el modelo resonante con SRRs y CSRRs, para las que se han propuesto nuevos modelos, desde los cuales se ha podido hacer un estudio donde se demuestra la dualidad entre ambas estructuras. Los aspectos más importantes tratados en cada uno de los capítulos de esta tesis pueden resumirse en los siguientes puntos:

- En el capítulo 1 se ha hecho una introducción a los metamateriales. En particular se han presentado las características de los medios efectivos, cuyo periodo es menor a la longitud de onda. Se ha estudiado su comportamiento en función de sus propiedades electromagnéticas efectivas (permitividad y permeabilidad), así como la posibilidad de obtener medios con características exóticas que no se encuentran en la naturaleza, como son los medios zurdos. Se han introducido brevemente los resonadores eléctricamente pequeños, los cuales suelen ser utilizados para la obtención de medios zurdos, que estarían englobados dentro del conjunto de los metamateriales. Además, se presentan diferentes configuraciones para la obtención de este tipo de estructuras y se han mostrado aplica-

ciones de este tipo de medios en el diseño de dispositivos de microondas, como pueden ser filtros de rechazo, desfasadores, antenas, etc.

- En el capítulo 2 se han presentado los resonadores eléctricamente pequeños que se han utilizado hasta el momento en metamateriales, como son el *split ring resonator* (SRR), el *broad-side coupled split ring resonator* (BC-SRR), el *complementary split ring resonator* (CSRR) y el *open split ring resonator* (OCSRR). Se han presentado las características más importantes de estos resonadores y sus modelos circuitales para las frecuencias donde pueden ser considerados eléctricamente pequeños. Asimismo, se han estudiado los principios por los que estos resonadores son eléctricamente pequeños, que vienen determinados principalmente por el acoplamiento que existe entre las partículas que forman el resonador. Se ha mostrado la utilidad de estos resonadores para su uso en la reducción del tamaño de dispositivos de microondas, donde se observa que el tamaño del dispositivo viene determinado por el tamaño del propio resonador. Con el fin de poder reducir aún más el tamaño de los dispositivos donde se usan estos resonadores, se presentan diferentes técnicas para obtener resonadores eléctricamente más pequeños que los utilizados hasta el momento. En los artículos A y B se presentan diferentes técnicas para conseguir reducir el tamaño de los resonadores. Una de estas técnicas consiste en acoplar más de una partícula para formar el resonador, de forma que se va incrementando el acoplamiento total que hace que el resonador disminuya su tamaño. Esta técnica puede ser utilizada en cualquier tipo de resonador basado en el SRR o los basados en su versión complementaria, el CSRR. Otra técnica consiste en utilizar dos niveles de metal, de forma que el acoplamiento que se produce entre las partículas que forman el resonador es a través del substrato donde se implementan. Esta técnica es la que se utiliza en el BC-SRR. En esta tesis se propone conectar estas dos capas de metal mediante vías, de forma que las partículas individuales que forman el resonador resultan ser mucho más largas, obteniéndose una frecuencia de resonancia más baja, además de aprovecharse el acoplamiento *broadside*, lo que se resume en una disminución del tamaño final del resonador. La utilización de estas técnicas dan lugar a diversas topologías obteniéndose resonadores con características muy diferentes, como se muestra en el artículo B. Se ha presentado una selección de los resonadores obtenidos al aplicar estas técnicas y se han caracterizado presentando su tamaño eléctrico, el factor de reducción que se obtiene al acoplar las partículas que forman el resonador, el acoplamiento entre partículas y su factor de

calidad. Además de la caracterización se muestran ejemplos de utilización de estos resonadores para demostrar la posibilidad de reducir el tamaño de líneas metamaterial utilizando estos resonadores.

Además de las técnicas de reducción del tamaño en resonadores, en el artículo C se ha presentado el *open complementary split ring resonador* (OCSR), que es el resonador complementario del OSRR. Este resonador es el complementario del OSRR, como el CSRR es el complementario del SRR. Para este resonador, que se presenta por primera vez en esta tesis, se ha caracterizado y analizado su respuesta frecuencial. Se propone y valida su modelo circuital. También se discute la posibilidad de controlar la eliminación de resonancias superiores a la primera, así como el control del cero de transmisión que aparece tras la primera frecuencia de resonancia.

- En el capítulo 3 se estudian y caracterizan líneas artificiales utilizadas para el diseño de metamateriales. En este capítulo se presenta la mayor parte del trabajo realizado en esta tesis, centrado principalmente en el estudio de las líneas de transmisión artificiales basadas en el modelo resonante. Se presenta de forma somera las líneas artificiales basadas en el modelo dual, artículo G. Estas líneas, consisten en cargar líneas de transmisión convencionales con capacidades en serie e inductancias en derivación, de forma que, tras un diseño adecuado, se pueden obtener rangos frecuenciales de propagación zurda, además de la posibilidad de controlar sus características eléctricas (impedancia característica y fase) más allá de lo que se puede controlar con las líneas convencionales.

La parte principal de este tema se centra en el estudio de las líneas de transmisión artificiales basadas en SRRs. Estas líneas consisten en acoplar resonadores como los SRRs a líneas de transmisión convencionales. En concreto, en esta tesis se han estudiado estas líneas en tecnología coplanar (CPW) y microstrip (artículos E y F). Estas líneas presentan una permeabilidad efectiva negativa ($\mu_{eff} < 0$) alrededor de la frecuencia de resonancia de los resonadores, de forma que se inhibe la propagación a dichas frecuencias. Además, si estas líneas se cargan con conexiones a masa, a modo de inductancia en derivación, se obtiene una estructura que presenta características de propagación zurda en un rango frecuencial alrededor de la resonancia del resonador. Para el estudio de estas líneas nos hemos centrado en el comportamiento de la celda unidad, para el cual se ha propuesto un nuevo modelo circuital que mejora el modelo anterior ya que explica fenómenos que se observan en la respuesta frecuencial electromagnética y que el modelo anterior no explicaba. El nuevo mo-

delo circuital se valida tanto para estructuras en tecnología CPW como microstrip, tanto para configuraciones con $\mu_{eff} < 0$ como para propagación zurda. También se presenta y valida un método de extracción de los parámetros del circuito equivalente para la estructura a partir de la respuesta frecuencial electromagnética simulada o medida. Haciendo uso del método de extracción propuesto, se discute la posibilidad de extraer los valores de los elementos del tanque resonante que modelan los resonadores. A partir del modelo circuital de la estructura se hace un estudio que relaciona el tamaño eléctrico del resonador con su ancho de banda (artículo H), obteniéndose que no es posible implementar un resonador que sea eléctricamente más pequeño y al mismo tiempo tenga un ancho de banda mayor, debido a que al disminuir el tamaño del resonador se disminuye el acoplamiento del mismo con la línea que lo excita, y este parámetro de acoplamiento es un factor importante en el ancho de banda.

Al igual que para el caso de líneas de transmisión artificiales basadas en SRRs, se ha hecho un estudio para las líneas artificiales basadas en CSRRs. Aunque este tipo de líneas no forma parte fundamental de esta tesis, se ha estudiado porque se presenta la dualidad que existe entre este tipo de línea basada en CSRRs con las basadas en SRRs, artículos D y G. Estas líneas consisten en acoplar CSRRs a una línea de transmisión convencional, obteniendo permitividad efectiva negativa ($\epsilon_{eff} < 0$) alrededor de la frecuencia de resonancia de los resonadores, de forma que se inhibe la propagación a dichas frecuencias. Además, si se practican unos cortes sobre la línea base, a modo de capacidades en serie, se obtiene una estructura que presenta características de propagación zurda en un rango frecuencial alrededor de la resonancia del resonador. Para esta estructura también se muestra su modelo circuital y una extracción de sus parámetros.

Para terminar con este capítulo, en los artículos I y D, se presenta la dualidad que existe entre las líneas artificiales basadas en SRRs y CSRRs y se muestran las diferencias generales que existen entre las líneas basadas en el modelo resonante y en el modelo dual. Respecto a la dualidad entre las líneas artificiales basadas en SRRs y CSRRs se ha obtenido que, además de que sus modelos circuitales son “duales circuitales”, su respuesta frecuencial varía de forma similar cuando se varía de igual forma la inductancia en derivación, para el caso de las líneas basadas en SRRs, y la capacidad en serie, para el caso de las líneas basadas en CSRRs, por lo que se concluye con que estas dos estructuras son estructuras prácticamente duales. Además del estudio sobre la dualidad,

se exponen las diferencias que hay entre el modelo resonante y el modelo dual en líneas artificiales, basadas principalmente en el hecho de que en el modelo resonante existe un cero de transmisión justo antes de la banda de transmisión, debido a la capacidad de acoplamiento en derivación, para el caso basado en CSRRs, y debido a la inductancia del resonador, para el caso basado en SRRs. Para el caso del modelo dual, no obstante, el cero de transmisión se presenta en el origen.

- En el capítulo 4 se presentan diferentes aplicaciones donde se han utilizado líneas metamaterial con resonadores propuestos en esta tesis, para el diseño de dispositivos con dimensiones menores a las que se obtendrían para un dispositivo convencional.

En el artículo J se han utilizado diferentes resonadores de los presentados en esta tesis para reducir el tamaño de un filtro notch. Este tipo de filtros pueden ser implementados mediante líneas de transmisión metamaterial basadas en el modelo resonante, en las que en torno a la resonancia del resonador acoplado a la línea, se obtiene $\mu_{eff} < 0$ si se acoplan SRRs o $\varepsilon_{eff} < 0$ si son CSRRs. En este trabajo hemos utilizado resonadores con dos niveles de metal acoplados a una línea microstrip, obteniendo un rango frecuencial donde se inhibe la propagación al obtener $\mu_{eff} < 0$. En los ejemplos propuestos se consigue reducir el tamaño del filtro en un factor 2 y 4 respecto al filtro diseñado con BC-SRRs. El diseño de estos filtros, utilizando diferentes resonadores, se han realizado manteniendo el tamaño del filtro, por lo que la disminución se ve reflejada en la frecuencia de operación del filtro. También se han diseñado los filtros de forma que su frecuencia de operación sea la misma, de forma que la reducción de tamaño se ve directamente sobre las dimensiones del filtro. En ambos casos se obtiene el mismo factor de reducción que existe entre los resonadores.

En el artículo K, se han utilizando líneas artificiales basadas en SRRs en tecnología CPW, para el diseño de divisores de potencia de reducidas dimensiones. En este caso se ha aprovechado la posibilidad de controlar la impedancia característica y fase de la celda para obtener una línea con $Z_B = 35.35\Omega$ y $\beta l = -90^\circ$ a una frecuencia de $1.5GHz$. Estas son las características que tiene que tener la línea de entrada de un divisor de potencia de dos salidas de 50Ω . En este caso se han diseñado las celdas con SRRs y con espirales (*spiral resonator*, SR) llegando a obtener un 40% y 60% de reducción en el tamaño del dispositivo respecto a uno convencional.

En el artículo L se ha diseñado un filtro paso bajo que presenta una

transición abrupta, al que además se le han eliminado los espurios que aparecen más cercanos a la banda de transmisión, obteniendo una banda de rechazo ancha. Para el diseño de este filtro se han utilizado OCSRrs en serie sobre una línea en tecnología microstrip. Al conectar los resonadores en serie, se inhibe la transmisión de la señal alrededor de la frecuencia de resonancia, obteniéndose una banda de rechazo. Al añadirle otros OCSRrs en cascada, con dimensiones diferentes, sintonizados a las frecuencias donde aparecen los primeros espurios, se consigue ampliar la banda de rechazo. El tamaño final del filtro obtenido es de 0.77 veces la longitud de onda a la frecuencia de corte del filtro, lo que confirma que este filtro es de pequeñas dimensiones.

En el artículo M se ha diseñado un filtro elíptico de 5º orden en tecnología CPW, utilizando la configuración de resonadores paralelo en serie y capacidades en derivación. Los resonadores paralelos en serie se han obtenido mediante OCSRrs y las capacidades en derivación se han obtenido con parches metálicos, por debajo de la línea CPW, conectados a masa mediante vías, minimizando el efecto inductivo de la línea. El filtro obtenido presenta una buena respuesta semejante a la función elíptica ideal hasta el doble de la frecuencia de corte y la banda de rechazo se mantiene por debajo de $25dB$ hasta cuatro veces la frecuencia de corte del filtro. Además, como los OCSRrs son resonadores eléctricamente pequeños, las dimensiones del filtro también lo son, siendo la longitud del filtro 0.13 veces la longitud de onda a la frecuencia de corte. Con estos resultados se demuestra que los OCSRrs pueden ser utilizados para el diseño de dispositivos de microondas de reducidas dimensiones.

Con los resultados obtenidos en esta tesis podemos afirmar que se han cumplido los objetivos iniciales, quedando de manifiesto la posibilidad de reducir las dimensiones de los resonadores utilizados para la síntesis de dispositivos basados en metamateriales, pudiendo reducir de forma notable las dimensiones de dichos dispositivos, útiles para el diseño de circuitos de comunicaciones. Además de proponer nuevos resonadores y técnicas para la reducción de su tamaño eléctrico, se han propuesto técnicas de extracción de parámetros que son muy útiles para la caracterización de los resonadores con topologías complicadas.

Como líneas futuras de investigación, las cuales ya se están llevando a cabo o empezarán a llevarse a cabo en un breve, está la obtención de líneas artificiales basadas en SRRs balanceadas en tecnología CPW, de forma que se puedan utilizar para el diseño de dispositivos de banda ancha, como por ejemplo filtros

paso alto. Al igual que se ha estudiado la relación que existe entre el ancho de banda y el tamaño de los resonadores basados en el SRR, se propone hacer este estudio para resonadores complementarios basados en el CSRR. Una vez propuestos métodos de extracción de parámetros para los modelos circuitales, a partir de la simulación electromagnética o medida de la estructura, sería de gran ayuda y de importante relevancia para el diseño de dispositivos basados en estas estructuras, estudiar el proceso inverso. La determinación de las dimensiones de la estructura a partir de los parámetros circuitales implicaría un ahorro importante en el tiempo de diseño, debido a la rapidez de simulación que se obtiene al trabajar con simuladores circuitales en lugar de simuladores electromagnéticos, cuyo tiempo de simulación es muy elevado respecto a la simulación circuital. Esta línea de investigación se ha abierto recientemente, utilizando un método basado en *space mapping* que permite la obtención de la topología de una estructura a partir de los parámetros eléctricos. Este proyecto se está realizando en colaboración con el grupo GAM (Grupo de Aplicaciones de Microondas) del ITEAM (Instituto de Telecomunicaciones y Aplicaciones Multimedia) de la Universidad Politécnica de Valencia. Otras líneas de investigación están en la combinación de resonadores para la obtención de dispositivos con varias bandas de operación. Por otro lado, dentro de las posibilidades que brindan las líneas de transmisión metamaterial, se está estudiando la aplicación de estas técnicas para dispositivos implementados sobre sustratos de papel, con lo que se puede obtener una importante reducción en el coste y peso de los dispositivos, como por ejemplo antenas para RFID.

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Apéndice A

Relación de Artículos

Relación de artículos que forman parte de esta tesis doctoral. Los artículos marcados con asterisco no pueden ser considerados como parte fundamental de esta tesis, por estar publicados en *proceedings* de congresos no indexados, o por no estar aceptados en el momento del depósito de esta tesis doctoral. Aún así han sido incluidos ya que en el texto se hace referencia a resultados que aparecen en los mismos.

- A.- * F. Aznar, J. Bonache, F. Martín, E. Ozbay, K. B. Alici, F. Bilotti, S. Tricarico, L. Vegni, J. D. Baena, L. Jelinek and R. Marqués, “Miniaturization and Characterization of Metamaterial Resonant Particles”, *European Microwave Conference (EuMC)*, Amsterdam (The Netherland), pp. 269-272. October 2008.
- B.- F. Aznar, J. García-García, M. Gil, J. Bonache and F. Martín, “Strategies for the miniaturization of metamaterial resonators”, *Microwave and Optical Technology Letters*, Vol. 50, No. 5, pp. 1263-1270. May 2008.
- C.- * F. Aznar, A. Vélez, M. Durán-Sindreu, J. Bonache and F. Martín, “Open complementary splits ring resonators (OCSRRs) modelling, analysis and prospective applications”, *New Journal of Physics*. Enviado.
- D.- * F. Aznar, M. Gil, J. Bonache, and F. Martín. “Revising the Equivalent Circuit Models of Resonant-Type Metamaterial Transmission Lines”, *IEEE MTT-S International Microwave Symposium Digest*, Atlanta (USA), pp. 323-326. June 2008.
- E.- F. Aznar, J. Bonache and F. Martín, “Improved circuit model for left-handed lines loaded with split ring resonators”, *Applied Physics Letters*, Vol. 92, pp. 043512. February 2008.

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- F.- F. Aznar, M. Gil, J. Bonache, L. Jelinek, J. D. Baena, R. Marqués and F. Martín, “Characterization of miniaturized metamaterial resonators coupled to planar transmission lines”, *Journal of Applied Physics*, Vol. 104, pp. 114501. December 2008.
- G.- F. Aznar, M. Gil, J. Bonache and F. Martín, “Modelling metamaterial transmission lines: a review and recent developments”, *Opto-Electronics Review*, Vol. 16, No. 3, pp. 135-145. September 2008. **Invited paper.**
- H.- * F. Aznar, J. Bonache and F. Martín, “Analysis of resonant-type metamaterial transmission lines on the basis of their equivalent circuit models”, *Asia Pacific Microwave Conference (APMC)*, Hong Kong (China). December 2008.
- I.- * F. Aznar, M. Gil, J. Bonache and F. Martín, “SRR- and CSRR-loaded metamaterial transmission lines: a comparison to the light of duality”, *International Congress on Advanced Electromagnetic Materials in Microwaves and Optics (Metamaterials)*, Pamplona (Spain). September 2008. **Invited.**
- J.- F. Aznar, M. Gil, J. Bonache, J. García-García and F. Martín, “Metamaterial transmission lines based on broad-side coupled spiral resonators”, *Electronics Letters*, Vol. 43, No. 9, pp. 530-532. April 2007.
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- L.- F. Aznar, A. Vélez, J. Bonache, J. Menés and F. Martín, “Compact low pass filters with very sharp transition bands based on open complementary split ring resonators”, *Electronics Letters*, Vol. 45, No. 6, pp. 316-317. March 2009.
- M.- * F. Aznar, A. Vélez, M. Durán-Sindreu, J. Bonache and F. Martín, “Elliptic-function CPW Low-Pass Filters Implemented by Means of Open Complementary Split Ring Resonators (OCSRRs)”, *IEEE Microwave Wireless Components Letters*. Enviado.

Artículo A

Referencia:

F. Aznar, J. Bonache, F. Martín, E. Ozbay, K. B. Alici, F. Bilotti, S. Tricarico, L. Vegni, J. D. Baena, L. Jelinek and R. Marqués, “Miniaturization and Characterization of Metamaterial Resonant Particles”, *European Microwave Conference (EuMC)*, Amsterdam (The Netherland), pp. 269-272. October 2008.

Resumen del artículo:

Este trabajo se centra en la miniaturización y caracterización de resonadores semidiscretos, de interés para la síntesis de estructuras basadas en metamateriales tales como líneas de transmisión metamaterial, superficies selectivas de frecuencia y elementos radiantes, entre otros. Las partículas consisten en patrones metálicos grabados en un dieléctrico y se inspiran en el *split ring resonator*, SRR (es decir, la partícula resonante anteriormente utilizada para la síntesis de metamateriales zurdos). Se discuten diferentes estrategias de miniaturización y se dan ejemplos. Se demuestra que mediante el uso de dos niveles de metal conectados mediante vías es posible lograr partículas de tamaño eléctrico muy pequeño (es decir, por debajo de $\lambda/100$, donde λ es la longitud de onda en el sustrato considerado a la frecuencia de resonancia). También se presenta un método para determinar los parámetros eléctricos de los resonadores metamaterial en configuraciones de línea de transmisión, y se discute la posibilidad de determinar las características de las partículas aisladas. Por último, se presentan ejemplos de aplicaciones de esta técnica. Este trabajo es de interés para la síntesis de los medios efectivos metamaterial basados en elementos resonantes.

Miniaturization and Characterization of Metamaterial Resonant Particles

¹Francisco Aznar, ¹Jordi Bonache, ¹Ferran Martín, ²Ekmel Ozbay, ²K. Boratay Alici, ³Filiberto Bilotti,

³Simone Tricarico, ³Lucio Vegni, ⁴Juan D. Baena, ⁴Lukas Jelinek and ⁴Ricardo Marqués

¹GEMMA/CIMITEC, Departament d'Enginyeria Electrònica, Universitat Autònoma de Barcelona, 08193 BELLATERRA, Spain
ferran.martin@uab.es

²Nanotechnology Research Center, Bilkent University, Ankara 06800, Turkey

³Department of Applied Electronics, University "Roma Tre", 00146 Rome, Italy

⁴Departamento de Electrónica y Electromagnetismo, Universidad de Sevilla, 41012 Sevilla, Spain

Abstract— This paper is focussed on the miniaturization and characterization of semi-lumped resonators, of interest for the synthesis of metamaterial-based structures such as metamaterial transmission lines, frequency selective surfaces, absorbers, and radiating elements, among others. The particles consist on metallic patterns etched on a dielectric and are inspired on the split ring resonator, SRR (that is, the formerly resonant particle used for the synthesis of left handed metamaterials). The different strategies for miniaturization are discussed and examples are given. It is shown that by using two metallic levels connected through vias it is possible to achieve very small electrical size for the particles (namely, below $\lambda/100$, where λ is the wavelength in the considered substrate at resonance). A method to determine the electrical parameters of the resonators in metamaterial transmission line configurations is also presented, and the possibility to determine the characteristics of the isolated particles is discussed. Finally, examples of application of this technique are presented. This work is of interest for the synthesis of effective media metamaterials based on resonant elements.

I. INTRODUCTION

Effective media metamaterials are artificial structures based on periodic inclusions of metals and/or dielectrics that exhibit controllable characteristics. Rather than being based on the composition of the elements forming the medium, the electromagnetic (or optical) properties of effective media metamaterials are based on their structure. Thanks to this controllability, it is possible to synthesize media with unusual properties, including negative refractive index media [1], or invisible cloaks [2], among others. Probably, the most relevant works that have led to the explosion of metamaterials are due to Pendry and co-workers [3], Smith and co-workers [4], and Veselago [5]. In [3], the split ring resonator (SRR) – see Fig. 1a was presented as a small size resonant particle useful for the synthesis of artificial (anisotropic) media with negative effective permeability. In [4], it was synthesized the first left handed medium by combining SRRs with metallic posts (the latter providing the required negative permittivity). Although highly anisotropic (the structure presented in [4] is a bulk one-dimensional left-handed medium), it represented the first synthesis of a medium with simultaneous negative permittivity and permeability (the properties of those media

were already anticipated by Veselago in 1968 [5], and have been experimentally demonstrated from the beginning of this millennium – see [6-9] and references therein).

Multiple works have been devoted to the synthesis of functional artificial structures based on SRRs. This includes, metamaterial transmission lines (and their microwave applications [10,11]), admittance surfaces [12], absorbers [13], radiating elements [14,15], imaging devices [16], etc. (see [9] for more details). One key aspect in many of these applications is the miniaturization of the resonant particles. To reduce the electrical size of SRRs, it is necessary to enhance the coupling between the individual rings [17]. In a uni-planar configuration (edge coupling), the limit is given by the minimum distance between etched metals, which is of the order of 50-100 μm for typical microwave substrates and etching systems (drilling machine or photo/etching system). With such critical dimensions, it is not possible to achieve particle diameters smaller than $\lambda/10$. The electrical size of the particle can be reduced by etching the individual rings face-to-face at different metal levels (broadside coupling), see Fig. 1(c). By using thin dielectrics, the distributed capacitance between the rings can be increased and the first resonance frequency significantly reduced (SRR diameters of about $\lambda/30$ have been demonstrated by using this approach [18]). Another possibility is to modify the topology so that the slits are etched in the same position and the inner and outer rings are cross connected [19,20]. The resulting topology is indeed a two-turn spiral (Fig. 1b) whose resonance frequency is half the resonance frequency of the SRR (the particle inductance is preserved but the capacitance is enhanced by a factor of four [19]).

In this paper, different strategies for particle miniaturization are presented. Miniaturization levels superior to $\lambda/100$ are demonstrated. In some cases, due to their complexity, it is difficult to derive analytical models for such particles. For this reason, we have developed a method to obtain the electrical parameters of the resonators when they are coupled to transmission lines. The possibility to obtain the parameters of the decoupled particles is discussed, and examples of particle characterization are provided.

II. STRATEGIES FOR PARTICLE MINIATURIZATION

The strategies for the miniaturization of metamaterial resonators either consist on using two metallic levels connected by vias and/or increasing the number of elements constituting the structure. For instance, it is possible to significantly reduce the electrical size of the resonators by implementing a two-turn spiral resonator with either turn etched face-to-face at both sides of a thin substrate and connected through a via (Fig 1f). By means of this topology, we take benefit of the broadside coupling as well as the higher capacitance inherent to the topology, as has been discussed in the previous section with regard to the two turn uni-planar spiral resonator (Fig 1b). We can further reduce the particle size by increasing the number of turns, as is illustrated in Fig. 1(g). With this particle we achieve a high level of miniaturization, which is of the order of $\lambda/100$ provided a substrate with 0.127mm thickness is considered.

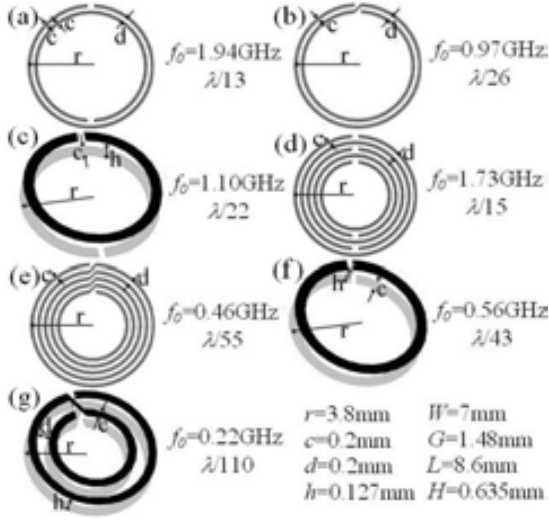


Fig. 1 Topologies of the different resonators considered, and indication of the resonance frequency and size obtained as is explained in the text with the structure depicted in Fig. 2. The considered substrate is Rogers RO3010 with dielectric constant $\epsilon_r=10.2$. Relevant dimensions are: metal strips width $c=0.2\text{mm}$, distance between strips $d=0.2\text{mm}$, external radius $r=3.8\text{mm}$, substrate thickness $h=0.127\text{mm}$ (for resonators formed by two metal layers). For the CPW structure the central strip width is $W=7\text{mm}$, the width of the slots is $G=1.35\text{mm}$, the length of the line is $L=8.6\text{mm}$ and the substrate thickness $H=0.635\text{mm}$.

By increasing the number of turns of the spiral resonator in a uni-planar metal configuration (Fig 1e), the resonance frequency can be significantly reduced, although it tends to saturate. It is also possible to introduce further rings to the topology of the SRR, as Fig. 1(d) illustrates. This also reduces the electrical size due to multiple coupling. The different topologies discussed, as well as their electrical size, are all depicted in Fig. 1. To obtain the electrical size of these particles we have coupled them to a CPW transmission line, as is depicted in Fig. 2, and we have obtained the simulated transmission coefficient of the structure (by using the *Agilent Momentum* commercial software). The transmission zero

coincides with the resonance frequency of the particle. Thus, we can easily obtain the particle size in terms of the wavelength at resonance. We must mention that the resonance frequency obtained by following this procedure corresponds to the resonance of the particle coupled to the line. That is, the presence of the CPW affects the particle to some extent (this aspect will be discussed later). Nevertheless, in many applications, the particles are coupled to a line and it is of interest to infer the parameter under these circumstances.

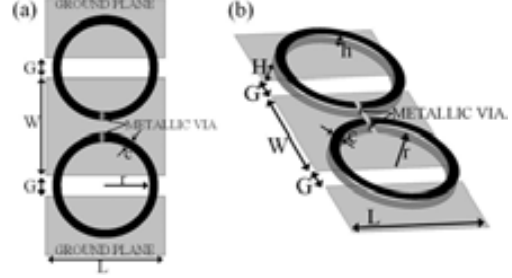


Fig. 2 Typical configuration of a CPW loaded with a pair of metamaterial resonators (Fig 1f) used to determine their resonance frequency, top (a) and 3D (b) view. Relevant dimensions are indicated in Fig. 1.

III. PARAMETER EXTRACTION

The structure of Fig. 2 can be described by the circuit model of Fig. 3(a) [21]. The resonators are described by the resonant tank L_s-C_s , M is the mutual inductive coupling between the line and the resonators, L is the line inductance and C the line capacitance. Notice that, contrary to the model reported in [21], in the model of Fig. 3(a) we have not used the magnetic wall concept since it is not actually necessary. This circuit model can be transformed to that of Fig. 3(b) with the following transformations:

$$C_s' = \frac{L_s}{2\omega_o^2 M^2} \quad (1)$$

$$L_s' = 2\omega_o^2 M^2 C_s \quad (2)$$

$$L' = L - L_s' \quad (3)$$

where $\omega_o=2\pi f_o$, f_o being the resonance frequency of the resonators:

$$f_o = \frac{1}{2\pi} \sqrt{\frac{1}{L_s C_s}} = \frac{1}{2\pi} \sqrt{\frac{1}{L' C_s}} \quad (4)$$

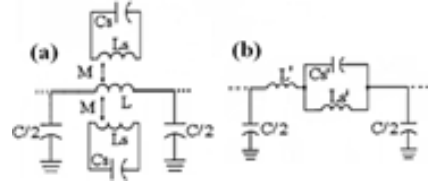


Fig. 3 Lumped element equivalent circuit model of a transmission line loaded with a pair of metamaterial resonant particles (a) and transformed model obtained by the procedure indicated in the text (b).

To determine the parameters of the circuit model of Fig. 3(b) we need four conditions. From the representation of the reflection coefficient, S_{11} , in the Smith Chart, two conditions are obtained. On one hand, we can determine the frequency that nulls the series reactance, f_z , from the intercept of S_{11} with the unit conductance circle. This is obvious since at this frequency, the real part of the admittance seen from the ports is simply the admittance of the opposite port, that is, $Y_o=1/Z_o=(50\Omega)^{-1}=0.02S$. This frequency is given by the following expression:

$$f_z = \frac{1}{2\pi} \sqrt{\frac{1}{L_s C_s'} + \frac{1}{L C_s'}} \quad (5)$$

On the other hand, the susceptance of the unit cell seen from the ports at f_z , which can be inferred from the Smith Chart, is:

$$B(\omega_z) = \frac{C\omega_z}{2} \quad (6)$$

with $\omega_z=2\pi f_z$. Another condition is obtained from the representation of transmission coefficient, S_{21} , since the transmission zero frequency, which coincides with the resonance frequency of the particles and can be easily identified, is given by expression (4). Finally, another condition can be deduced from the phase of the transmission coefficient, $\phi_{S_{21}}$. At the frequency where $\phi_{S_{21}}=90^\circ$, $f_{\pi/2}$, the electrical length of the unit cell, $\phi=\beta l$ (β being the phase constant and l the length of the unit cell), is $\phi(f_{\pi/2})=-90^\circ$. Since the dispersion relation of a periodic structure consisting of cascaded unit cells as those of Fig. 2 is given by:

$$\cos\phi = 1 + \frac{Z_s(\omega)}{Z_p(\omega)} \quad (7)$$

it follows that

$$Z_s(\omega_{\pi/2}) = -Z_p(\omega_{\pi/2}) \quad (8)$$

with $\omega_{\pi/2}=2\pi f_{\pi/2}$. Expressions (4)-(6) and (8) are the four conditions needed to univocally determine the circuit parameters of Fig. 3(b). To illustrate the accuracy of the reported method and the validity of the model, we have applied it to determine the parameters of the structure depicted in Fig. 2. Fig. 4 shows the comparison between the electromagnetic and circuit simulation (the circuit parameters have been inferred from the simulated S-parameters of the structure, but it is also possible to infer them from measurement). Good agreement has been obtained, indicating the validity of the circuit model and parameter extraction.

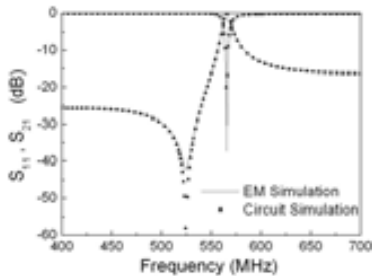


Fig. 4 Frequency response of the structure of Fig. 2 obtained by electromagnetic and circuit simulation. The dimensions are depicted in Fig.1 and the circuit parameters in Table I(f).

IV. APPLICATION OF THE PARAMETER EXTRACTON TECHNIQUE TO THE MINIATURIZED RESONATORS

The method reported in the previous section has been applied to obtain the parameters of the circuit model of Fig. 3(b), corresponding to the CPW transmission lines loaded with the resonant elements considered in Section II. The results are given in Table I. Additionally, in this table are given the values of the elements of the original circuit (Fig. 3a), which have been inferred from transformation (1)-(3) by considering $M=Lf$, where f is the fraction of the slots occupied by the resonant elements. This assumption is roughly valid if (i) the resonant elements can be described through a quasi-static analysis, and (ii) the magnetic flux lines generated by the line in the region of the resonators penetrate the whole resonator area. To satisfy the latter requirement it is necessary to use moderate or narrow substrates.

TABLE I
ELEMENT VALUES FOR THE DIFFERENT RESONATORS CONSIDERED IN FIG. 1

	(a)	(b)	(c)	(d)	(e)	(f)	(g)
C (pF)	2.51	2.19	2.11	3.55	3.00	2.05	2.45
For the circuit model in Fig. 3(b)							
L' (nH)	1.93	2.13	2.00	1.76	2.18	2.21	2.40
L _{S'} (nH)	0.37	0.42	0.37	0.44	0.52	0.38	0.53
C _{S'} (pF)	17.9	63.4	56.4	19.4	231	206	979
For the circuit model in Fig. 3(a)							
M (nH)	1.72	1.90	1.98	1.02	1.26	2.18	2.17
L (nH)	2.31	2.55	2.37	2.20	2.71	2.60	2.92
L _S (nH)	15.7	17.2	21.2	4.81	6.04	24.7	17.9
C _S (pF)	0.43	1.55	0.99	1.76	20.0	3.22	28.9

It is worth mentioning that the estimated resonator parameters correspond to the particle in the presence of the CPW on the opposite substrate side. However, it is possible to estimate the parameters of the isolated particle (that is, without the presence of the CPW metal level). To this end we make the following assumption: the resonators are considerably distant to the CPW. Under this conditions, the capacitance C_s roughly coincides with the capacitance of the isolated particle, C_l . The inductance of the isolated particle can be inferred from the method of images according to:

$$L_l = L_s + M' \quad (9)$$

where M' is the mutual coupling between the resonator and its image.

We have applied this procedure to the SRR particle since the analytical model of this particle is well known [18]. The parameters of the isolated resonator coincide to a good approximation with those inferred from the method reported in [18] for substrate thicknesses in the vicinity of 0.6mm-0.8mm. For thicker substrates, the values inferred from the analytical model can also be obtained by varying (decreasing) the filling factor f (that is, the second assumption for the validity of $M=Lf$ is not satisfied). For thinner substrates the distributed capacitance between the individual rings of the SRR is influenced by the presence of the CPW metal level, and this complicates C_l calculation. In deriving (9), it has been considered that the effect of the slots (CPW) is negligible.

According to these words, to accurately derive the parameters of the isolated particles, we need to know M accurately, and this is not so simple because we can not consider that the optimized substrate thicknesses for the SRR hold for other particles. Indeed, this has been corroborated since it has been found that the optimized substrate thickness for the broadside coupled SRR is different. Nevertheless, for many applications based on the transmission line model of Fig. 3(b), the reported parameter extraction techniques is very valuable, since we are able to determine not only the resonator parameters, but also the line parameters very accurately. For the isolated particles, the method is only able to provide approximate values of the resonator parameters. This method is of special interest for those structures based on complex resonators which can not be easily modelled.

V. CONCLUSIONS

In conclusion, we have proposed small size planar resonators of interest for the synthesis of effective media metamaterials. Dimensions as small as $\lambda/100$ (λ being the wavelength at resonance) have been demonstrated. We have also proposed a technique for the determination of resonator parameters when they are coupled to a planar transmission line. This is of interest in many applications based on transmission line metamaterials. The technique has been applied to various resonant particles. The procedure for the determination of the parameters of the isolated particles has also been discussed, and its limitations have been pointed out. This paper is of interest for the microwave community involved in metamaterial synthesis and their applications.

ACKNOWLEDGMENT

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Artículo B

Referencia:

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Resumen del artículo:

En este artículo, se presentan y discuten diferentes posibilidades para mejorar la miniaturización de partículas resonantes metamaterial. Mediante el uso de dos capas de metal conectadas por vías, se proponen nuevas topologías compactas útiles para la síntesis de metamateriales (incluyendo estructuras zurdas). Se ha realizado un análisis comparativo de las diferentes topologías, en términos de tamaño y factor de calidad descargado. A su vez, se exponen las pertinentes conclusiones.

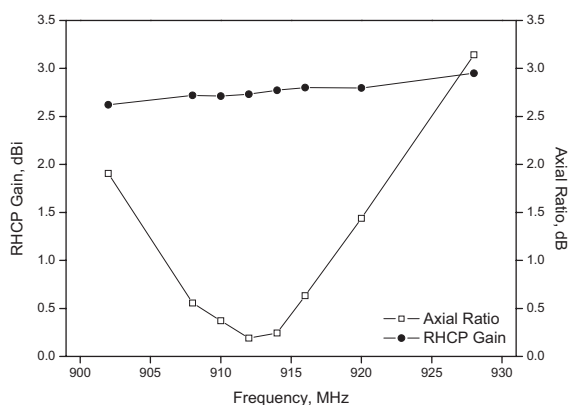


Figure 5 Measured RHCP gain and axial ratio performance

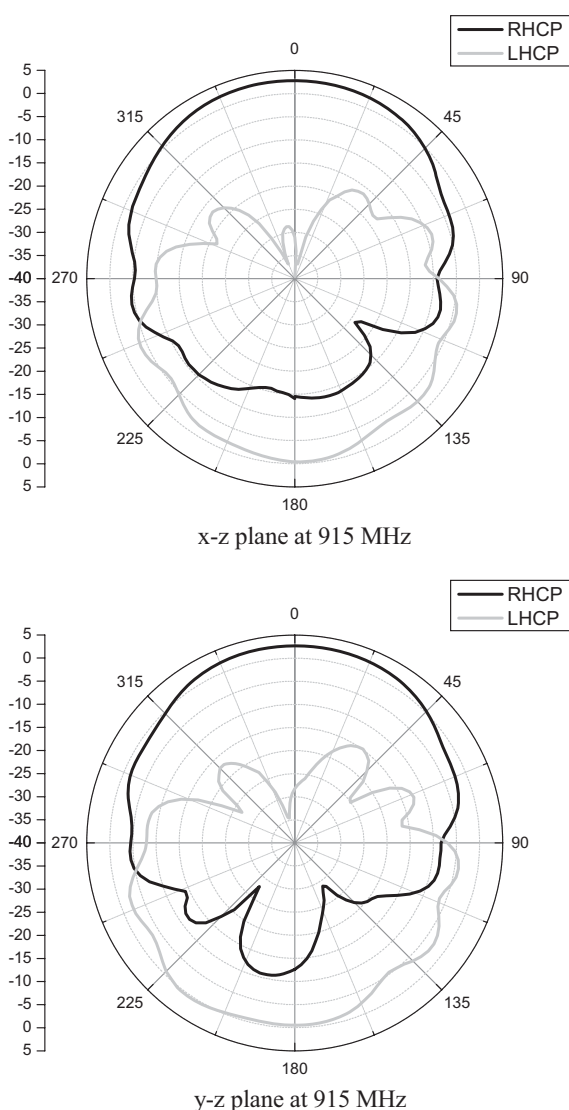


Figure 6 Measured radiation patterns at 915 MHz in the x - z and y - z planes

3. EXPERIMENTAL RESULTS

The proposed aperture-coupled circularly polarized antenna for UHF RFID applications has been fabricated and experimentally analyzed. It should be noticed that a thick high dielectric substrate with a relative permittivity of $\epsilon_r = 8$ is employed for size reduction. Generally, the impedance of an aperture-coupled antenna can be matched when the aperture size is below resonant size. In this article the aperture size is determined to limit the level of back radiation to below -20 dB relative to the main lobe and rectangular stubs are used for effective impedance matching. Figure 4 shows the simulated and measured return loss characteristics. The measured impedance bandwidth (VSWR [1] 1.5) spans 60 MHz from 880 to 940 MHz, which covers the entire North American UHF RFID frequency band. In Figure 5, the measured axial ratio in the broadside direction is plotted. It is observed that a 3 dB axial ratio bandwidth is about 30 MHz or 3.28% with respect to a center frequency of 915 MHz. Figure 5 also illustrates the measured right-handed circular polarization (RHCP) gain within the CP bandwidth. A peak gain of 2.69 dBi and an average gain of 2.58 dBi are obtained in the frequency band of interest. The measured RHCP gain patterns in the two principle planes of x - z and y - z are shown in Figure 6.

4. CONCLUSION

In this article, a novel UHF RFID reader antenna for a handheld applications was presented. The prototype of the proposed aperture-coupled antenna has been fabricated and the measured return loss, axial ratio, and CP gain characteristics are excellent in the North American UHF RFID frequency band. The compact size of $100 \text{ mm} \times 100 \text{ mm} \times 9.6 \text{ mm}$ enables the proposed antenna to be easily integrated with portable RFID reader systems.

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STRATEGIES FOR THE MINIATURIZATION OF METAMATERIAL RESONATORS

Francisco Aznar, Joan García-García, Marta Gil, Jordi Bonache, and Ferran Martín

GEMMA/CIMITEC, Departament d'Enginyeria Electrònica, Universitat Autònoma de Barcelona, 08193 BELLATERRA (Barcelona), Spain; Corresponding author: Francisco.Aznar@uab.es

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ABSTRACT: In this article, different possibilities to enhance the miniaturization of metamaterial resonant particles are presented and dis-

cussed. By using two metal layers connected by vias, new compact topologies useful for the synthesis of metamaterials (including left handed structures) are proposed. A comparative analysis of the different topologies, in terms of size and unloaded quality factor is carried out, and relevant conclusions are inferred. © 2008 Wiley Periodicals, Inc. Microwave Opt Technol Lett 50: 1263–1270, 2008; Published online in Wiley InterScience (www.interscience.wiley.com). DOI 10.1002/mop.23348

Key words: metamaterials; left handed transmission lines; electromagnetic resonators; device miniaturization

1. INTRODUCTION

Metamaterials have been a subject of great interest in the scientific community since the beginning of this millennium, when it was experimentally demonstrated by Smith and coworkers [1, 2]. The existence of left handed wave propagation in an artificial medium composed of metallic wires and split ring resonators (SRRs) [3], and negative refraction in an artificial prism fabricated by means of the same metallic inclusions [2]. This represented the corroboration of the seminal (theoretical) work carried out by Veselago in 1968, where the exotic properties of artificial structures with simultaneous negative permeability and permittivity were anticipated [4].

The study of metamaterials has been very active in recent years. In microwave engineering, applications include the design of compact filters [5] and other microwave components with superior performance and/or exhibiting novel functionalities (see the monographs [6–8] to gain insight on the multiple possibilities of metamaterials in microwave engineering). From the scientific point of view, the demonstration of invisible cloaks operating at microwave frequencies [9] and beyond, and the possibility to extend the metamaterial concepts and applications to the THz and optical regions are hot topics at present [10–12].

To achieve most of the relevant characteristics of metamaterials, the artificial structures must exhibit effective media properties. To this end, the constituent building blocs (SRRs, etc.) must be electrically small, that is, with dimensions significantly smaller than the signal wavelength at the frequencies of interest. In the case of SRRs, this sub-wavelength behavior is achieved through the coupling between the inner and outer rings of the structure [13]. However, besides the subwavelength dimensions, other important requirements in magnetically driven resonators, such as SRRs and other related particles, is the presence of a strong magnetic moment at resonance. Having in mind these considerations, several resonant particles of interest for the synthesis of metamaterials have been already proposed by Marqués and coworkers [14, 15]. The main objective of this work is to demonstrate that by combining two metal levels (at both sides of a dielectric layer) connected by vias with the appropriate topology, it is possible to design new resonant particles with high level of miniaturization. A comparative analysis of the different particles in terms of dimensions and quality factor is pointed out. This analysis reveals that the application of the proposed strategies for area reduction does not represent a severe degradation in the quality factor. The immediate consequence of this fact is that, beyond the applications of these new resonators to the synthesis of metamaterials, they can also be potentially used to reduce the area of planar passive components (the application of conventional SRRs to compact planar filters has been already demonstrated [16]).

2. PRINCIPLE FOR RESONATOR'S MINIATURIZATION

Most of the subwavelength metamaterial resonant particles are constituted by at least two coupled resonators. Size reduction in

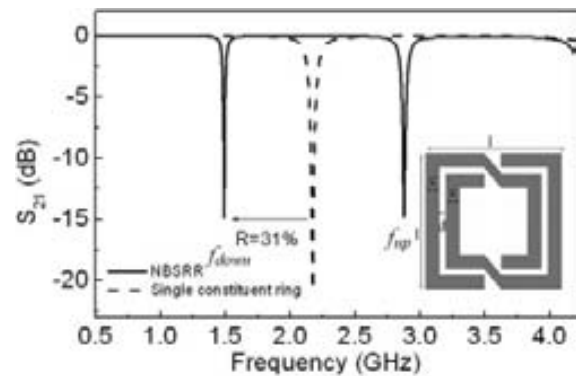


Figure 1 Simulated insertion losses of a microstrip line loaded with a NB-SRR (solid line) and loaded with a single resonator (dashed line). In the inset is represented the topology of the NB-SRR and relevant dimensions, where $l = 9.4$ mm, $c = 0.4$ mm, $d = 0.05$ mm (the Rogers RO3010 substrate with thickness $h = 1.27$ mm and dielectric constant $\epsilon_r = 10.2$ has been considered). The simulation has been performed by means of the Agilent Momentum commercial software

these particles is due to the interaction between these individual resonators. From the physical point of view, this interaction means that there is an effective energy interchange between the individual resonators through the electromagnetic field. Depending on whether the energy interchange is mainly produced through the electric or the magnetic field, such coupling is identified as electric or magnetic. It can be shown that the nature of the coupling can be determined by analysing the symmetry of the current distribution in the resonators at the resonances [13, 16].

In general, from the point of view of the equivalent circuit model of two coupled resonators, the individual resonant particles can be modeled by LC tanks and their electric and magnetic coupling by a mutual capacitance and inductance, respectively. As a consequence of the interaction between the individual resonators, the first resonance frequency of the coupled resonators is driven to small values and, hence, the resulting resonant element (formed by the two coupled individual resonators) can be made electrically small [17]. Actually, the coupling produces a split-off of the fundamental frequency of either single resonator, and two new resonance frequencies appear: one of them smaller (f_{down}) and the other higher (f_{up}) than the resonance frequency of the individual particles. Regardless of the kind of coupling, the coupling coefficient can be estimated by [17]:

$$M = \frac{f_{\text{up}}^2 - f_{\text{down}}^2}{f_{\text{up}}^2 + f_{\text{down}}^2} \quad (1)$$

As an illustrative example, Figure 1 shows the transmission coefficient of a microstrip line loaded with a nonbianisotropic SRR (NB-SRR). At the resonance frequencies of the particle a notch in the transmission coefficient arises. As can be appreciated, the first resonance frequency of the NB-SRR, f_{down} , is substantially smaller (31%) than the fundamental frequency of the single ring (we will call this the reduction ratio, R , from now on). This has been achieved by reducing the space between the rings ($d = 0.05$ mm). According to expression 1, the coupling coefficient has been estimated to be $M = 0.58$. Obviously, f_{down} can be made smaller by further decreasing d . Similar features can be observed in metamaterial resonators formed by nonidentical individual rings exhibiting distinct (although close) resonance frequencies. In this case,

f_{down} is driven below the smaller of the two resonance frequencies of the individual resonators, whilst f_{up} appears above the higher of these resonances [13]. The coupling between the individual resonators and the resulting displacement of f_{down} towards small values is the fundamental principle used by Pendry et al. for the synthesis of electrically small resonant particles, such as the SRR [3]. In the vicinity of the first resonance, an array of such particles can be considered to form a continuous medium, provided the dimensions of the particles can be made small as compared with the wavelength. Size reduction can also be achieved in particles not consisting on two coupled resonators. This is the case of the spiral resonator (SR) [18–21], where selfcoupling within the particle is responsible for resonator miniaturization. However, in this case we can not talk in terms of a coupling coefficient.

3. NEW DESIGN STRATEGIES FOR MINIATURIZATION

The use of two different metal layers connected by vias in the design of metamaterial resonators provides clear advantages in different ways. The first evident improvement arises from the possibility of using a second metallic level to increase the length of the particles without expanding their area. On the other hand, the two metal layers and the vias provide further flexibility and the possibility to enhance both interring or selfcoupling, specially by etching the metallic strips of the elements face-to-face in the substrate (broad-side coupling). Figure 2(a) shows the layout and resonance frequencies of a subwavelength resonant particle based on the SRR, in which the inner and outer rings have been extended through vias to the bottom side of the substrate. The results of this figure show that the reduction ratio is around $R = 25\%$, which is higher than that obtained by considering SRRs with the conventional topology proposed by Pendry et al. (around 19%). We attribute this to the enhancement of edge coupling between the two individual resonators caused by their higher length.

In Figure 2(b), we have proceeded in a similar way, but this time trying to etch the individual resonators face-to-face. The reduction ratio is this time $R = 39\%$, which is attributed to broad-side coupling. Obviously, R can be further enhanced by simply using thinner substrates. Although the structures shown in Figures 2(a) and 2(b) are quite similar from the point of view of area, strip width, and even geometry, the results show that the subwavelength property becomes more intense in the structure of Figure 2(b).

At this point, we would like to mention that an important aspect of these particles for their use in the synthesis of metamaterials is their magnetic moment at resonance. To exhibit a strong magnetic behavior, it is important that the resonator topology preserves the cancellation of the different components of current distribution in the particle. Although the analysis of the polarizabilities of the proposed particles is complicated, the inspection of the current distributions inferred from electromagnetic solvers can lead us to useful information concerning the suitability of the particle for its use in the synthesis of metamaterials. This aspect will be considered in Section 5. In the next section, we deal with another relevant aspect which is intimately related to particle losses, that is, the quality factor.

4. LOADED AND UNLOADED QUALITY FACTOR

In the most general definition, the quality factor is the ratio between the total stored energy and the dissipated energy at resonance. This important figure establishes an upper limit for the performance of microwave components based on electromagnetic resonators, and the quality factor is the usual way to have into

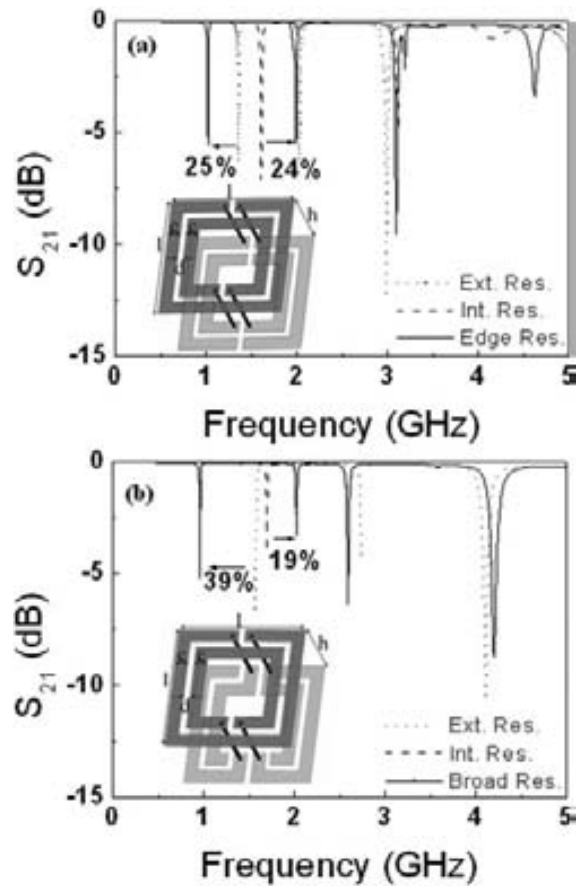


Figure 2 Simulated insertion losses of a microstrip line loaded with the indicated resonators and with the individual resonators (external and internal) forming the overall resonator. In (a) interresonators coupling is edge type; in (b) interresonator's coupling is both edge and broadside type. The values of the parameters are: $l = 8$ mm, $c = 0.8$ mm, $d = 0.4$ mm, $h = 1.27$ mm in a $\epsilon_r = 10.2$ Rogers RO3010 substrate

account the effects of losses. The loaded quality factor (Q_L) can be estimated from the following expression:

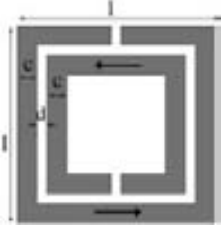
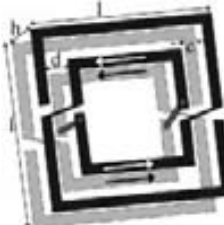


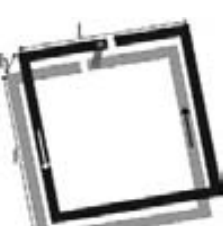
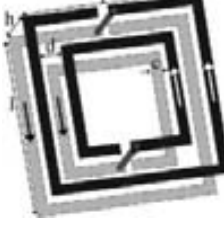
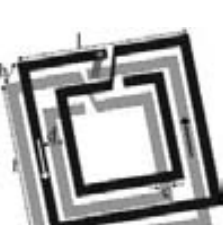
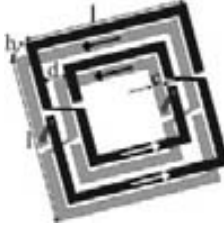
$$Q_L = \frac{f_0}{\Delta f_{-3\text{dB}}} \quad (2)$$

where f_0 is the resonance frequency and $\Delta f_{-3\text{dB}}$ is 3 dB the bandwidth. The unloaded quality factor of the resonator can be obtained from Q_L according to:

$$Q_u = \frac{Q_L}{1 - |S_{21}|} \quad (3)$$

where S_{21} is the magnitude of the transmission coefficient at resonance. The unloaded quality factor, which is the actual figure of merit of the resonators, can be severely influenced by device miniaturization. The reason is that the quality factor depends inversely on the current density in the resonator at the resonance frequency and therefore, since the miniaturization implies in general an increment of the current density, a certain level of degra-

TABLE 1 3D Representations of the New Proposed Metamaterial Resonators and Relevant Characteristics for the Two Considered Substrate Thickness

	3D DIAGRAM	$h = 1.27$ mm	$h = 0.254$ mm		3D DIAGRAM	$h = 1.27$ mm	$h = 0.254$ mm
1		$f_0 = 0.80$ GHz $\lambda_g/9.5$ $R = 30.2\%$ $M = 0.59$ $Q_u = 150$	$f_0 = 1.02$ GHz $\lambda_g/7.6$ $R = 35.0\%$ $M = 0.62$ $Q_u = 161$	5		$f_0 = 0.35$ GHz $\lambda_g/21$ $R = 21.2\%$ $M = 0.50$ $Q_u = 114$	$f_0 = 0.18$ GHz $\lambda_g/43$ $R = 61.6\%$ $M = 0.87$ $Q_u = 118$
2		$f_0 = 0.85$ GHz $\lambda_g/9$ $R = 25.6\%$ $M = 0.45$ $Q_u = 129$	$f_0 = 0.49$ GHz $\lambda_g/16$ $R = 68.9\%$ $M = 0.95$ $Q_u = 129$	6		$f_0 = 0.84$ GHz $\lambda_g/9$ $R = 24.6\%$ $M = 0.45$ $Q_u = 136$	$f_0 = 0.49$ GHz $\lambda_g/15$ $R = 66.5\%$ $M = 0.85$ $Q_u = 113$
3		$f_0 = 0.43$ GHz $\lambda_g/18$ $Q_u = 96$	$f_0 = 0.25$ GHz $\lambda_g/32$ $Q_u = 94$	7		$f_0 = 0.31$ GHz $\lambda_g/24$ $R = 26.4\%$ $M = 0.64$ $Q_u = 130$	$f_0 = 0.20$ GHz $\lambda_g/39$ $R = 17.6\%$ $M = 0.50$ $Q_u = 120$
4		$f_0 = 0.18$ GHz $\lambda_g/41$ $Q_u = 106$	$f_0 = 0.09$ GHz $\lambda_g/83$ $Q_u = 95$	8		$f_0 = 0.34$ GHz $\lambda_g/22$ $R = 23.7\%$ $M = 0.61$ $Q_u = 135$	$f_0 = 0.32$ GHz $\lambda_g/24$ $R = 30.6\%$ $M = 0.33$ $Q_u = 143$

dation in Q_u can also be expected. However, it is not expected that the proposed strategies for miniaturization lead to a significant increment of the current densities at resonance. Thus, we expect that the quality factor of the proposed resonant particles is not severely degraded. This will be corroborated in the next section.

5. COMPARATIVE ANALYSIS OF DIFFERENT SUBWAVELENGTH RESONATORS

In Table 1, different resonant particles that use two metal levels and vias are depicted, and their characteristics (R , M , and Q_u) are compared. For comparison purposes, the well known SRR topology proposed by Pendry et al. [3] (first row) and the broad-side coupled SRR (BC-SRR) proposed by Marqués et al. [22] (second row) are also included. The first column depicts the structure of the resonator in three dimensions. The arrows indicate the direction of current flow at the different metallic strips (black arrows for the bottom metallic layer and white arrows for the top metallic layer) at a certain given time under resonance. The representation of the

current density distribution (obtained by means of Agilent Momentum) is interesting because it gives an idea of the magnetic behavior of the particles at resonance. No current cancellation is observed in these particles. Therefore these are of interest for the synthesis of metamaterials and, in general, in those applications requiring magnetically coupled resonators. In fact, the different topologies shown in the table have been designed with the aim to avoid the destructive interferences between the magnetic fields produced by the currents in the strips at both sides of the substrate. All the simulations have been performed by considering the Rogers RO3010 substrate with $\epsilon_r = 10.2$, and two different thicknesses: $h = 0.254$ and 1.27 mm. In all the layouts the values of the indicated geometric parameters are $l = 15.28$ mm, $d = 0.2$ mm, and $c = 0.6$ mm. In the second and third columns are depicted the first resonance frequency (renamed f_0), R , M , Q_u and the size of the particle (side) expressed in terms of the guided wavelength, for the two considered substrate thicknesses. Q_u has been obtained from

electromagnetic simulations (by considering ohmic and dielectric losses) through the use of expressions (2) and (3), where the particles have been coupled to two microstrip lines (input and output lines) with via shorts in the extremes, and the particle being placed in between at a fixed distance from the extremes of each line. With this set-up the coupling between the particles and the lines varies (the particles are not necessarily located in those positions where the current in the line is a maximum) and this may produce certain influence on the value of Q_u . To properly compare the Q_u of the different particles, the same coupling is convenient, but this is difficult to achieve in practice.

The first and the second rows of Table 1 are dedicated to the conventional SRR and the BC-SRR. It is interesting to observe the significant reduction in the electrical size for the thinnest substrate in the case of the BC-SRR. This is obviously explained by the enhancement of coupling that can be achieved between metals etched face to face in a narrow substrate. This size reduction is also visible in other particles of Table 1 where broadside coupling is also present.

Particles 3 and 4 are significantly different from the other particles presented in the table since their subwavelength property does not arise from the interaction between two resonators, but from selfcoupling. The structure of these particles is based on the SR, which has been extended through vias to the bottom side of the substrate in such a way that the magnetic field created by the current flow in the strips does not cancel at resonance. The difference between the results obtained for $h = 1.27$ and 0.254 mm points out the importance of the broad-side capacitance between the strips etched at the two metallic levels. Obviously, for these particles we have not represented M and R , since these magnitudes

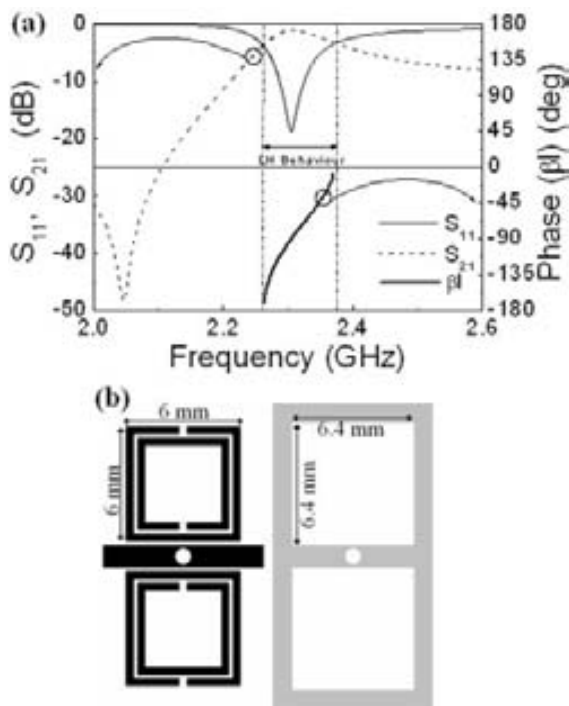


Figure 3 Left-handed cell implemented by using SRRs with open windows in the ground plane. (a) S-parameters and β ; (b) top and bottom view of the layout

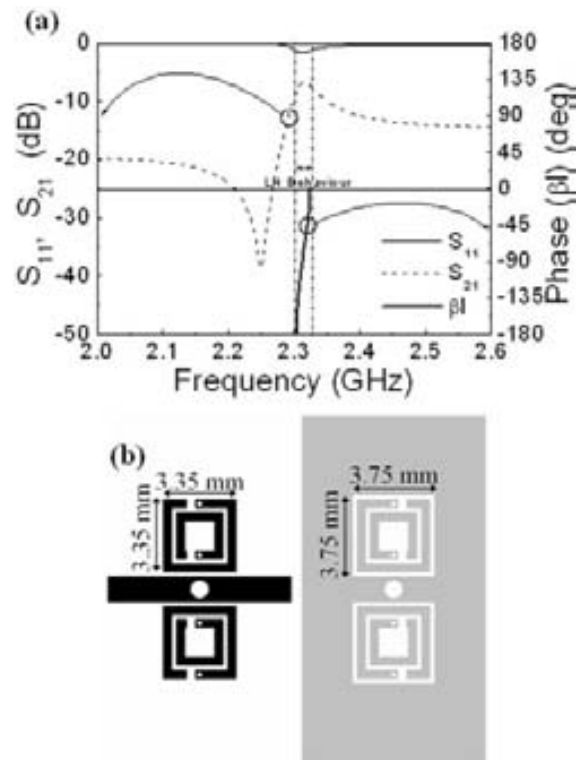


Figure 4 Left-handed cell implemented by using the particle number 7 of Table 1 with open windows in the ground plain. (a) S-parameters and β ; (b) top and bottom view of the layout

have significance only if the particle is composed of two individual resonators.

The particles 5 and 6 are composed of two coupled resonators. In both cases the geometry has been specially designed to magnify the broadside coupling. The increment in the coupling coefficient that is obtained by reducing the substrate thickness varies from $M = 0.50$ to 0.87 for the Particle 5 and from $M = 0.45$ to 0.85 for the Particle 6. This significant dependence of the coupling coefficient with the substrate thickness proves that in both cases the coupling is mainly governed by the broadside capacitance. This is obvious in Particle 6, but it is not evident in Particle 5, where different metallic strips share the same metal level and are very close. On the other hand, the presence of face-to-face metallic strips does not necessarily imply the existence of significant broadside coupling. Particle 8 is a clear example of this situation. Although the geometry of this particle contains metallic strips aligned at both sides of the substrate, it forces the current to close through an edge capacitance at the first resonance. Thus, by the effect of the geometry, the edge capacitance becomes the dominant coupling mechanism at the first resonance. As a consequence, the reduction of substrate width does not substantially decrease the electrical size of the particle.

The different examples collected in Table 1 marshal a representative sample of the miniaturization possibilities of metamaterial resonators by the inclusion of vias and the use of a second metallic layer. Concerning Q_u , it can be appreciated that the SRR provides the best result. However, we do not obtain significant degradation in this figure for the other resonator topologies. Nevertheless, we must mention that the technique to obtain Q_u is not

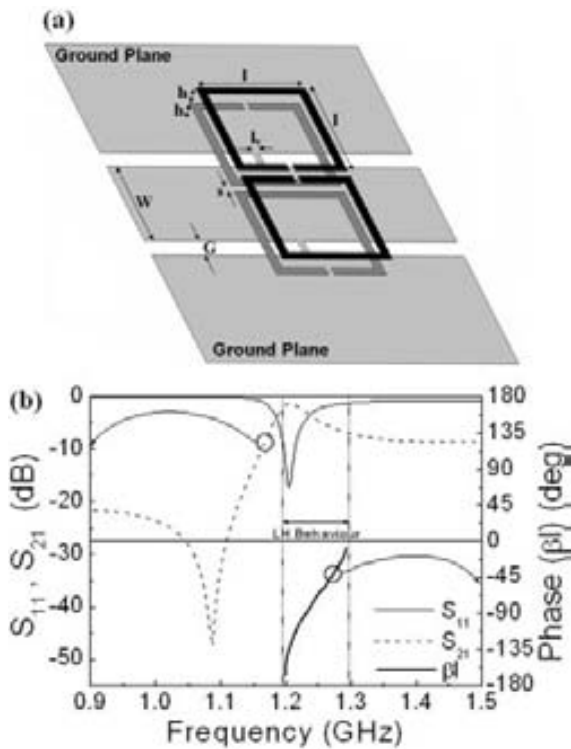


Figure 5 Left-handed cell implemented by using BC-SRRs (particle number 2 of Table 1) coupled to a CPW with shunt inductances. (a) 3D representation; (b) S-parameters and βl

very accurate due to the strong sensitivity of Q_u with the modulus of the transmission coefficient at resonance (Expression 3) and to the different coupling between the resonators and the input and output lines (as has been mentioned before). Hence, the results provided in the table are estimative, rather than exact values of Q_u .

6. ILLUSTRATIVE EXAMPLES OF APPLICATION TO THE DESIGN OF LEFT HANDED STRUCTURES

The proposed structures can replace the conventional SRRs in many applications, like, for instance, the design of left handed transmission lines. In Figures 3 and 4, the frequency responses, including the dispersion relation, of one-cell left handed microstrip structures based on the conventional SRR (see Fig. 3) and on the resonator labeled 7 in Table 1 (see Fig. 4), are depicted (the characteristics of the Rogers RO3010 substrate have been considered, with dielectric constant $\epsilon_r = 10.2$ and thickness $h = 1.27$ mm). The βl diagram, which has been inferred from the phase of the transmission coefficient, ϕ_T , according to

$$\cos(\beta l) = \frac{\cos(\phi_T)}{|S_{21}|} \quad (4)$$

where $|S_{21}|$ is the magnitude of the transmission coefficient, indicates that the phase and group velocities are antiparallel in the region of interest (to derive Expression 4, it has been assumed that the attenuation constant is null in the transmission band). The central frequency of the left handed band is centered in the vicinity of 2.3 GHz in both cases. However the structure of Figure 4 is much smaller for the reasons that have been explained before. The

significant insertion losses in Figure 4 are attributed to the smaller mutual inductance between the line and the particle. It is expected that particle excitation is enhanced by using coplanar waveguide (CPW) configurations. These CPW-based cells, which were proposed previously [23], consist on a CPW line with thin metal wires that connect the central strip to the ground planes (acting as shunt inductances) and two coupled resonators. A three-dimensional (3D) representation of this structure, using BC-SRRs (Particle 2 in Table 1) and its frequency response and dispersion are depicted in Figure 5. The dimensions for the CPW and the shunt inductances are as follows: CPW line width $W = 7$ mm, slots width $G = 1.35$ mm, and shunt inductances width $l_v = 0.4$ mm. The geometrical dimensions for the resonators are: side of the outer ring $l = 7.8$ mm, strip width $c = 0.6$ mm, and separation between resonators $s = 0.55$ mm. Simulations have been performed using the Agilent Momentum simulator. The considered substrate has a dielectric constant of $\epsilon_r = 10.2$ and the three metal levels are separated a distance of 0.635 mm. The characteristics of the frequency response and dispersion of the structure of Figure 5 are similar to those of Figures 3 and 4.

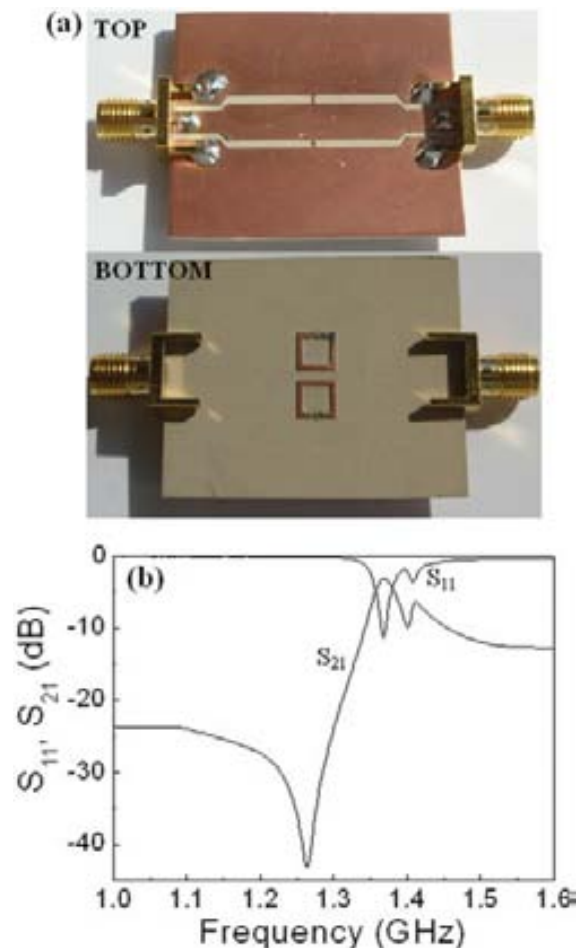


Figure 6 Fabricated prototype of a left handed cell corresponding to a CPW loaded with resonators labeled as number 3 in Table 1 (a), and measured frequency response (b). [Color figure can be viewed in the online issue, which is available at www.interscience.wiley.com]

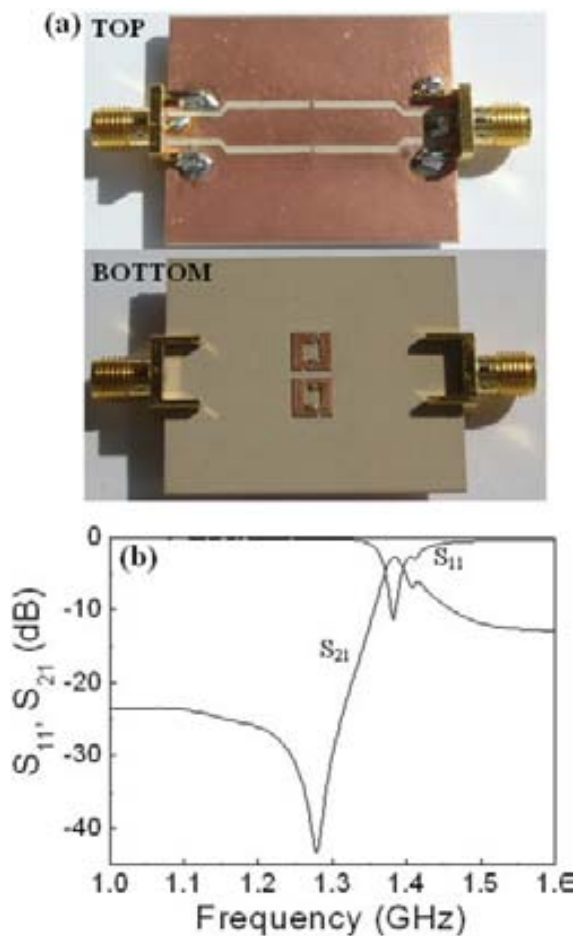


Figure 7 Fabricated prototype of a left-handed cell corresponding to a CPW loaded with resonators labeled as number 5 in Table 1 (a), and measured frequency response (b). [Color figure can be viewed in the online issue, which is available at www.interscience.wiley.com]

Two additional left handed cells in CPW technology have been designed and fabricated, but using two of the new resonators. One of them contains the resonator number 3 (see Fig. 6) and the other one contains the resonator number 5 (see Fig. 7). The measured frequency responses (obtained by means of the Agilent HP8720ET vector network analyzer) are also depicted in these figures. The left handed cells have been fabricated using the Rogers RO3010 substrate with relative permittivity $\epsilon_r = 10.2$ and thickness $h = 0.635$ mm. Actually, we have used two different substrates. In one of them we have etched the rings (at both sides) and we have drilled the holes for vias metallization. In the other, the CPW structure and the inductive wires have been etched. After aligning both substrates, they have been soldered and subjected to pressure to minimize the presence of air in the contacting region. CPW dimensions are $W = 5$ mm, $G = 1.16$ mm and the wire inductance has a width of $l_v = 0.4$ mm in both cells. The geometrical parameters of the resonator number 3 are as follows: the side length is $l = 4.8$ mm, the strip width $c = 0.6$ mm and separation between the resonators $s = 1.36$ mm. For the cell with the resonators number 5 the dimensions are $l = 5$ mm, $c = 0.6$ mm, the separation between adjacent strips $d = 0.2$ mm and $s = 1.16$

mm. In view of the measured frequency responses of these left handed cells, insertion losses in the allowed band are within acceptable limits (a significant improvement as compared to Figure 4 has been obtained by using a CPW configuration). The small dimensions of the resonators make the cell to be electrically very small. The cell size is actually given by the side length of the resonators, and this is as small as $\lambda_g/24$ and $\lambda_g/23$ for the resonators of Figures 6 and 7, respectively. In future works, the application of these miniaturized left handed cells to the design of microwave components is expected. Their application to the design of reflectors is another potential application.

7. CONCLUSIONS

As a conclusion, the use of two metal levels connected by vias in the design of subwavelength resonant particles provides the possibility to introduce broadside coupling in the designs and hence to enhance the miniaturization. The geometry and the topology of the particles play a very important role in size reduction since the mere presence of face-to-face metal strips does not guarantee broadside coupling. Several merit figures have been pointed out as numerical indicators of particle performance. The analysis of the proposed particles shows that by enhancing the coupling between the constituent elements of the resonators or self coupling, it is possible to enhance miniaturization without severe quality factor degradation. Some of the proposed resonant particles have been used to design several compact left handed cells in microstrip and CPW technology. The frequency responses are indicative of better performance in CPW-loaded structures. Work is in progress to apply the proposed particles to the design of planar passive microwave devices, where miniaturization is the key point.

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MINIATURIZED MICROSTRIP CROSS-COUPLED BANDPASS FILTER USING NOVEL STEPPED IMPEDANCE RESONATORS WITH A DESIRABLE UPPER STOPBAND

Guohua Zhao and Changjun Liu

School of Electronics and Information Engineering, Sichuan University, Chengdu 610064, China; Corresponding author: cjliu@ieee.org

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ABSTRACT: A miniaturized microstrip cross-coupling bandpass filter using new stepped impedance resonators (SIRs) with simultaneous suppression of the 2nd harmonic passband response is presented. Four improved and miniaturized SIRs constitute the compact filter, which exhibits a sharp transition band due to two transmission zeros at both sides

of the passband. Two quarter-wavelength open-ended stub resonators are applied to provide extra transmission zeros for suppressing the 2nd harmonic passband. A bandpass filter with a center frequency at 2.5 GHz was designed and fabricated, of which experimental results validated the proposed filter design. © 2008 Wiley Periodicals, Inc. *Microwave Opt Technol Lett* 50: 1270–1273, 2008; Published online in Wiley InterScience (www.interscience.wiley.com). DOI 10.1002/mop.23343

Key words: microstrip bandpass filter; cross-coupling; harmonic suppression; miniaturization; stepped-impedance resonator

1. INTRODUCTION

In modern wireless communication systems, high-selectivity small-size microstrip bandpass filters with excellent out-of-band rejections are required to enhance the system performance and reduce the fabrication cost. To obtain a sharp transition band, one effective method is to introduce two transmission zeros and locate them at either side of the passband, respectively. Cross-coupled configurations, which exhibit elliptic function-like responses [1], are suitable and commonly applied.

Conventional microstrip bandpass filters are designed with half-wavelength or quarter-wavelength resonators. To reduce filter size, cross-coupled filters using folded half-wavelength hairpin resonators were developed [2]. Stepped impedance resonators have attracted much attention due to their controllable first spurious passband [3, 4]. With modified stepped impedance hairpin resonators, those microstrip filters were more compact and had better stopband extensions [5].

The planar bandpass filters made of resonators inherently have the spurious passbands at multiple of the center frequency, which limit the upper stopbands. Many methods have been proposed to solve the problem, such as providing different electric length for even and odd modes [6], using uniplanar compact electric band-gap (EBG) structure [7] or defected ground structure (DGS) [8], creating transmission zeros by open-ended stub resonators [9], and so on.

In this letter, we propose a novel miniaturized microstrip filter structure that is suitable for realizing the high selectivity and suppressing the 2nd harmonic passband. In the design, four improved folded half-wavelength hairpin SIRs are placed as a two-by-two array to achieve the cross-coupling. Therefore, the cross-coupled capacitance is introduced directly between two adjacent resonators with two transmission zeros introduced. In addition, two quarter-wavelength open-ended stubs are attached at the edges of the resonators in the main-coupling path. One extra transmission zero is introduced at the first harmonic frequency. A compact high-selectivity bandpass filter at 2.5 GHz with harmonic suppression is optimally designed, fabricated, and measured.

2. RESONATOR AND FILTER DESIGN

Figure 1 shows layout of the proposed filter. Four symmetric basic resonators are placed oppositely to each other and form a four-square contour. Tapped-line input/output ports are connected to two resonators, in which the distance between the feed point and the resonator center is L_c . The resonator consists of two triangular patches connected to both ends of a high impedance microstrip line section, which is bent three times to reduce circuit area. The resonator is an improved variation of the SIR with square-shape as well. The main coupling principle is identical to the hairpin SIR's. The SIR section is used in each element, of which the size is adjustable with the relevant impedance ratio [10].

Figure 2 shows the equivalent circuit of the proposed filter. In each resonator, the distributed inductance L_1 and L_2 are formed by

Artículo C

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Resumen del artículo:

Este artículo se centra en el comportamiento físico y las posibles aplicaciones de un nuevo tipo de partícula plana resonante: el *open complementary split ring resonator* (OCSRRs). Estos resonadores complementarios son homólogos de los *open split ring resonator* (OSRRs), ya presentado por algunos de los autores, y consisten en un par de “ganchos” concéntricos grabados en una capa de metal y opuestamente orientados. Al igual que los OSRRs, los OCSRRs son resonadores abiertos que pueden ser excitados por medio de una fuente de voltaje o corriente. Se propone un modelo circuital de la partícula, que es validado experimentalmente mediante la excitación de las partículas por medio de una línea de transmisión coplanar (CPW). También se muestra que los OCSRRs presentan frecuencias de resonancia de orden superior, que pueden ser suprimidas selectivamente mediante la introducción de elementos adicionales, llevando a cabo un simple análisis basado en paridad de modos. Debido al pequeño tamaño eléctrico de los OCSRRs, tales partículas son útiles para la síntesis de metamateriales planares y componentes de microondas.

Open complementary split ring resonators (OCSRRs): modelling, analysis and prospective applications

F Aznar, A Vélez, M Durán-Sindreu, J Bonache and F Martín

GEMMA/CIMITEC, Departament d'Enginyeria Electrònica.
Universitat Autònoma de Barcelona. 08193 BELLATERRA (Barcelona). Spain.
E-mail: Ferran.Martin@uab.es

Abstract

This paper is focussed on the physics and prospective applications of a new type of planar resonant particles: the open complementary split ring resonators (OCSRRs). These resonators are the complementary counterpart of the open split ring resonators (OSRRs), previously presented by some of the authors, and consist on a pair of concentric hooks etched on a metal layer in opposite orientation. Like OSRRs, OCSRRs are open resonators that can be excited by means of a voltage or current source. An accurate circuit model of the particle is proposed and experimentally validated by exciting the particle by means of a coplanar waveguide (CPW) transmission line. It will be also shown that OCSRRs exhibit higher order resonance frequencies, which can be selectively suppressed by introducing additional elements, as derived from a simple analysis based on mode parity. Due to the small electrical size of OCSRRs, such particles are useful for the synthesis of planar metamaterials and microwave components.

1. Introduction

Since the proposal of the split ring resonator (SRR – figure 1(a)) as an electrically small non-magnetic resonant particle useful for the synthesis of negative permeability media [1], and the subsequent application of this particle to the implementation of the first artificial material exhibiting left handed wave propagation [2], there has been an intensive research activity in the field of metamaterials. Several recently published textbooks have dealt with this hot topic [3-8]. Metamaterials are artificial structures composed of small size inclusions (or “atoms”) based on combinations of metals and dielectrics. As long as the unit cell of these structures is made small as compared to the guided wavelength, effective (or continuous) media properties arise, and their electromagnetic (or optical) behaviour depends on how these unit cells are structured, rather than on their composition. Thanks to these characteristics, unusual properties can be achieved, including left handed wave propagation, super-resolution, negative refractive index or cloaking, among others.

There are several approaches for the synthesis of metamaterials. One of these approaches is based on the use of resonant particles like the SRR, or other electrically small resonators (spiral resonators [9-11], broad-side coupled split ring resonators [12], chiral resonators [13], or other resonant structures with more complex topology [14]). These resonant elements have been used for the synthesis of bulk 1D, 2D and 3D metamaterials, and also for the synthesis of planar metamaterial structures, such as frequency selective surfaces [15] or metamaterial transmission lines [16]. In 2004, it

was presented the complementary split ring resonator (CSRR) [17], which has been demonstrated to be a very interesting particle for the implementation of composite right/left handed transmission lines in microstrip technology. The CSRR is the complementary counterpart of the SRR and hence it consists on a pair of slot rings with apertures in opposite orientation, etched on a metallic screen (figure 1(b)). The CSRR can be excited by means of an axial time varying electric field, and, etched in the ground plane of a microstrip line, it provides a negative effective permittivity to the structure, as has been previously discussed [17,18] (other topologies inspired on the CSRR topology have also been used for the synthesis of one-dimensional negative permeability media [19]).

In figure 1 are also depicted the equivalent circuit models of these resonant particles (the SRR and the CSRR) [20]. The SRR is a closed resonant tank, where the inductance L_s is the inductance of a ring of average radius and identical width as the individual rings forming the particle, and the capacitance is $C_s=C_o/4$, C_o being the distributed (edge) capacitance between the two concentric rings. Namely, the capacitance is given by the series connection of the capacitances corresponding to the upper and lower halves of the particle [20,21]. The CSRR has been also modelled as a closed resonator. The capacitance C_c is the capacitance of a disk of radius $r_o-c/2$ surrounded by a metallic plane at a distance c of its edge, and the inductance is the parallel combination of the two inductances connecting the inner disk to the ground, as reported in [20]. In the absence of a dielectric slab and for perfectly conducting and infinitely thin metallic layers, the SRR and the CSRR are strictly dual particles, their circuit parameters are related by $C_c=4(\epsilon_o/\mu_o)L_s$ and $C_o=4(\epsilon_o/\mu_o)L_o$, and their resonance frequencies are identical [20]. Under real conditions (presence of a dielectric substrate and metal layers with finite conductivity and thickness), some deviations from duality (mainly due to the presence of the substrate) are expected, and the resonance frequencies of SRRs and CSRRs with identical dimensions are very close, but not identical.

In 2004, it was also presented the open split ring resonator (OSRR) [22]. This is the open version of the SRR (see the topology of this particle and its equivalent circuit in figure 1(c)). Thus, it is an open resonator, which can be modelled by means of a series LC circuit, where the inductance is identical to that of the SRR, and the capacitance is C_o , i.e. the distributed (edge) capacitance between the two concentric rings along the whole circumference. According to this, it follows that for identical dimensions and substrate, the resonance frequency of the OSRR is half the resonance frequency of the SRR.

This paper is focussed on a novel resonator, which is the missing particle in the list of figure 1: the open complementary split ring resonator (OCSRR). Section 2 is devoted to the modelling of this particle; in section 3, this model is validated through experiment; in section 4, we discuss the deviations from the prediction of the OCSRR model at high frequencies; finally, prospective applications and the main conclusions are highlighted in section 5 and 6, respectively.

2. Circuit model of the OCSRR

The OCSRR is obtained from the SRR topology by opening it (as in the OSRR) and applying duality (as in the CSRR). It is thus the complementary counterpart of the OSRR, and for this reason it has been called OCSRR (see a typical layout in figure 2). However, the term “open” in the nomenclature of this particle must be considered with some caution. “Open resonator” means that the particle exhibits two metallic terminals

for excitation (through a voltage or current source), as opposed to closed particles, where these terminals are absent and the particle is magnetically or electrically excited. Thus the nomenclature of the OSRR is very clear on account of its topology (the metallic terminals –ports– are indicated in figure 1(c)). In the case of the OCSRR, what does the term “open” mean? In other words: where are the metallic connecting terminals? Such terminals are the metal regions below and above the particle, as indicated in figure 2. Between the upper and lower terminals, there is an electric short through the metal between the inner and outer slot rings forming the particle, but there is also capacitive connection through the capacitances across the slots. Thus, according to this, the circuit model of the particle is an open parallel resonant tank, as figure 2(b) illustrates. Indeed, from the previous analysis on SRRs [21], OSRRs [22] and CSRRs [20] (summarized in section 1), it follows that the inductance is L_o , that is, the inductance of the metallic strip between the ring slots, and the capacitance is C_c , i.e. the same as the capacitance of the CSRR. Therefore, it is expected that the resonance frequency of the OCSRR is one half the resonance frequency of the CSRR, although in practice some deviation is expected due to the presence of the metal region between the inner metal of the slot rings and the lower electrode. Obviously, the model of the OCSRR is valid in a limited range of the spectrum (as occurs in SRRs, CSRR, etc). This aspect will be discussed in section 4. Concerning the calculation of L_o , this electrical parameter is given by the inductance of the central strip of the OCSRR surrounded by the inner and outer metal of the particle, as if these metallic regions were the ground planes of a circular CPW structure. Thus, L_o can be calculated from well known formulas giving the per-unit length inductance of a CPW structure without backside metal [23]. C_c can be obtained from a variational method reported in [20].

3. Validation of the model and results

To experimentally determine the resonance frequency of the OCSRR, the particle can be excited by using a coplanar waveguide (CPW) configuration. The topology of the test structure is shown in figure 3. Two symmetrically placed OCSRRs have been etched in the slots of the CPW. To avoid the excitation of parasitic modes (slot mode) the different ground plane regions have been connected through vias and metallic strips etched in the back side of the substrate, as depicted in figure 3.

According to the OCSRR model presented in section 2, the test structure can be modeled by cascading a pair of shunt resonant tanks between two transmission lines (the latter modeling the CPW sections of the structure), as figure 3(b) illustrates. By designing the host CPW with a characteristic impedance of $Z_o=50\Omega$, the transmission coefficient between the input and the output port must be maximum at the resonance frequency of the OCSRR (provided that the impedance of the ports is also 50Ω , as usual). The reason is that the shunt branch opens at this frequency, and the injected power must be totally transmitted to the load (except certain power that may be lost due to conductor and dielectric losses). Thus, the resonance frequency can be determined from the simulated (by means of electromagnetic solvers) or measured transmission, S_{21} , or reflection, S_{11} , coefficients of the test structure.

The simulated (by using the *Agilent Momentum* electromagnetic software) and measured (by using the *Agilent 8720ET* vector network analyzer) frequency responses of the test structure shown in figure 3 are depicted in figure 4(a). For the measurement, the standard Open/Short/Load/Through calibration technique employing the Agilent 85052D calibration kit has been applied. It is also depicted in this figure the circuit simulation that has been obtained by extracting the parameters of the LC resonant tank.

Losses have been excluded and the parameters have been inferred from the electromagnetic simulation by curve fitting. Actually this is very simple since the resonance frequency of the OCSRR is given by the notch in the reflection coefficient. Thus we have only one degree of freedom and the determination of the inductance and capacitance of the OCSRR from curve fitting is straightforward. The circuit simulation is in good agreement with the electromagnetic simulation and measurement. We have also obtained the resonance frequency of a CSRR with identical dimensions, that has been etched in the ground plane of a microstrip line implemented on the same substrate, that is, the *Rogers RO3010* with dielectric constant $\epsilon_r=10.2$, dielectric thickness $h=1.27\text{mm}$ and metal thickness $t=35\mu\text{m}$ (see figure 4(b)). The resonance frequency of the CSRR has been inferred according to the method reported in [24], where it is also indicated the configuration of the structure. For the OCSRR, the resonance frequency obtained from measurement and simulation are, $f_0^{exp}=0.95\text{GHz}$ and $f_0^{sim}=1.01\text{GHz}$, respectively, and the extracted inductance and capacitance of the particle are $L=10.41\text{nH}$ and $C=2.33\text{pF}$. For the CSRR, $f_0^{exp}=1.96\text{GHz}$, $f_0^{sim}=2.014\text{GHz}$, $L=2.28\text{nH}$ and $C=2.74\text{pF}$. These results are in agreement with the model of the OCSRR, from which it follows that the resonance frequency of this particle is half the resonance frequency of the CSRR (with identical dimensions), the capacitances are very similar, and the ratio of inductances is close to 4.

We have considered additional OCSRRs and CSRRs with different radius (and $c=d=0.2\text{mm}$), and we have determined the resonance frequency for all of them. The results (see table 1) show that by increasing the radius of the particles, the ratio between the resonance frequency of the CSRR and the OCSRR is closer to two. This is an expected result since the effect of the inter-metallic region between the central strip and the metal inside the OCSRR vanishes for sufficiently large particles. Notice also that for the first row ($r_{ext}=7\text{mm}$), the ratios between the resonance frequencies and the reactive parameters of the particles are very close to the theoretical values. We have also analyzed the effects of varying c and d , leaving unaltered the external radius of the particle. The results are depicted in table 2. It is also clear according to this table that as c and d decrease, the model predictions are more accurate.

Table 1. Resonance frequency for the OCSRR and CSRR with the indicated external radius. In all the cases $c=0.2\text{mm}$ and $d=0.2\text{mm}$.

r_{ext} (mm)	f_{OCSRR} (GHz)	C_{OCSRR} (pF)	L_{OCSRR} (nH)	f_{CSRR} (GHz)	C_{CSRR} (pF)	L_{CSRR} (nH)
7	0.54	4.92	18.28	1.07	4.92	4.50
5	0.78	3.41	12.21	1.55	3.68	2.87
3	1.33	2.08	7.05	2.75	2.84	1.18

Table 2. Resonance frequency for the OCSRR and CSRR with the indicated values of c and d . In all cases $r_{ext}=7\text{mm}$.

$c=d$ (mm)	f_{OCSRR} (GHz)	C_{OCSRR} (pF)	L_{OCSRR} (nH)	f_{CSRR} (GHz)	C_{CSRR} (pF)	L_{CSRR} (nH)
0.3	0.58	4.14	17.36	1.16	4.09	4.64
0.4	0.65	3.69	16.72	1.23	3.98	4.23
0.5	0.69	3.42	15.72	1.30	3.62	4.22
0.6	0.74	3.14	15.02	1.35	3.44	4.04

The results shown in this section validate the reported model of the new proposed particle. In the vicinity of the resonance frequency of the OCSRR, the particle

is accurately described by means of an open parallel resonant tank, with an inductance four times larger than that of the CSRR, and identical capacitance. In the next section deviations from the predictions of the model at higher frequencies are analyzed and discussed.

4. Discussion

The description of the OCSRR (and the other related particles such as the SRR, CSRR and OSRR) by means of a simple lumped element model is restricted to a limited range of frequencies. The reason is that at high frequencies, the particles are not electrically small. In general, for metamaterial and circuit design, a proper description of the particles in the vicinity of their first resonance suffices. However, some deviations from the predictions of the model at moderate frequencies are present, and it is of interest to analyze their origin. This is the main aim of this section. In figure 5, we depict the frequency response of the structure shown in figure 3(a), inferred from electromagnetic simulation, and also the circuit simulation that has been inferred from the circuit of figure 3(b) with the extracted parameters of the OCSRR. The frequency range has been extended to roughly 5 times the first resonance frequency of the OCSRR, in order to clearly appreciate the moderate to high frequency effects. As in figure 4, there is a deep notch in the transmission coefficient (transmission zero), located at 2.86GHz, preceded by a very narrow spike (at 2.42GHz). These effects are not predicted by the model of figure 3(b). The transmission zero present at 2.86GHz is related to the fact that the metallic region present between the central strip of the CPW and the inner metallic region of the OCSRR introduces an extra inductance to the model. This inductance does not appreciably affect the frequency response in the vicinity of the resonance frequency of the OCSRR (at roughly 1GHz), but is responsible for the transmission zero present at a higher frequency. Such transmission zero can be accounted for by introducing an additional inductance to the model, as depicted in figure 6(a). At the frequency where the shunt branch shorts, namely:

$$f_z = \frac{1}{2\pi} \sqrt{\frac{L_i + L_0}{L_i C_C L_0}} \quad (1)$$

signal propagation is precluded, and the injected power is reflected back to the source (transmission zero frequency). By taking into account the effects of such inductance, the transmission zero frequency is predicted by the circuit model, as shown in figure 5.

The spike present at 2.42GHz is due to the second resonance frequency of the OCSRR, but its effects are difficult to model in practice. At roughly 3.83GHz, there is an additional transmission band, which is related to the third resonance frequency of the OCSRR, and can be accounted for by the model simply by introducing an additional resonator, as shown in figure 6(b). The inductance and capacitance of this new resonator can be determined from the electromagnetic simulation of the structure as for the resonator describing the first resonance frequency of the OCSRR. The circuit simulation of the complete model is also depicted in figure 5. Except by the spike preceding the transmission zero, the frequency response obtained from circuit simulation is in good agreement with the electromagnetic simulation.

To get more insight on the characteristics of the OCSRR-loaded CPW transmission lines, we have modified the width and length of the metallic region present between the central strip of the CPW and the inner metallic region of the OCSRR. By

doing this, we modify the inductance series connected to the resonant tank modeling the OCSRR, with the result of a shift in the transmission zero frequency. This has been corroborated from the electromagnetic simulation of the structure with different dimensions, length (D) and width (X), for the above cited metallic region (figure 7). Notice that dimensions of the structure in figure 7 are different than those of figure 3.

We have also obtained the magnetic currents, provided by the *Agilent Momentum* commercial software, obtained at the relevant frequencies of the structure in figure 3 (figure 8). At 1.02GHz, the first resonance frequency of the OCSRR, the magnetic currents at both rings forming the OCSRRs are co-directional. This means that the excited mode of the OCSRR is an odd mode, which can be interpreted as a slot mode of the CPW structure constituted by the inter-rings region of the slot rings and the surrounding metallic regions. This corroborates the relation with the OSRR [22] when the Babinet principle is applied, just as in the SRR and the CSRR [18]. Notice that the maximum current (only one current maximum appears) takes place in the region between the rings and the central strip of the CPW. This is coherent with the maximum magnetic current at the first resonance frequency of a complementary spiral resonator (CSR), not shown, which takes place in the region connecting the inner and outer loop of the spiral. Notice also that in the region between the rings and the central strip of the CPW, the magnetic currents are contra-directional. Thus, in this region, the mode is even.

At 2.42GHz, where the spike appears, the magnetic currents in the rings are contra-directional, although the inner ring is scarcely excited. In the inter-connecting region between the rings and the central strip of the CPW, the currents are co-directional, which means that in this region the mode is odd. With the presence of the connecting strip (and vias) in the bottom side of the structure, this mode is minimized and this explains that the inner ring is scarcely excited. As can be seen in the current diagrams, at this frequency, power is not transmitted to the right hand side of the structure, and hence it is reflected back to the source, giving a notch in transmission.

Finally, at 3.83GHz (the third resonance frequency of the OCSRR), the magnetic currents in the rings exhibit three maximums: one in the external ring, another one in the internal ring, and the third one in the region between the rings and the central strip of the CPW. As occurs at the third resonance of the CSR, in the region where the current of one ring maximizes, the other ring exhibits nearly zero current. Nevertheless, as at the first resonance, the magnetic currents in the region between the rings and the CPW are contra-directional, which means that in this region the mode is even.

The parity of the different modes suggests that it is possible to selectively cancel them. The first resonance frequency can be merely cancelled by connecting the inner and outer metallic region of the OCSRR, by means of metallic strips etched in the back substrate side and metallic vias, as figure 9 illustrates (notice that in figure 9, the simulated structure is that of figure 7). Indeed, the third resonance frequency is also partially cancelled by introducing such additional elements (see figure 9(b)). Further cancellation of the second resonance is achieved by connecting the metallic parts present at both sides of the slots in the region between the CPW and the OCSRR (access region), as illustrated in figure 10. The reason is that the access region exhibits an odd mode at this frequency, as has been explained before. Thus, through electric connection of the external metallic regions, the mode is partially suppressed. This also prevents the slot mode of the CPW structure at the first resonance of the OCSRR (the resonance frequency of interest). For this reason, we are forced to use this electrical connection by introducing the back side metallic strip and vias. This is effective in suppressing the slot mode of the CPW structure. However, it has been found that to substantially suppress

the second resonance frequency of the OCSRR, it is necessary to introduce several strips in parallel or, alternatively, a wide strip. The responses with several strips are illustrated in figure 11, where it can be seen that the spike is practically suppressed. Obviously, it is possible to effectively and simultaneously suppress the second and third resonance by combining the two proposed strategies.

To summarize this discussion, it has been found that the phenomenology associated to OCSRRs coupled to CPW structures is very rich, with the presence of several resonances that can be selectively suppressed by the addition of additional elements (backside metallic strips and vias). The suppression of higher order resonances is of interest to improve the wideband behaviour of microwave components based on these resonant elements (for instance in filters). This aspect will be discussed in the next section.

5. Potential applications

Due to the small electrical size of OCSRRs (at the first resonance frequency), these particles may find applications in the design of microwave circuits where size reduction is mandatory. Such particles are open parallel resonators that can be applied to the design of microwave filters [25] and diplexers, among other microwave components. In [25], a wideband band pass filter based on OCSRRs was proposed. In this paper, it is demonstrated that OCSRRs can be series connected in microstrip technology to implement low pass filters with very sharp transition bands and wide rejection band. By series connecting the resonator to the line, signal transmission is inhibited in the vicinity of the resonance frequency, and a stop band is generated. By cascading OCSRRs with different dimensions, it is possible to enhance the rejection band with the result of a low pass structure with wide band rejection. A prototype device example is depicted in figure 12, and the simulated and measured frequency responses of this device are depicted in figure 13. The current diagrams obtained from electromagnetic simulation at the indicated frequencies (see figure 14) reveal which OCSRRs are responsible for signal filtering in the different regions. As long as OCSRRs are electrically small, the size of the resulting filter is small. Its length is only 0.77λ , where λ is the guided wavelength at the filter cutoff frequency.

OCSRRs are also of interest for the design of metamaterial transmission lines. Specifically, OCSRRs can be combined with OSRRs in order to implement composite right left handed transmission lines [26] in CPW (as has been already demonstrated by the authors [27]) or in microstrip technology. The potential of these artificial lines is very high since they are small and allow for further design flexibility, as compared to conventional lines, through dispersion and impedance engineering. To demonstrate the application possibilities of these composite right left handed lines based on the combination of OCSRRs and OSRRs, we present here an implementation in microstrip technology. The structure is depicted in figure 15 and consists on an artificial line exhibiting backward wave propagation at low frequencies, and forward wave propagation at high frequencies. The dimensions of the elements have been chosen so that there is a continuous transition between the backward and forward wave transmission regions, and the structure behaves like a wideband band pass filter. Again, filter dimensions are very small and filter performance is good (see figure 16). It is remarkable that the filter exhibits a good rejection above the passband of interest. This is in part due to the fact that the second resonance of the OCSRR is partly suppressed, so that the spurious related to such resonance is not appreciable in the measured filter characteristics. Such lines based on a combination of OCSRRs and OSRRs are being

applied at present by the authors to the design of other microwave filters and dual-band components based on metamaterial concepts.

6. Conclusions

In conclusion, it has been shown that OCSRRs are electrically small open parallel resonators that can be modelled by means of a parallel resonant tank where the elements are closely related to those of the CSRR. We have presented the circuit model of a CPW structure loaded with a pair of OCSRRs, and the model has been validated through simulation and experiment. It has been found that a transmission zero arises at higher frequencies. The origin of this transmission zero has been interpreted and accounted for in the model by introducing an extra inductance. We have also shown that OCSRRs exhibit higher order resonance frequencies, that correspond to even or odd modes of OCSRR excitation. From this analysis, we have inferred the strategy for resonance suppression (by means of properly allocated metallic strip and vias connecting the different ground plane regions), which has been validated through electromagnetic simulation. Finally, we have highlighted prospective applications of OCSRRs in microwave engineering. OCSRRs are the open version of CSRRs, in the same form that OSRRs are the open version of SRRs. With this new particle we enhance the design capabilities of microwave circuits and metamaterial-inspired structures based on resonant elements.

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Figure captions

Figure 1. Typical topology and equivalent circuit model of the SRR (a), CSRR (b) and OSRR (c).

Figure 2. Typical topology of the OCSRR (a) and equivalent circuit model (b). The two terminals –ports– of the open resonator are indicated.

Figure 3. Test structure for the determination of the resonance frequency of the OCSRRs, based on a CPW loaded with a pair of OCSRRs (a) and equivalent circuit (b). The host CPW has a slot and central strip widths of $G=1\text{mm}$ and $W=3.64\text{mm}$, respectively. The length of the structure is 10.4mm (excluding access lines). The dimensions of the OCSRRs are, in reference to figure 2, $c=0.6\text{mm}$, $d=0.6\text{mm}$ and $r_{ext}=5\text{mm}$. The structure has been implemented on the *Rogers RO3010* substrate with dielectric constant $\epsilon_r=10.2$, dielectric thickness $h=1.27\text{mm}$ (metal thickness is $t=35\mu\text{m}$). The vias (connecting the upper and lower metallic layers) and the back substrate side metallization are indicated in black. The backside metal consists thus on a rectangle with a strip width of 0.2mm .

Figure 4. Simulated and measured transmission (S_{21}) and reflection (S_{11}) coefficients for the structure shown in figure 3 (a), and a microstrip line loaded with a CSRR (b). In both cases a 50Ω host line has been considered, and the resonators are identical. The circuit simulations obtained by extracting the circuit elements from the electromagnetic simulation are also depicted. Notice that the resonance frequency of the CSRR is neither given by the reflection zero nor given by the transmission zero. It is indicated in the figure, and it has been obtained following the method reported in [24]. It will be later discussed in section 4 how to add further elements to the model of figure 3b to properly describe the behavior of the OCSRR-loaded CPW above the first resonance frequency.

Figure 5. Electromagnetic simulation (transmission and reflection coefficients) of the structure shown in figure 3(a). In the figure are also included the transmission and reflection coefficients obtained from the circuit model of figure 3(b) (with extracted parameters), the circuit model of figure 3(b) with an additional inductance to take into account the transmission zero (see figure 6(a)), and the circuit of figure 3(b) with an additional resonant element to account for the third resonance of the OCSRR (see figure 6(b)). The elements values are: $L_0=10.41\text{nH}$, $C_C=2.33\text{pF}$, $L_i=1.53\text{nH}$, $L_r=0.69\text{nH}$ and $C_r=2.45\text{pF}$. The first, f_0 , second, f_1 , and third, f_2 , resonance frequencies of the OCSRR, as well as the transmission zero frequency, f_z , of the structure, are indicated.

Figure 6. Improved circuit models of the CPW structure loaded with a pair of OCSRR to take into account the transmission zero (a) and the transmission zero and third resonance of the OCSRR (b).

Figure 7. Test structure for the determination of the effects of varying the dimensions of the metallic region present between the central strip of the CPW and the inner metallic region of the OCSRR (a) and simulated responses with several lengths, D , (b) and widths, X , (c). The host CPW has a slot and central strip widths of $G=1\text{mm}$ and $W=3.64\text{mm}$, respectively. The length of the structure is 14.4mm . The dimensions of the OCSRRs are, in reference to figure 2, $c=0.4\text{mm}$, $d=0.4\text{mm}$, $r_{ext}=7\text{mm}$, $X=1.6\text{mm}$ in (b)

and $D=3.7\text{mm}$ in (c). For the substrate, the dielectric constant $\epsilon_r=10.2$ and thickness $h=1.27\text{mm}$.

Figure 8. Magnetic current diagrams of the OCSRR obtained at the relevant frequencies of figure 5. Given that the structure is symmetric, one half of the structure has been represented.

Figure 9. OCSRR with metallic strips connecting the outer and inner metallic region in order to suppress the 1st and 3rd resonances (a), and simulated frequency response (b).

Figure 10. Strategy to suppress the 2nd resonance of the OCSRR with only one metallic strip (a), adding a thin strip (b) and adding a wide strip (c).

Figure 11. Frequency response for the structures represented in figure 10.

Figure 12. Layout of the fabricated low pass filter prototype. The dimensions of the four smaller OCSRRs are (in reference to figure 2), $c=0.2\text{mm}$, $d=1.83\text{mm}$ and $r_{ext}=2.9\text{mm}$. The dimensions of the five larger OCSRRs are, $c=0.2\text{mm}$, $d=2.43\text{mm}$ and $r_{ext}=3.5\text{mm}$. Other filter dimensions are indicated. The device has been fabricated on a substrate with dielectric constant $\epsilon_r=11$ and thickness $h=1.27\text{mm}$.

Figure 13. Measured and simulated transmission (S_{21}) and reflection (S_{11}) coefficients for the filter of figure 12.

Figure 14. Current diagrams of the filter of figure 12 at the indicated frequencies.

Figure 15. Layout (a) and photograph (b) of a composite right left handed line implemented in microstrip technology and designed to behave as a band pass filter. The substrate is the *Rogers RO3010* with thickness $h = 0.254 \text{ mm}$ and dielectric constant $\epsilon_r = 11.2$. Dimensions are: $l = 22.3 \text{ mm}$, $W = 0.21\text{mm}$. For OCSRR: $r_{ext} = 2.7 \text{ mm}$, $c = 0.2 \text{ mm}$, $d = 1.2 \text{ mm}$. For OSRR: $r_{ext} = 4\text{mm}$, $c = 0.4 \text{ mm}$, $d = 0.2 \text{ mm}$.

Figure 16. Simulated and measured transmission (S_{21}) and reflection (S_{11}) coefficients for the filter of figure 15.

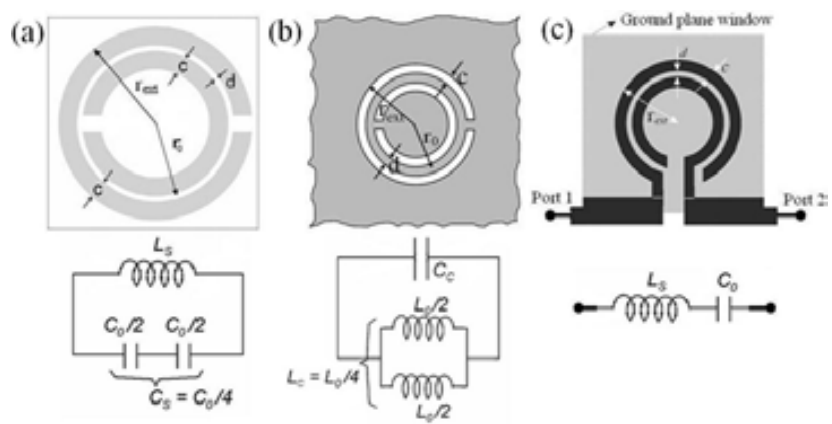


Figure 1, F. Aznar *et al.*

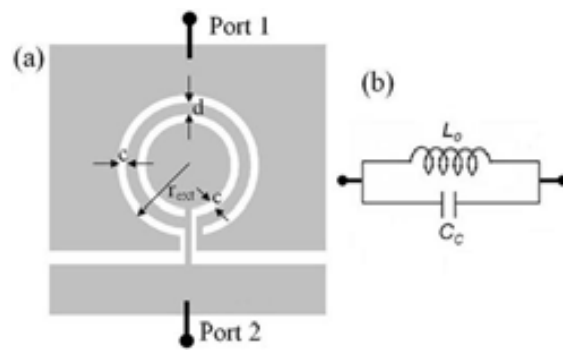


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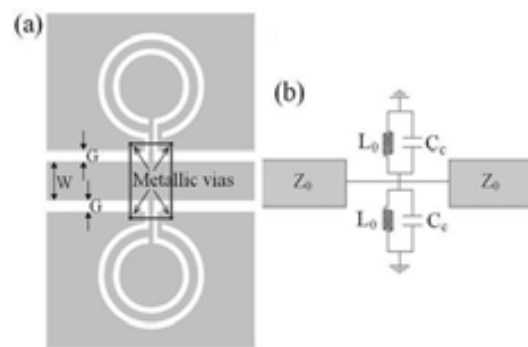


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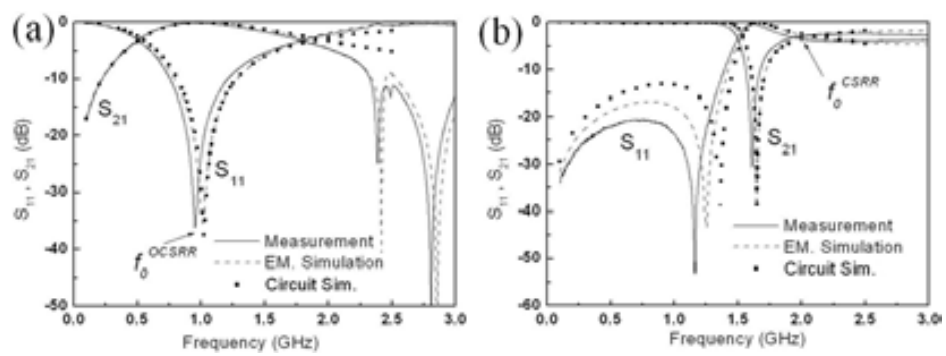


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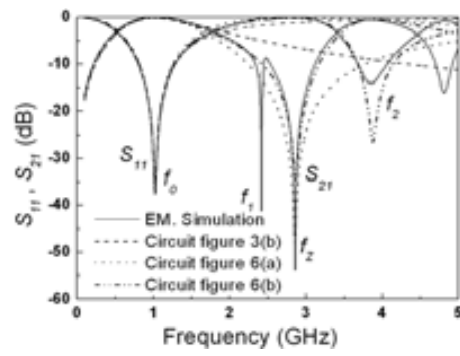


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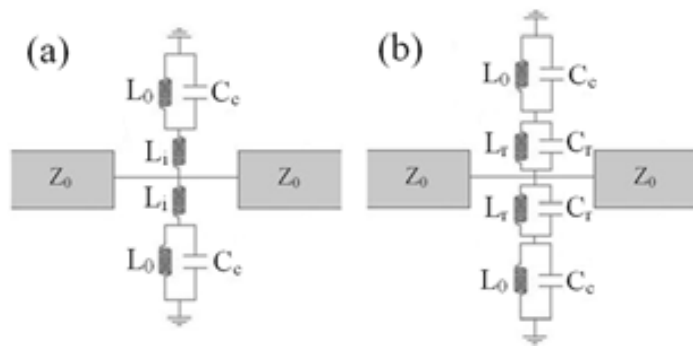


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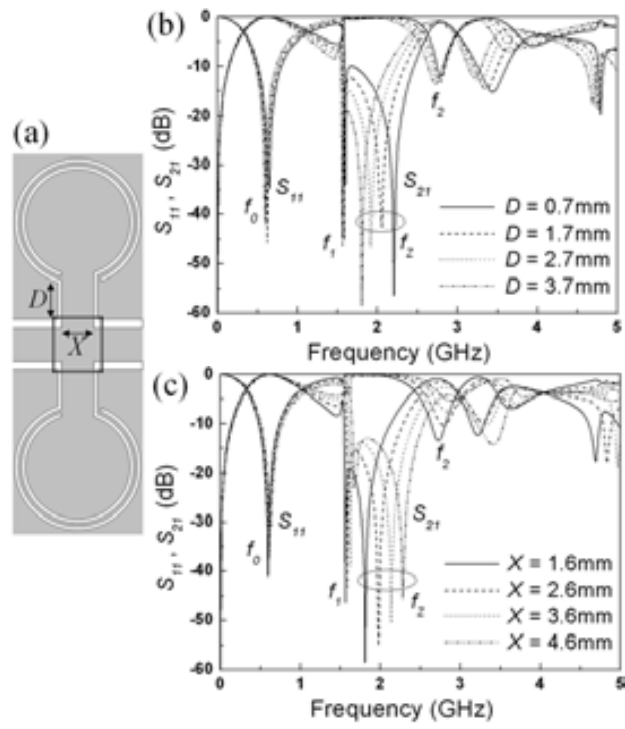


Figure 7, F. Aznar *et al.*

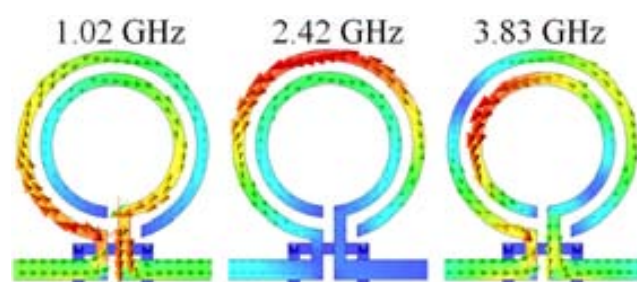


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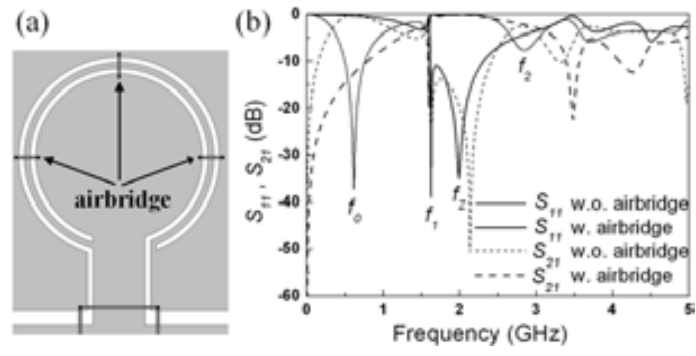


Figure 9, F. Aznar *et al.*

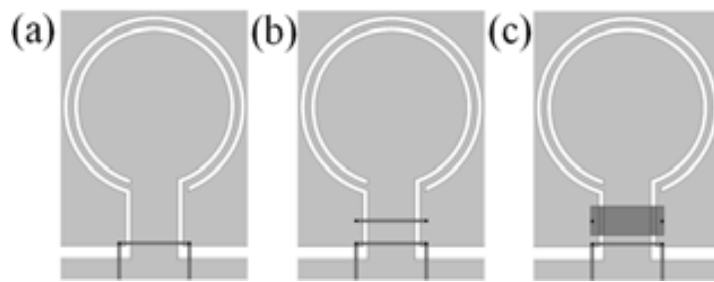


Figure 10, F. Aznar *et al.*

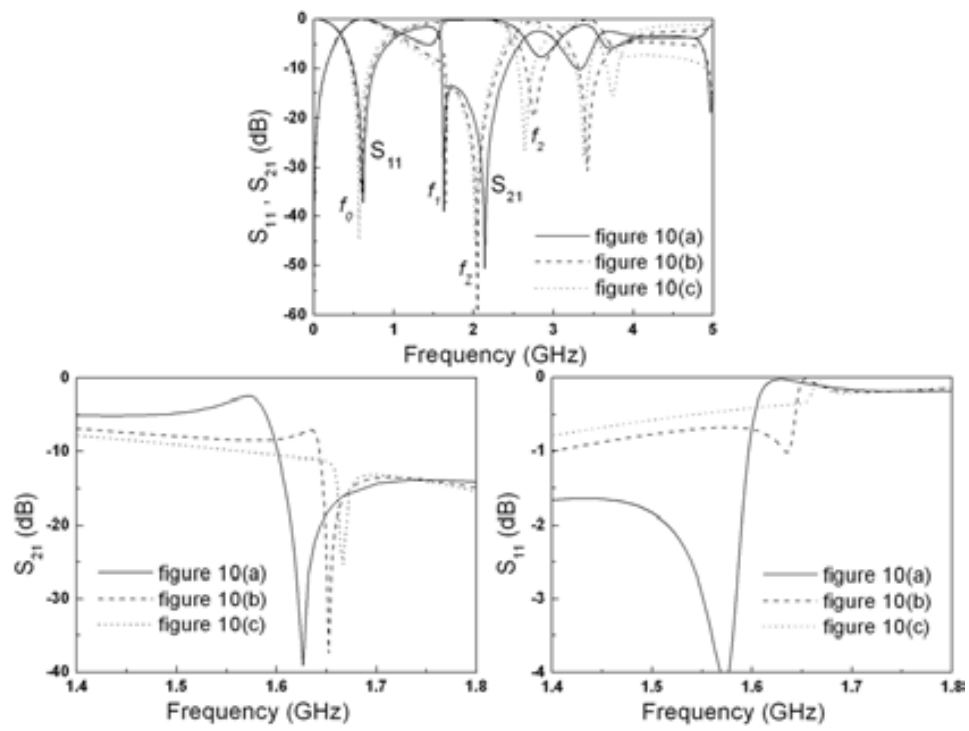


Figure 11, F. Aznar *et al.*

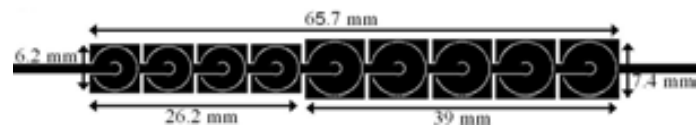


Figure 12, F. Aznar *et al.*

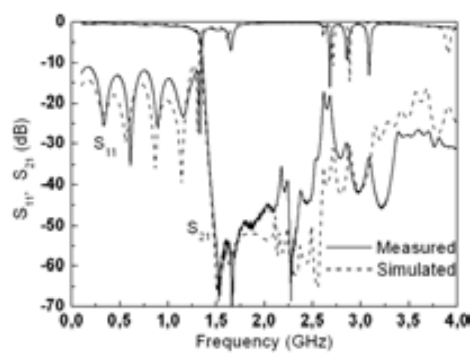


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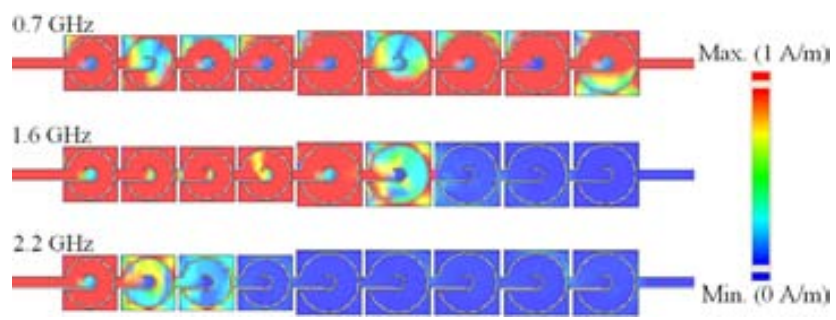


Figure 14, F. Aznar *et al.*

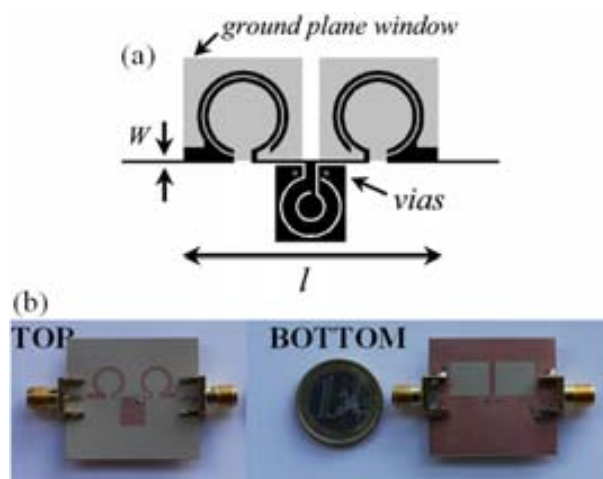


Figure 15, F. Aznar *et al.*

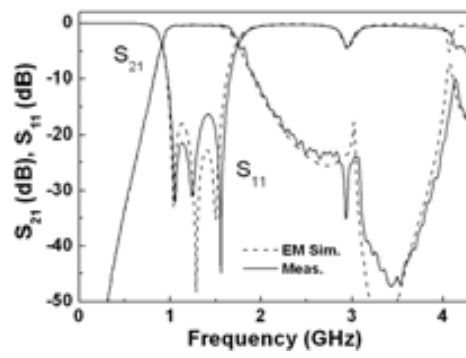


Figure 16, F. Aznar *et al.*

Artículo D

Referencia:

F. Aznar, M. Gil, J. Bonache, and F. Martín. “Revising the Equivalent Circuit Models of Resonant-Type Metamaterial Transmission Lines”, *IEEE MTT-S International Microwave Symposium Digest*, Atlanta (USA), pp. 323-326. June 2008.

Resumen del artículo:

En este trabajo, se demuestra que los modelos circuitales existentes para líneas zurdas resonantes cargadas con *split ring resonators* (SRRs) o con *complementary split ring resonator* (CSRRs), necesitan una revisión con el fin de tener en cuenta los diferentes elementos de las estructuras y para describir con precisión su comportamiento. La conclusión principal y novedad de este artículo es que los modelos ya existentes son formalmente correctos, pero algunos de sus parámetros no tenían una interpretación física hasta ahora. Sin embargo, estos parámetros están relacionados con los parámetros de los modelos mejorados, que se proponen y analizan exhaustivamente aquí por primera vez. Se incluye un análisis comparativo de líneas cargadas con SRR y CSRR, a la luz de estos nuevos modelos. Los resultados de este trabajo son de interés para el diseño de circuitos de microondas basados en CSRRs y SRRs.

Revising the Equivalent Circuit Models of Resonant-Type Metamaterial Transmission Lines

Francisco Aznar, Marta Gil, Jordi Bonache and Ferran Martín

GEMMA/CIMITEC, Departament d'Enginyeria Electrònica. Universitat Autònoma de Barcelona.
08193 BELLATERRA (Barcelona), Spain. E-mail: Ferran.Martin@uab.es

Abstract — In this work, it is shown that the previously reported and accepted equivalent circuit models of resonant type left-handed lines, either loaded with split ring resonators (SRRs), or with complementary split ring resonators (CSRRs), need a revision in order to properly account for the different elements of the structures and to accurately describe their behavior. The main relevant conclusion and novelty of this paper is that the already existing models are formally correct, but some of the reactive parameters do not have the physical interpretation given so far. However, these parameters are related to the parameters of the improved models, which are proposed and exhaustively analyzed here for the first time. A comparative analysis of SRR- and CSRR-loaded lines, to the light of these new models, is also included. The results of this work are of interest for the design of CSRR- and SRR-based microwave circuits.

Index Terms — Metamaterial transmission lines, split ring resonators (SRRs), complementary split ring resonators (CSRRs), equivalent circuits.

I. INTRODUCTION

Resonant-type metamaterial transmission lines, either loaded with split ring resonators (SRRs) [1] or with complementary split ring resonators (CSRRs) [2] (Fig. 1), were proposed in 2003 and 2004, respectively, as an alternative to artificial lines loaded with series capacitances and shunt inductances [3-5] (CL-loaded approach). In all cases, the structures exhibit a composite right/left handed (CRLH) behavior [6,7], that is, backward wave propagation at low frequencies, and forward wave propagation at higher frequencies. The propagation characteristics of these artificial lines have been interpreted to the light of their lumped element equivalent circuits. For SRR- and CSRR-loaded lines, the reported models are depicted in Fig. 2 [8]. To achieve CRLH wave propagation in CSRR-loaded lines, series capacitive gaps are required, whereas shunt inductive elements are necessary in SRR-based lines. In [8], a physical interpretation was given to the different reactive parameters of both circuit models. However, in this work it is demonstrated that, by extracting such parameters, we are not actually able to link all of them to the different elements of the structures (SRRs, CSRRs, series gaps, shunt inductive strips and host line), in spite of the excellent agreement between circuit and electromagnetic simulations (or measurement) of the structures. As will be shown, the circuits of Fig. 2 are inferred through circuit transformation from other circuit models

where the reactive parameters have a direct correspondence with the different planar components and, hence, a clear physical interpretation. The relationships between the elements of the original and transformed models are obtained. From these relations, the behavior of the structures is interpreted, and interesting implications on circuit design are pointed out and discussed.

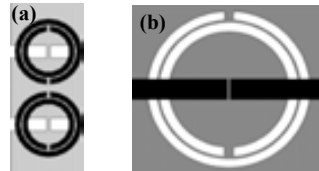


Fig. 1. Typical unit cell of a CRLH line based on SRRs (a) and CSRRs (b). In (a), the SRRs are etched in the back substrate side of a coplanar waveguide (CPW) transmission line (the upper metal level is depicted in gray). In (b), the host line is a microstrip transmission line and the CSRR is etched in the ground plane (depicted in gray).

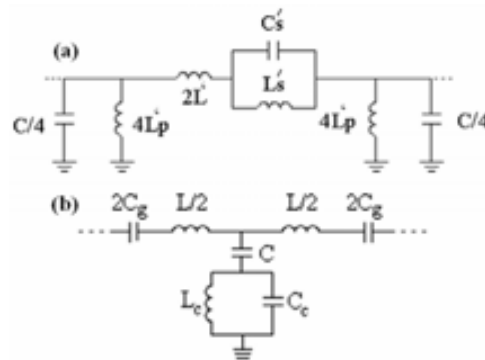


Fig. 2. Lumped element equivalent circuit model of the unit cell of a CRLH line based on SRRs (a) and CSRRs (b).

II. IMPROVED MODEL FOR CSRR-LOADED CRLH LINES

According to previous works [8], the T-circuit of Fig. 2(b) models a CSRR-loaded CRLH line, where the CSRRs are described by the resonant tank L_c-C_c , L models the line inductance, C_g the gap capacitance and C the electric coupling between the line and the CSRRs. The parameter extraction method reported in [9] reveals that there is very

good agreement between electromagnetic simulation (or measurement) and circuit simulation. This means that the T-circuit model of Fig. 2(b) provides a good description of device behavior. It was argued that C is composed of the line capacitance and the fringing capacitance of the gap. However, the extremely high values of C previously reported [9] can not be explained to the light of this interpretation. It has been found that C increases by increasing gap distance. On the other hand, it has been found that C_g does not experience a significant decrement when gap distance is increased. The results of table I, corresponding to the extracted parameters of identical structures with different gap distance, corroborate these statements. The considered substrate is the *Rogers RO3010* with thickness $h=1.27\text{mm}$ and dielectric constant $\epsilon_r=10.2$. Line width is 1.15mm and gap width is 4.8mm . CSRR dimensions are: internal radius $r_i=2.69\text{mm}$, width $c=0.36\text{mm}$ and separation $d=0.24\text{mm}$.

TABLE I: EXTRACTED PARAMETERS OF IDENTICAL CSRR-LOADED UNIT CELLS WITH DIFFERENT GAP DISTANCE

Gap space (mm)	C_g (pF)	C (pF)	L (nH)	L_c (nH)	C_c (pF)
1.56	0.45	21.3	3.30	1.46	3.14
0.76	0.49	8.8	4.10	1.58	2.90
0.16	0.54	3.8	5.35	1.69	2.70

To properly model the series gap and the coupling between the line and the CSRR, we must consider the π -circuit model of the series gap. Thus, the new proposed equivalent circuit of the CSRR-loaded CRLH line is that depicted in Fig. 3, where C_L is the line capacitance, C_f is the fringing capacitance of the gap and C_s is the series capacitance of the gap. Obviously, from π -T transformation, the circuit model of Fig. 2(b), which is the reported model of microstrip lines loaded with CSRRs and series gaps, is obtained, but the values of C_s and C do not actually have a physical interpretation. Indeed, C_g and C can be expressed in terms of C_s and $C_{par}=C_f+C_L$ according to:

$$C_g = 2C_s + C_{par} \quad (1)$$

$$C = \frac{C_{par}(2C_s + C_{par})}{C_s} \quad (2)$$

and expressions (1) and (2) explain the behavior of C and C_g when the gap distance is increased (C_s decreased). We have considered isolated gaps, and we have estimated C_s and C_f from the S-parameters obtained through electromagnetic simulation. By using equations (1) and (2) we have obtained the values of C and C_g . Reasonable agreement with those values inferred from the parameter extraction method (shown in table I) results. These values are [$C_g=0.45\text{pF}$, $C=17.8\text{pF}$], [$C_g=0.53\text{pF}$, $C=7.63\text{pF}$] and [$C_g=0.70\text{pF}$, $C=3.67\text{pF}$], for the structures with a gap distance of 1.56mm , 0.76mm and 0.16mm , respectively. These results are of interest because they reveal that it is possible to obtain high values of C (by decreasing C_s), regardless of the substrate thickness. These high values of C are typically necessary to enhance bandwidth and to drive the transmission zero frequency

$$\omega_z = 2\pi f_z = \frac{1}{\sqrt{L_c(C+C_c)}} \quad (3)$$

to small values. However, this can be achieved without the penalty of a small C_g , since C_f and C_L do also contribute to this capacitance (see expression 1). The increment of L in table I when the gap distance is decreased is due to the strip length variation. CSRR parameters (L_c and C_c) vary since these parameters are influenced by the geometry of the upper metallic layer (strips and series gap).

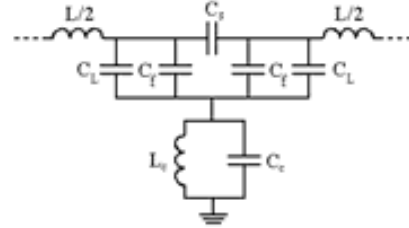


Fig. 3. Improved lumped element equivalent circuit model of the unit cell of a CRLH line based on CSRRs.

III. IMPROVED MODEL FOR SRR-LOADED CRLH LINES

In [8], it was demonstrated that the π -circuit of Fig. 2(a) is obtained by transformation of the series impedance of a primary model, intended to describe the physics of SRR-loaded CRLH transmission lines. In such primary model, the series impedance is composed of the line inductance, L , inductively coupled to the SRRs (described by a resonant tank L_s - C_s) through a mutual inductance M , whereas the shunt impedance accounts for the line capacitance, C , and the shunt inductive strips (or vias in microstrip technology), modeled by the inductance L_p (due to symmetry considerations, the magnetic wall concept was used in [8]). It has been found through parameter extraction (using a technique similar to that reported in [9] for CSRR-loaded lines) that the model of Fig. 2(a) provides a good description of device behavior. However, the parameters of the primary model, related to those of the circuit of Fig. 2(a) through

$$C_s' = \frac{L_s}{\omega_o^2 M^2} \quad (4)$$

$$L_s' = \omega_o^2 M^2 C_s \quad (5)$$

$$L_p' = L_p \quad (6)$$

$$L' = L \quad (7)$$

are not representative of the different elements of the structures. In particular, it has been found that the transmission zero frequency, ω_z , depends on the characteristics of the inductive strips, something not accounted for by the model (ω_z should coincide with the resonance frequency of the SRRs, $\omega_b=(L_s C_s)^{-1/2}=(L_s' C_s')^{-1/2}$, according to the model). The simulated transmission coefficients of two identical structures (unit cells), one with shunt strips (CRLH

line) and the other one without such elements (negative permeability line), clearly points out the variation of ω_z (Fig. 4).

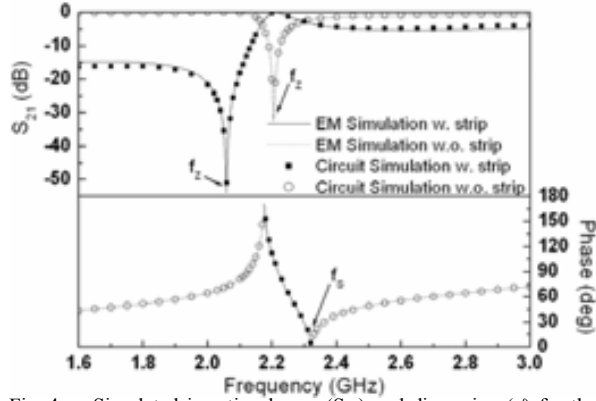


Fig. 4. Simulated insertion losses (S_{21}) and dispersion (ϕ) for the indicated SRR-loaded CPW unit cells with and without shunt inductive strips. The circuit simulation with extracted parameters is also included (symbols). The considered substrate has a thickness of $h=1.524\text{mm}$ and a dielectric constant of $\epsilon=10.2$. For the CPW, central strip width is $W=7\text{mm}$ and slot width is $G=1.48\text{mm}$. The width of the shunt strips is 0.2mm . SRR dimensions are: internal radius $r_o=2.4\text{mm}$, width $c=0.6\text{mm}$ and separation $d=0.2\text{mm}$. We have actually represented the modulus of the phase.

Thus, although the circuit of Fig. 2(a) is formally correct, the previously proposed primary circuit [8] is only a first order approximation, and it needs a revision. However, such improved circuit must be expressible as that of Fig. 2(a). This is the main aim of this section. The proposed improved model of SRR-loaded CRLH lines is depicted in Fig. 5(a). In contrast to the work in [8], the magnetic wall concept is not used here since it is not actually necessary. In this new model, which is neither a π - nor a T-circuit, the line inductance has been divided in two parts, and the inductance of the shunt strips, L_p , is located at the intermediate position. This provides a more accurate description of the strips location and perfectly explains the variation of ω_z when the shunt strips are removed. From the elements of the admittance matrix of the circuit of Fig. 5(a), we can obtain the equivalent π -circuit model. After a straightforward calculation, the series and shunt impedances of such π -circuit are found to be:

$$Z_s(\omega) = j\omega \left(2 + \frac{L}{2L_p} \right) \left[\frac{L}{2} - L_s' + \frac{L_s'}{1 - L_s' C_s' \omega^2} \right] \quad (8)$$

and

$$Z_p(\omega) = j\omega \left(2L_p + \frac{L}{2} \right) \quad (9)$$

with

$$L_s' = 2M^2 C_s \omega_o^2 \frac{\left(1 + \frac{L}{4L_p} \right)^2}{1 + \frac{M^2}{2L_p L_s}} \quad (10)$$

$$C_s' = \frac{L_s}{2M^2 \omega_o^2} \left(1 + \frac{M^2}{2L_p L_s} \right)^2 \left(1 + \frac{L}{4L_p} \right) \quad (11)$$

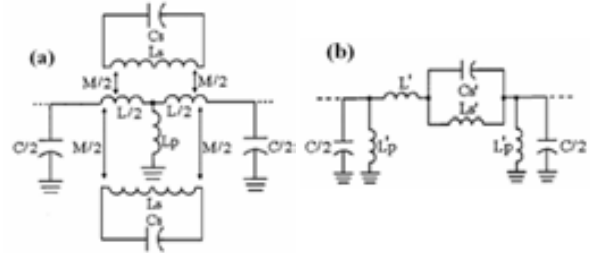


Fig. 5. Improved lumped element equivalent circuit model of the unit cell of a CRLH line based on SRRs (a) and transformed model (b).

From these results, it is clear that: (i) the model of Fig. 5(a) can be formally expressed as that of Fig. 2(a) (see Fig. 5(b)); (ii) expressions (4)-(7) are no longer valid; the link between the parameters of Figs. 5(a) and 5(b) is given by expressions (10), (11) and

$$L' = \left(2 + \frac{L}{2L_p} \right) \frac{L}{2} - L_s' \quad (12)$$

$$L_p' = 2L_p + \frac{L}{2} \quad (13)$$

(iii) the transmission zero verifies $\omega_z < \omega_o$ and it depends on L_p since it is given by $\omega_z = (L_s' C_s')^{-1/2}$, that is:

$$\omega_z = \omega_o \left(1 + \frac{M^2}{2L_p L_s} \right)^{-1/2} \quad (14)$$

We have extracted the parameters of the model of Fig. 5(b) for those structures whose insertion losses are depicted in Fig. 4. By using expressions (10)-(13) and $M=L/f$ (f being the fractional area of the slots occupied by the SRRs [8]), we have obtained the parameters of the model of Fig. 5(a). Such parameters are represented in table II. It is remarkable that (i) the circuit simulation with extracted parameters provides a very good fit to the electromagnetic simulation (see Fig. 4), and (ii) the parameters of the model of Fig. 5(a) are roughly the same with and without the presence of the shunt strips. The parameter values that have been obtained from the isolated elements (SRRs, shunt strips and line elements) are also represented in table II (3rd row). The variations are

attributed to the interaction between the CPW, the SRRs and the strips. Namely, line parameters (C and L) are affected by the presence of SRRs, and SRR parameters (C_c and L_s) are influenced by the presence of the CPW structure.

TABLE II: EXTRACTED PARAMETERS OF SRR-LOADED CPW UNIT CELLS WITH AND WITHOUT SHUNT STRIPS

Shunt strips	L_p (nH)	C (pF)	L (nH)	L_s (nH)	C_s (pF)
Yes	0.36	2.16	2.27	14.81	0.35
No	----	2.16	2.29	14.69	0.35
Estimated	0.57	1.14	2.60	12.55	0.50

IV. COMPARATIVE ANALYSIS AND DISCUSSION

It is interesting to mention that the transmission zero frequency can be controlled with the shunt strip (equation 14), in a similar form that ω_z depends on the characteristics of the gap (through C) in CSRR-loaded lines (expression 3). Specifically, widening the inductive strip and the capacitive gap (enhancing gap separation) has the same effect: lowering the transmission zero frequency. Notice that this effect is due to the fact that the strips/gaps are located above the position of the SRRs/CSRRs. Actually, this similarity is not surprising, if one takes into account that CSRR-loaded microstrip lines and SRR-loaded CPWs roughly exhibit a dual behavior. Another interesting aspect of SRR-loaded CPW lines is that the frequency where the phase nulls, ω_s , does not depend on L_p . In a π - or T-circuit, such phase is given by:

$$\cos \phi = 1 + \frac{Z_s(\omega)}{Z_p(\omega)} \quad (15)$$

Thus, by forcing $Z_s(\omega)=0$, we obtain such frequency, that is:

$$\omega_s = 2\pi f_z = \frac{1}{\sqrt{C_s \left(L_s - 2 \frac{M^2}{L} \right)}} \quad (16)$$

and it does not depend on L_p , as anticipated. It has been verified from electromagnetic simulation (Fig. 4) that identical value of ω_s is obtained with and without the presence of shunt strips. Analogously, ω_z does not depend on the presence of the series gap in CSRR-loaded microstrip lines. According to the model of Fig. 3, this frequency is obtained when $Z_p(\omega)=\infty$, and this occurs at the resonance frequency of CSRRs, $\omega_0=(L_c C_c)^{-1/2}$ (such frequency is the same for the 3 considered cases of table I). Again, we obtain a similar behavior concerning the frequency where the phase nulls in both the SRR- and CSRR-loaded lines. This further supports the dual behavior of the considered structures.

V. CONCLUSIONS

In conclusion, the previously existing and accepted models of SRR- and CSRR-loaded metamaterial transmission lines have been revised. The new proposed models accurately account

for the different elements of the structures. It has been demonstrated that although the new models can be transformed to π - or T-circuits formally identical to the previous ones, the elements of these new transformed circuits are very different, and they are able to perfectly explain the electromagnetic behavior of the structures. The roughly dual behavior of the two considered structures has also been discussed.

ACKNOWLEDGEMENT

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Artículo E

Referencia:

F. Aznar, J. Bonache and F. Martín, “Improved circuit model for left-handed lines loaded with split ring resonators”, *Applied Physics Letters*, Vol. 92, pp. 043512. February 2008.

Resumen del artículo:

En este artículo, se presenta y discute el modelo mejorado del equivalente circuital de una línea de transmisión zurda basada en *split ring resonators* (SRRs). Está rigurosamente demostrado que aunque el anterior modelo circuital aceptado para estas líneas de transmisión metamaterial (un circuito en π) proporciona una buena descripción del comportamiento del dispositivo, sus parámetros eléctricos no describen la realidad física de la estructura. Por el contrario, los parámetros del nuevo modelo circuital equivalente representan los diferentes elementos de la estructura, incluyendo los SRRs, los elementos inductivos en derivación y la línea base. También se muestra que el modelo propuesto puede ser transformado a un modelo en π que es formalmente idéntico al modelo anterior de línea zurda basada en SRRs. Con esta transformación, se interpretan perfectamente las principales características re-levantes de estas líneas zurdas.

Improved circuit model for left-handed lines loaded with split ring resonators

F. Aznar,^{a)} J. Bonache, and F. Martín^{b)}

CIMITEC, Departament d'Enginyeria Electrònica, Universitat Autònoma de Barcelona, Bellaterra, 08193 Barcelona, Spain

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In this letter, an improved lumped element equivalent circuit model for left-handed lines based on split ring resonators (SRRs) is presented and discussed. It is rigorously demonstrated that although the previously accepted circuit model of these metamaterial transmission lines (a π circuit) provides a good description of device behavior, its electrical parameters do not actually describe the physics of the structure. Conversely, the parameters of the improved equivalent circuit model are representative of the different elements of the structure, including the SRRs, the shunt inductive elements and the host line. It is also shown that the proposed model can be transformed to a π model which is formally identical to the previous reported model of SRR-based left-handed lines. With this transformation, the main relevant characteristics of these left-handed lines are perfectly interpreted. © 2008 American Institute of Physics. [DOI: 10.1063/1.2839600]

Resonant-type metamaterial transmission lines were proposed by Martín *et al.*¹ in 2003 by loading a coplanar waveguide (CPW) structure with split ring resonators (SRRs).² Such resonators, consisting on a pair of concentric metallic rings with slits etched on opposite sides, are electrically small³ and were formerly used in 2000 by Smith *et al.*⁴ for the synthesis of a bulk left-handed medium. Indeed, the left-handed SRR-based CPW reported in Ref. 1 was inspired on the medium of Smith *et al.*⁴ By etching SRRs in the back substrate side, beneath the slots, and shunt connected metallic strips between the central strip and ground plane, a one-dimensional effective medium with simultaneous negative permeability (due to the presence of the SRRs) and permittivity (thanks to the shunt strips) in a narrow band was achieved.¹ It was also demonstrated^{1,5} that by removing the shunt connected strips of the structure, a stop-band characteristic is obtained, and this was interpreted as due to the negative and highly positive effective permeability of the structure above and below SRR's resonance, respectively.

Alternatively, the behavior of these SRR-based structures (with and without shunt connected strips) can be interpreted to the light of the lumped element equivalent circuit models of the unit cells, which were reported in Ref. 1 and are reproduced here for completeness (see Fig. 1). As was reported in Ref. 1, L and C account for the line inductance and capacitance, respectively, C_s and L_s model the SRR, M is the mutual inductive coupling between the line and the SRRs, and L_p is the inductance of the shunt strips. In the models of Fig. 1, the magnetic wall concept has been used. From the transmission line approach of metamaterials,^{6,7} it follows that the structure exhibits left-handed wave propagation in those regions where the series reactance and shunt susceptance are negative, whereas the requirement for an effective medium with negative permeability is a negative series reactance and a positive shunt susceptance. According to this, the models of Fig. 1 perfectly explain the propagation

characteristics of the structures. However, the unit cell model of the left-handed structure is not able to accurately describe its behavior. The reasons are investigated in this paper, where we propose an improved model which can be transformed to that of Fig. 1(c) with modified parameters.

According to the π -circuit models of Fig. 1, the structures should exhibit a transmission zero (all injected power is returned back to the source) at that frequency ω_z , where the series branch opens [$Z_s(\omega_z)=\infty$, where $Z_s(\omega)$ is the impedance of the series branch], and it occurs at the resonance frequency of the SRRs ω_0 for both circuit models. In other words, the transmission zero should be located at identical frequency (provided identical SRRs are used) in both structures (with and without shunt strips). Moreover, according to the models of Fig. 1, the allowed band in the left-handed CPW structure should be slightly shifted toward positive values as compared to the stop band in the negative-permeability line (structure without shunt strips). This is due

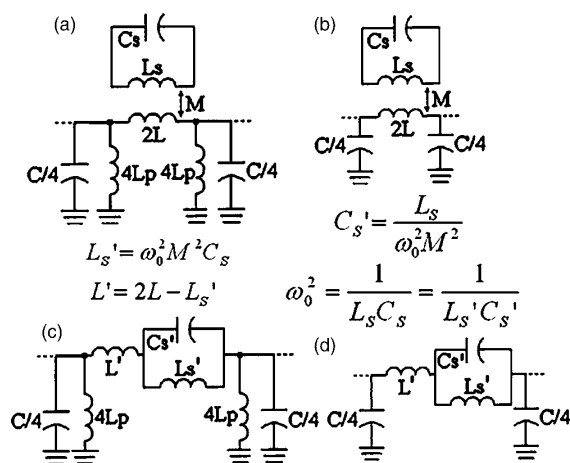


FIG. 1. Lumped element equivalent circuit for the basic cell of the left-handed (a) and negative-permeability (b) CPW structures loaded with SRRs. These circuits can be transformed to those depicted in (c) and (d), according to the indicated transformations.

^{a)}Electronic mail: francisco.aznar@uab.es.

^{b)}Tel.: 34935813522. Fax: 34935812600.

Electronic mail: ferran.martin@uab.es.

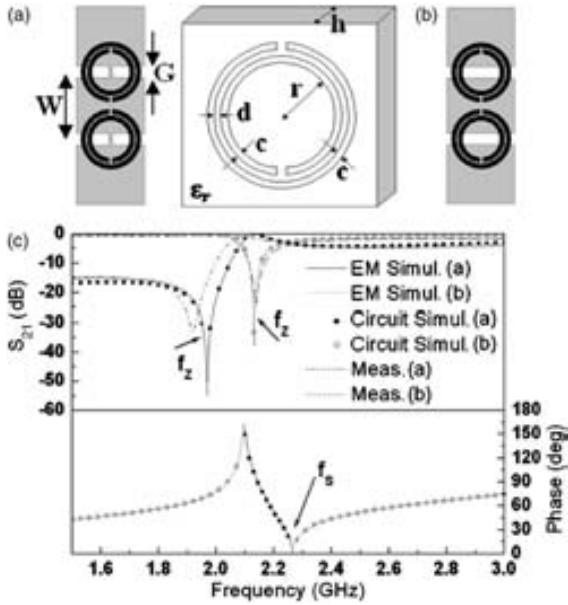


FIG. 2. Layout of the considered CPW structures with SRRs and shunt strips (a) and with SRRs only (b); simulated and measured transmission coefficient S_{21} and simulated dispersion relation (c). The considered substrate is the Rogers RO3010 with thickness $h=1.27$ mm and dielectric constant $\epsilon_r=10.2$. Relevant dimensions are rings width $c=0.6$ mm, distance between the rings $d=0.2$ mm, and internal radius $r=2.4$ mm. For the CPW structure, the central strip width is $W=7$ mm and the width of the slots is $G=1.35$ mm. Finally, the shunt strip width is 0.2 mm. The results of the electrical simulation with extracted parameters are depicted by using symbols. We have actually represented the modulus of the phase since it is negative for the left-handed line. Discrepancy between measurement and simulation for circuit (a) is attributed to fabrication related tolerances.

to the fact that the left-handed band is located to the right of $\omega_0=\omega_z$, whereas the stop band extends below and above ω_0 , as has been explained above. In the negative-permeability and left-handed CPW structures reported in Ref. 1, these features apparently occur. However, these structures are composed of four cascaded unit cells, and this may obscure device behavior. Namely, to better identify the position of the transmission zero and the allowed band (left-handed structure), it is convenient to consider a single unit cell line. Under these conditions, the transmission zero is not obscured by surface waves, and the transmission peak in the left-handed structure can be easily identified since the effects of losses are minimized. Thus, we have considered two identical single cell structures, one with shunt strips, the other one without shunt strips (Fig. 2). The transmission (S_{21}) and reflection (S_{11}) coefficients of these structures have been obtained from full wave electromagnetic simulation using the AGILENT MOMENTUM commercial software. We have also obtained the dispersion relation of this structure, following the standard procedure:

$$\cos(\beta l) = \frac{A+D}{2}, \quad (1)$$

where β is the phase constant, l is the unit cell length, and A and D are the diagonal elements of the transmission ($ABCD$) matrix, which can be inferred from the simulated reflection and transmission coefficients.⁸ These results are depicted in Fig. 2. Two relevant characteristics are readily visible: (i) the

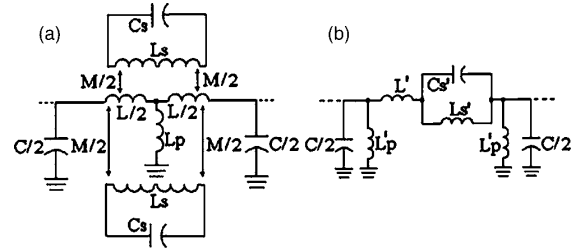


FIG. 3. Proposed improved circuit model for the basic cell of the left handed CPW structure (a). Transformation of the model to a π circuit (b), according to the procedure described in the text.

transmission zero is shifted toward negative values in the left-handed line, as compared to the negative-permeability line, and (ii) the phase shift, $\phi=\beta l$, nulls at identical frequency in both structures. From the π -circuit model, the dispersion diagram can be inferred as follows:

$$\cos(\beta l) = 1 + \frac{Z_s(\omega)}{Z_p(\omega)}, \quad (2)$$

where $Z_s(\omega)$ and $Z_p(\omega)$ are the series and shunt impedances of the π -circuit model. According to the results of Fig. 2, the models of Fig. 1 cannot be both correct since they predict identical transmission zero frequency. Thus, we must assume that at least one of the π circuits shown in Fig. 1 is incorrect, as it is. Specifically, the series branch must be different in order to explain the different values of the transmission zero frequency. However, surprisingly, in spite of the different series impedances, these impedances must null at identical frequency, since $\phi=0$ when $Z_s(\omega)=0$ (the left-handed and negative-permeability structures exhibit zero-degree phase shift at the same frequency).

As will be shown, this phenomenology can perfectly be explained by considering an improved circuit model for the left-handed line (leaving the circuit model of the negative-permeability structure unaltered). This is depicted in Fig. 3(a) (the magnetic wall concept is not considered since this is not actually necessary). In this model, the reactive parameters have the same interpretation, as in the models of Fig. 1. However, the inductance of the shunt inductive strips L_p is now located between the two inductances ($L/2$) that model each line section, to the left and right of the position of the shunt strips. This improved model reflects the location of the inductive strips, as was reported in Ref. 9. The model is neither a π circuit nor a T circuit. Thus, the transmission zero frequency and the frequency where the phase shift nulls cannot be directly obtained from it. The main relevant aspect of this letter is to demonstrate that the model of Fig. 3(a) can be transformed to a π circuit, formally identical to that of Fig. 1(c) with modified parameters. From this transformation, the shift of the transmission zero toward lower frequencies as compared to the structure without shunt strips, and the preservation of the frequency where $\phi=0$, can perfectly be explained.

Due to symmetry considerations and reciprocity, the admittance matrix of the circuit of Fig. 3(a) (which is a biport) must satisfy $Y_{12}=Y_{21}$ and $Y_{11}=Y_{22}$. From these matrix elements, the equivalent π -circuit model can be obtained according to Ref. 8,

$$Z_s(\omega) = -(Y_{21})^{-1}, \quad Z_p(\omega) = (Y_{11} + Y_{21})^{-1}. \quad (3)$$

Y_{21} is inferred by grounding port 1 and obtaining the ratio between the current at port 1 and the applied voltage at port 2. Y_{11} is simply the input admittance of the biport, seen from port 1, with a short circuit at port 2. After a straightforward but tedious calculation, the elements of the admittance matrix are obtained, and by applying Eqs. (3), we finally obtain

$$Z_s(\omega) = j\omega \left(2 + \frac{L}{2L_p} \right) \left[\frac{L}{2} + M^2 \frac{1 + \frac{L}{4L_p}}{L_s \left(\frac{\omega_0^2}{\omega^2} - 1 \right) - \frac{M^2}{2L_p}} \right], \quad (4a)$$

$$Z_p(\omega) = j\omega \left(2L_p + \frac{L}{2} \right), \quad (4b)$$

with $\omega_0 = (L_s C_s)^{-1/2}$. Expression (4a) can be rewritten as

$$Z_s(\omega) = j\omega \left(2 + \frac{L}{2L_p} \right) \left[\frac{L}{2} - L'_s + \frac{L'_s}{1 - L'_s C'_s \omega^2} \right], \quad (5)$$

with

$$L'_s = 2M^2 C_s \omega_0^2 \frac{\left(1 + \frac{L}{4L_p} \right)^2}{1 + \frac{L}{2L_p L_s}}, \quad C'_s = \frac{L_s}{2M^2 \omega_0^2} \left(\frac{1 + \frac{M^2}{2L_p L_s}}{1 + \frac{L}{4L_p}} \right)^2. \quad (6)$$

These results indicate that the improved circuit model of the unit cell of the left-handed lines loaded with SRRs and shunt inductors [Fig. 3(a)] can be formally expressed as the π circuit of Fig. 1(c), but with modified parameters [Fig. 3(b)]. This modified parameters are related to the parameters of the circuit of Fig. 3(a), according to Eqs. (6) and

$$L' = \left(2 + \frac{L}{2L_p} \right) \frac{L}{2} - L'_s, \quad L'_p = 2L_p + \frac{L}{2}. \quad (7)$$

The transmission zero frequency ω_z for the circuit of Fig. 3 is no longer given by the resonance frequency of the SRRs ω_0 , but it is smaller, that is, $\omega_z \ll \omega_0$. On the other hand, the frequency where $\phi=0$, ω_s , is obtained by forcing $Z_s(\omega)=0$. This gives

$$\omega_s = \frac{1}{\sqrt{C_s \left(L_s - 2 \frac{M^2}{L} \right)}}. \quad (8)$$

Despite that $Z_s(\omega)$ is a function of L_p , unexpectedly, ω_s does not depend on the shunt inductance. This explains that ω_s is identical for both the left handed and negative-permeability structures.

This analysis reveals that the previous reported circuit model of left-handed lines loaded with SRRs and shunt inductive elements [Fig. 1(c)] is formally correct [it is formally identical to that of Fig. 3(b)]. The weakness relies on the physical interpretation of the elements of that model. It has been demonstrated in this letter that these elements [i.e., the elements of the circuit of Fig. 3(b)] do not have any physical meaning. However, they are related to the elements of the circuit model of Fig. 3(a), which describe the different components of the left-handed unit cell. To further confirm the validity of the proposed circuit model of left-handed lines

TABLE I. Extracted parameters for the SRR-loaded CPW unit cells with and without shunt strips, and estimated values (third row).

Shunt strips	L_p (nH)	C (pF)	L (nH)	M (nH)	L_s (nH)	C_s (pF)
Yes	0.36	2.44	2.17	1.18	11.23	0.50
No	...	2.44	2.21	1.20	11.22	0.50
Estimated	0.56	2.13	2.60	1.45	12.55	0.50

based on SRRs, we have considered two identical structures (unit cells), one with shunt connected strips (left-handed line), the other without these elements, and we have extracted the parameters of both structures (from the electromagnetic simulation) according to a procedure similar to that described in Ref. 10. The circuit of Fig. 3(b) has been considered for the left-handed unit cell, whereas for the structure without shunt strips, we have considered the circuit model that results by forcing $L_p = \infty$. The electrical simulation of both circuits is also depicted in Fig. 2 to ease the comparison with the electromagnetic simulation. The agreement is excellent. On the other hand, we have inverted Eqs. (6) and (7) in order to obtain the parameters of the model of Fig. 3(a), and they show a very good agreement with those parameters inferred in the structure without shunt strips (see Table I), which means that the presence of the shunt strip does not affect the parameters of the structure, hence having a clear physical interpretation. Alternatively, L_s and C_s can be inferred from the geometry of the SRRs by following the method reported in Ref. 7, L and C can be estimated from the dispersion relation and the characteristic impedance of the artificial line loaded with SRRs (i.e., without shunt strips) at low frequencies, and L_p can be obtained from independent full wave electromagnetic simulations of a zero-length CPW structure loaded with a pair of shunt strips (and modeled through a shunt inductance). The results are also given in Table I.

In conclusion, the previously existing and accepted models of SRR-loaded CPW metamaterial transmission lines have been revised. For SRR-loaded left-handed lines, some effects were found not to be explained by the former circuit model, and an improved model has been proposed and exhaustively analyzed. It has been concluded that the revised model can be transformed to a π circuit formally identical to that previously reported but with modified parameters.

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Artículo F

Referencia:

F. Aznar, M. Gil, J. Bonache, L. Jelinek, J. D. Baena, R. Marqués and F. Martín, “Characterization of miniaturized metamaterial resonators coupled to planar transmission lines”, *Journal of Applied Physics*, Vol. 104, pp. 114501. December 2008.

Resumen del artículo:

En este artículo, se propone un método para obtener las características eléctricas de resonadores metamaterial acoplados a líneas de transmisión planares. Esta técnica de extracción de parámetros se basa en la comparación entre la medida (o la simulación electromagnética) de las características de transmisión y reflexión de una línea cargada con tales resonadores y sus modelos circuitales. Las partículas resonantes consideradas en este estudio son *split ring resonators*, *spiral resonators*, y otras partículas resonantes eléctricamente pequeñas basadas en dos niveles de metal. El interés en esta técnica radica en la falta de modelos analíticos que proporcionen los parámetros eléctricos de los diferentes resonadores considerados (debido a su complejidad). De los parámetros extraídos, se concluye que el modelo circuital predice con mucha precisión la respuesta frecuencial de las estructuras consideradas. Hay un creciente interés en la síntesis de líneas de transmisión metamaterial con tamaños de celda unidad extremadamente pequeños. Con el fin de alcanzar dimensiones cerca de (o incluso por debajo) $\lambda_g/100$, es imprescindible recurrir a resonadores con complejas topologías que combinan acoplamiento *broadside* y largas tiras metálicas (grabadas en distintos niveles de metal conectados a través de vías). Esto justifica la técnica propuesta. Por tanto, este trabajo es útil para ayudar a la síntesis de componentes de microondas basados en líneas de transmisión metamaterial de tipo resonante.

Characterization of miniaturized metamaterial resonators coupled to planar transmission lines through parameter extraction

F. Aznar,^{1,a)} M. Gil,¹ J. Bonache,¹ L. Jelinek,² J. D. Baena,³ R. Marqués,² and F. Martín^{1,b)}

¹Departament d'Enginyeria Electrònica, GEMMA/CIMITEC, Universitat Autònoma de Barcelona, 08193 Bellaterra (Barcelona), Spain

²Departamento de Electrónica y Electromagnetismo, Universidad de Sevilla, Avda. Reina Mercedes, s/n, 41012 Seville, Spain

³Departamento de Física, Universidad Nacional de Colombia, Carrera 30 N°45-30 Bogotá D.C., Colombia

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In this paper, a method for obtaining the electrical characteristics of metamaterial resonators coupled to planar transmission lines is proposed. This parameter extraction technique is based on the comparison between the measured (or full wave electromagnetic simulated) transmission and reflection characteristics of a host line loaded with such resonators and those obtained from its lumped element equivalent circuit model (previously reported by some of the authors). The resonant particles considered in this study are split ring resonators, spiral resonators, and other electrically small resonant particles based on two metal levels. The interest in this technique lies in the lack of analytical models providing the electrical parameters of several of the considered ultrasmall resonator topologies (due to their complexity). From the extracted parameters, it is concluded that the circuit models predict very accurately the frequency responses of the considered structures for the different resonators under study. There is an increasing interest in the synthesis of metamaterial transmission lines with extremely small unit cell size. In order to achieve dimensions close to (or even below) $\lambda_g/100$, it is imperative to use complex resonator topologies that combine broadside coupling and extremely elongated metallic strips (etched in different metal levels connected through vias). This justifies the proposed technique. Thus, this work is useful in aiding the synthesis of microwave components based on resonant type metamaterial transmission lines. © 2008 American Institute of Physics. [DOI: 10.1063/1.3021109]

I. INTRODUCTION

Resonant type metamaterial transmission lines were first proposed in 2003 by Martín *et al.*¹ by periodic loading of a coplanar waveguide (CPW) structure with split ring resonators (SRRs) (Ref. 2) and shunt connected strips. In such structures, the metallic strips provide a negative value of the effective dielectric permittivity ϵ_{eff} below a certain frequency (plasma frequency f_p), whereas the SRRs are responsible for the negative effective permeability μ_{eff} in a narrow band above their resonance frequency f_o . By designing the structure with $f_o < f_p$, ϵ_{eff} and μ_{eff} are simultaneously negative in a certain frequency band, and left handed (or backward) wave propagation in that region occurs. Alternatively, left handed transmission lines can be implemented by loading a host line with complementary split ring resonators (CSRRs) and series (capacitive) gaps.^{3,4} In this case, the negative effective permittivity and permeability are provided by the CSRRs and the gaps, respectively. The physics and applications of these resonant type metamaterial transmission lines are exhaustively discussed in the book coauthored by two of the authors⁵ (other books that focus on the CL-loaded approach of metamaterial transmission lines have also been recently published^{6,7}).

There are two main aspects of metamaterial transmission lines that make them attractive for RF/microwave circuit design: (i) their small dimensions and (ii) the possibility of tailoring (to some extent) the dispersion diagram of such lines. The small dimensions are due to the electrically small size of the metamaterial resonators (SRRs and CSRRs, among others), which is in turn due to the high coupling between the individual rings forming the particles^{8,9} (or to self-coupling in certain resonators based on spiral topologies). The dispersion diagram of metamaterial transmission lines can be controlled by virtue of the higher number of degrees of freedom of these resonant type metamaterial transmission lines. This makes it possible to achieve functionalities not realizable with conventional lines, such as the synthesis of multiband components¹⁰ or components with enhanced operative bandwidths.¹¹⁻¹⁴

Most resonant type metamaterial transmission lines and applications reported so far have been implemented by means of CSRRs etched either in the ground plane⁵ or in the conductor strip¹⁵ of microstrip transmission lines. These lines have been exhaustively studied in the literature.¹⁶⁻¹⁸ With regard to SRR-loaded lines, a circuit model describing such lines has been previously reported. The validation of this model has been done by comparing experimental data obtained from fabricated prototypes with circuit simulations,¹ where the circuit parameters have been inferred from models valid under very restrictive conditions¹⁹ (i.e.,

^{a)}Electronic mail: francisco.aznar@uab.es.

^{b)}Tel.: 34 93 581 35 22. FAX: 34 93 581 26 00. Electronic mail: ferran.martin@uab.es.

such circuits elements have not been derived from parameter extraction). This explains certain discrepancies between the results of simulation and experiment in previous works.¹ In this work, a method for extracting the circuit parameters of metamaterial transmission lines loaded with SRRs or with other resonators magnetically coupled to the line is proposed. As will be shown, the circuit models with extracted parameters accurately describe the behavior of these metamaterial transmission lines, regardless of the resonator type and host line configuration (microstrip, CPW, etc.).

One relevant aspect of magnetically coupled metamaterial resonators is the possibility of significantly reducing their size by using two metal levels or the combination of two metal levels and metallic vias. For instance, broadside coupled split ring resonators (BC-SRRs), where two open rings are etched on opposite sides (face to face) of a dielectric slab, can be made electrically very small provided the thickness of the dielectric is also very small.¹⁹ There has also been report of the notable size reduction in broadside coupled spiral resonators (BC-SRs),²⁰ where the spirals are implemented by etching different metal loops at both sides of a dielectric layer connected through metallic vias. For certain resonator types, the complexity is so important that the analytical determination of the effective inductance and capacitance is not possible. Moreover, in simpler resonators (such as SRRs) analytical models do exist, but they are valid under very restrictive conditions, that is, without the presence of the host line. Thus, the parameter extraction technique proposed in this work is of interest since it directly provides the circuit elements of those circuits modeling the resonator-loaded lines, this being of interest for metamaterial and microwave circuit design. It will be also shown that the parameter extraction is coherent with the parameters of the isolated particles. Obviously, this will be done on the basis of transmission lines loaded with resonators whose electrical characteristics can be analytically inferred. This will further strengthen the validity of the proposed models and the parameter extraction technique.

II. THE MODEL AND PARAMETER EXTRACTION

The former model describing the unit cell of metamaterial transmission lines loaded with magnetically coupled resonators (SRRs, SRs, etc.) and shunt connected metallic elements is depicted in Fig. 1(a).¹ In Figs. 1(b) and 1(c) the layouts of typical unit cells corresponding to left handed CPW and microstrip lines, respectively, are depicted. In the CPW configuration, the SRRs are paired in the lower substrate side (where they are etched), beneath the slots of the structure and centered with the shunt strips. The negative effective permittivity is achieved by means of the shunt connected strips. In microstrip lines, the SRRs are etched in pairs on the upper substrate side, adjacent to the conductor strip. In this case, the metallic vias are responsible for the negative permittivity of the structure. As pointed out in Ref. 1, the validity of the lumped element model [Fig. 1(a)] in describing the structures in Figs. 1(b) and 1(c) is subject to the small electrical size of the unit cells. In addition to this, the coupling between adjacent resonators and losses is ne-

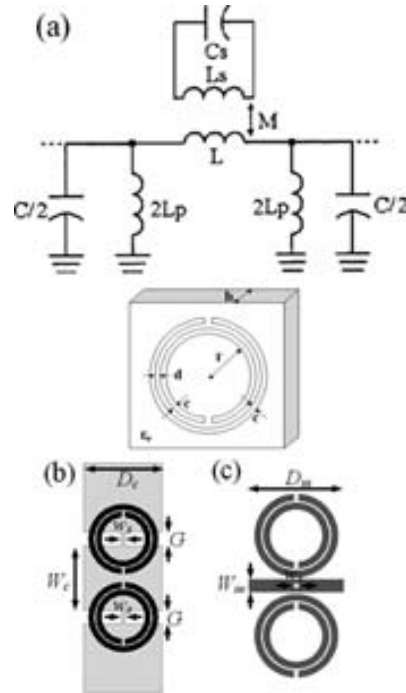


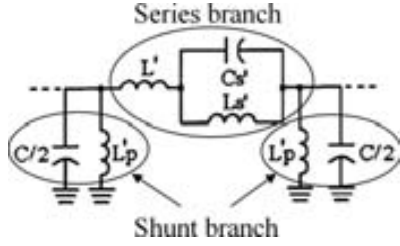
FIG. 1. Lumped element equivalent circuit for the basic left handed cell (a). Layouts of the considered CPW (b) and microstrip (c) structures combining SRRs with shunt strips and one via, respectively.

glected. It has been found that the coupling between adjacent resonant elements is significant only in square or rectangular tiny spaced resonators.²¹ [This coupling is the origin of magnetoinductive waves in chains of SRRs (Refs. 22–25) or electroinductive waves in CSRR arrays.²⁶] In reference to Fig. 1(a), L and C are the per-section inductance and capacitance of the line, L_p models the inductance of the vias (microstrip structure) or metallic strips (CPW configuration), the magnetically coupled resonators are described by means of the resonant tank constituted by the inductance L_s and the capacitance C_s , and, finally, their magnetic coupling to the line is described by the mutual inductance M . As compared to the circuit model in Ref. 1, the model in Fig. 1(a) describes the whole unit cell of the artificial lines. That is, we have not applied now the magnetic wall concept, in spite of the symmetry of the structure (in reference to a central plane oriented along the line and orthogonal to it) and excitation. The reason is that in order to extract the parameters, comparison to either simulation (electromagnetic) or experiment is required (as will be discussed later), and it involves the whole structure. Following a similar analysis to that reported in Refs. 1 and 27, the circuit in Fig. 1(a) can be transformed to the circuit model depicted in Fig. 2, where

$$L'_s = 2M^2 C_s \omega_o^2, \quad (1)$$

$$C'_s = \frac{L_s}{2M^2 \omega_o^2}, \quad (2)$$

$$L' = L - L'_s, \quad (3)$$

FIG. 2. Transformed π -circuit model of the basic left handed cell.

$$L'_p = 2L_p. \quad (4)$$

However, it has been recently demonstrated by the authors that in order to accurately describe the frequency response of left handed lines loaded with magnetically coupled resonators through a lumped element model, it is necessary to locate the inductance of the shunt strips (or vias) as Fig. 3 illustrates.²⁸ This circuit can also be transformed into that in Fig. 2. However, the transformation equations are no longer those given by expressions (1)–(4) but by²⁸

$$L'_s = 2M^2 C_s \omega_o^2 \frac{\left(1 + \frac{L}{4L_p}\right)^2}{1 + \frac{M^2}{2L_p L_s}}, \quad (5)$$

$$C'_s = \frac{L_s}{2M^2 \omega_o^2} \left(\frac{1 + \frac{M^2}{2L_p L_s}}{1 + \frac{L}{4L_p}} \right)^2, \quad (6)$$

$$L' = \left(2 + \frac{L}{2L_p}\right) \frac{L}{2} - L'_s, \quad (7)$$

$$L'_p = 2L_p + \frac{L}{2}. \quad (8)$$

Notice, however, that in the absence of the shunt connected strips or vias, $L_p \rightarrow \infty$ and expressions (5)–(8) coincide with expressions (1)–(4). As explained in Ref. 28, the model in Fig. 3 is an improved equivalent circuit of SRR-loaded left

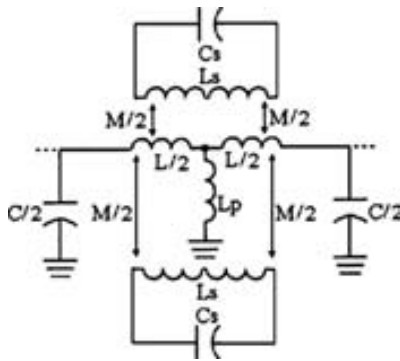


FIG. 3. Circuit model modified to describe more accurately the physical behavior of the left handed cells.

handed lines, where the element parameters have the physical interpretation given before. [Notice that L_p cannot be linked to the shunt strips or vias in the model in Fig. 1(a).] However, this model also transforms to the π -circuit in Fig. 2 [with transformation equations (5)–(8)]. The elements of the π -circuit depicted in Fig. 2, which models the unit cell of resonant type left handed transmission lines loaded with magnetically coupled resonators, can be inferred from the measured or simulated reflection and transmission coefficients of such unit cells according to the method described below.

A. The parameter extraction technique

Since the number of parameters of the circuit model in Fig. 2 is five, we also need five conditions to univocally determine such parameters. From the representation of the reflection coefficient of a single unit cell, S_{11} , in the Smith chart, two conditions are obtained. On one hand, we can determine the frequency that nulls the series reactance, f_s , from the intercept of S_{11} with the unit conductance circle. This is obvious since at this frequency, the real part of the admittance seen from the ports is simply the admittance of the opposite port, that is, $Y_o = (Z_o)^{-1} = (50 \Omega)^{-1} = 0.02 \text{ S}$. Hence, S_{11} must be allocated in the unit conductance circle at f_s , as illustrated in the example provided in Sec. II B. This frequency is given by the following expression:

$$f_s = \frac{1}{2\pi} \sqrt{\frac{1}{L'_s C'_s} + \frac{1}{L' C'_s}}. \quad (9)$$

On the other hand, the susceptance of the unit cell seen from the ports at f_s , which can be inferred from the Smith chart, is

$$B(\omega_s) = \frac{CL'_p \omega_s^2 - 2}{L'_p \omega_s}, \quad (10)$$

with $\omega_s = 2\pi f_s$. Another condition concerns the parallel resonator of the series branch. Namely, the resonance frequency of this resonator is given by

$$f_z = \frac{1}{2\pi} \sqrt{\frac{1}{L'_s C'_s}}. \quad (11)$$

Notice that this frequency does not coincide with the intrinsic resonance frequency of the magnetically driven resonator, f_o (which is the resonance frequency of the tank formed by L_s and C_s). The frequency f_z [expression (11)] can be easily obtained from the transmission coefficient S_{21} of the unit cell since at this frequency the series branch is opened and the whole power injected from the input port is reflected back to the source. Thus, the transmission coefficient nulls (transmission zero frequency) and f_z can be easily identified from the representation of the transmission coefficient in a decibel scale.

Another condition can be deduced from the phase of the transmission coefficient, $\phi_{S_{21}}$. At the frequency where $\phi_{S_{21}} = 90^\circ$, $f_{\pi/2}$, the electrical length of the unit cell, $\phi = \beta l$ (β being the phase constant and l the length of the unit cell), is

$\phi(f_{\pi/2}) = -90^\circ$.¹⁷ Since the dispersion relation of a periodic structure consisting of cascaded unit cells, as those in Fig. 2, is given by

$$\cos \phi = 1 + \frac{Z_s(\omega)}{Z_p(\omega)}, \quad (12)$$

with Z_s and Z_p being the series and shunt impedances, respectively, of the π -circuit model, it follows that

$$Z_s(\omega_{\pi/2}) = -Z_p(\omega_{\pi/2}), \quad (13)$$

with $\omega_{\pi/2} = 2\pi f_{\pi/2}$. Expressions (9)–(11) and (13) are four of the five conditions needed to univocally determine the circuit parameters in Fig. 2. Indeed, by removing the shunt connected vias or strips in the layouts in Fig. 1, we can represent the corresponding reflection coefficient on a Smith chart and obtain the susceptance seen from the ports at that frequency where S_{11} intercepts the unit conductance circle. Since this is simply the susceptance of the line capacitance (provided L_p has been removed), we can thus univocally determine C . Hence, this is the fifth condition that is required to extract the circuit parameters of the circuit model in Fig. 2.

B. Validation of the model and results

The presented method has been applied in extracting the circuit model parameters (Fig. 2) of different structures. In this section, we will illustrate the validity of both, the model and the parameter extraction method, by means of the application of the technique to a microstrip line and to a CPW left handed structure. The microstrip line is loaded with a pair of SRRs and a shunt connected via,²⁷ as depicted in Fig. 1(c). The dimensions of the SRRs are (in reference to Fig. 1) $c = 0.6$ mm, $d = 0.2$ mm, and $r = 2.4$ mm; the diameter of the via is $w_v = 0.4$ mm; and the conductor strip width is $W_m = 1.15$ mm and the length $D_m = 8.6$ mm. The considered substrate is Rogers RO3010 with thickness $h = 1.27$ mm and dielectric constant $\epsilon_r = 10.2$. The reflection coefficient of the structure (obtained from full wave electromagnetic simulation by means of the AGILENT MOMENTUM commercial software) is depicted on a Smith chart in Fig. 4(a) and both the reflection and transmission coefficients are simultaneously depicted in a decibel scale in Fig. 4(b). The phase of the transmission coefficient (also obtained from full wave electromagnetic simulation) is also depicted in Fig. 4(b). We have applied the parameter extraction technique to this structure (the results are given in Table I). From the extracted parameters, we have obtained the frequency response (reflection and transmission coefficients and phase response) from the electrical simulation (using the AGILENT ADS software) of the circuit model. The results of these circuit simulations are also depicted in Fig. 4(b) to be easily compared with the results obtained from the electromagnetic simulation of the layout. As can be appreciated, the circuit model describes the behavior of the structure in Fig. 1(c) with excellent accuracy [the circuit and electromagnetic simulations in Fig. 4(b) are roughly undistinguishable]. This points out the validity of the model and the ability of the reported technique to provide the circuit parameters of the structure.

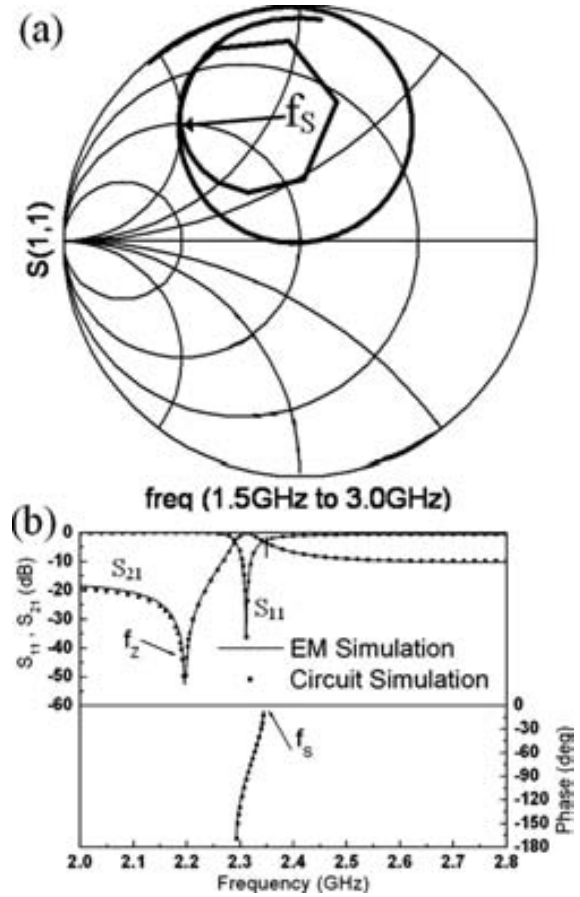


FIG. 4. Reflection coefficient on the Smith chart (a); frequency response (reflection, S_{11} , and transmission, S_{21} , coefficients) depicted in a decibel scale and the dispersion relation (b) for a left handed cell based on a microstrip structure. The considered substrate is Rogers RO3010 with thickness $h = 1.27$ mm and dielectric constant $\epsilon_r = 10.2$. Relevant dimensions are ring width $c = 0.6$ mm; distance between the rings, $d = 0.2$ mm; and internal radius $r = 2.4$ mm. For the microstrip structure the strip line width is $W_m = 7$ mm and the length is $D_m = 8.6$ mm; the diameter of via is $w_v = 0.4$ mm. The results of the circuit simulation with extracted parameters are depicted using symbols.

The same parameter extraction method has been used for the CPW transmission line loaded with a pair of SRRs and two signal-to-ground strips.¹ The SRRs are located beneath the slots of the CPW structure with the slits aligned with the

TABLE I. Extracted parameters for the circuit shown in Fig. 2 for microstrip and CPW structures with and without shunt metallic elements.

Structure	C (pF)	L' (nH)	L'_p (nH)	C'_s (pF)	L'_s (nH)
Including the shunt metallic elements					
Microstrip	1.72	11.86	2.04	3.16	1.66
CPW	2.35	4.32	1.81	4.42	1.45
Removing the shunt metallic elements					
Microstrip	1.72	3.02	...	45.11	0.11
CPW	2.35	2.06	...	20.29	0.27

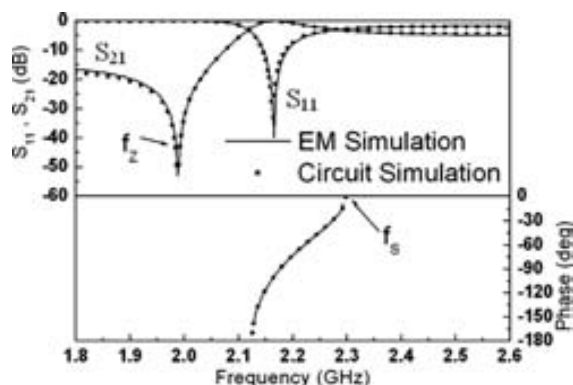


FIG. 5. Frequency response (reflection, S_{11} , and transmission, S_{21} , coefficients) depicted in a decibel scale and the dispersion relation for a left handed cell based on a CPW structure. The considered substrate is Rogers RO3010 with thickness $h=1.27$ mm and dielectric constant $\epsilon_r=10.2$. Relevant dimensions are ring width $c=0.6$ mm; distance between the rings, $d=0.2$ mm; and internal radius $r=2.4$ mm. For the CPW structure the central strip width is $W_c=7$ mm, the width of the slots is $G=1.48$ mm, and the length is $D_c=8.6$ mm; the shunt strip width is $w_s=0.4$ mm. The results of the electrical simulation with extracted parameters are depicted using symbols.

shunt connected strips. In reference to Fig. 1(b), the SRR dimensions are $c=0.6$ mm, $d=0.2$ mm, and $r=2.4$ mm; the slot and central strip widths of the host CPW are $G=1.48$ mm and $W_c=7$ mm, respectively, with a length $D_c=8.6$ mm and the shunt metallic strip width is $w_s=0.4$ mm. The relative permittivity of the considered dielectric is $\epsilon_r=10.2$ with thickness $h=1.27$ mm (Rogers RO3010). The electrical (from the extracted parameters given in Table I) and electromagnetic simulations of the structure are depicted in Fig. 5. Again, the agreement between circuit and electromagnetic simulations is excellent.

To further demonstrate this parameter extraction method, it has also been applied for the two previous structures after removing the shunt strips and the via in the CPW and the microstrip structure, respectively. In these cases, the electrical circuit has four elements (Fig. 2 excluding L'_p). The extracted parameters can be seen in Table I, whereas the frequency responses are presented in Fig. 6 with the phase for these two structures. Comparing the frequency responses for the left handed structures in Figs. 4(b) and 5 with the negative-permeability structures in Figs. 6(a) and 6(b), the conservation of the zero-degree phase frequency f_s and a shift in the transmission zero frequency f_z can be observed.²⁸

In order to generalize this model and its parameter extraction technique, a CPW left handed cell has been loaded with some different subwavelength resonators.^{20,29} The resonators that have been considered and their lengths in terms of the guided wavelength (λ_g) are broadside coupled nonbi-anisotropic split ring resonator (BC-NB-SRR) ($\lambda_g/29$), broadside coupled spiral resonator with two turns [BC-SR(2)] ($\lambda_g/49$), and broadside coupled spiral resonator with four turns [BC-SR(4)] ($\lambda_g/117$). Figure 7 shows the topologies of these three resonators, the layout of the three metal layer configuration of a CPW loaded with a pair of metamaterial resonators and shunt strips, and the frequency response

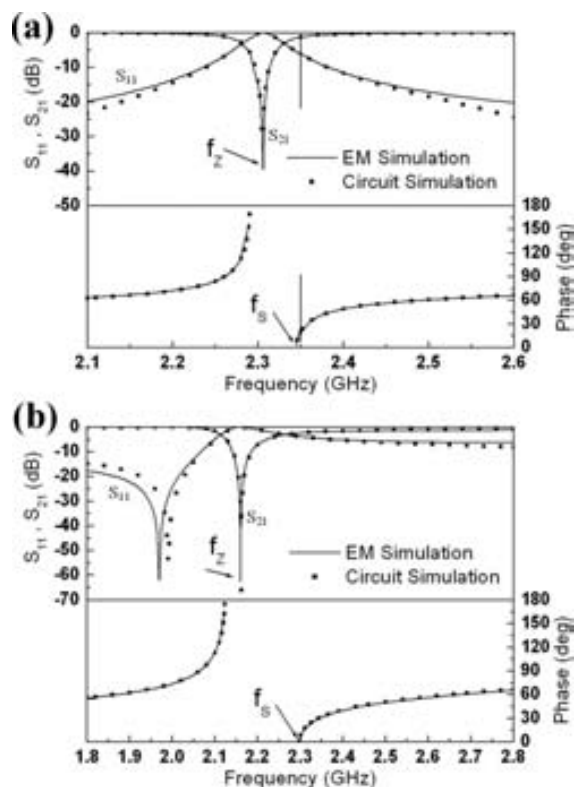


FIG. 6. Simulated reflection, S_{11} , and transmission, S_{21} , coefficients, and simulated dispersion relation for negative-permeability cells based on microstrip (a) and CPW (b) structures. The layouts are represented in Fig. 1 removing the shunt elements. The considered substrate and the relevant dimensions are the same as those in Figs. 4 and 5.

obtained from the electromagnetic simulation together with the electrical response corresponding to the circuit obtained from the parameter extraction. Both responses are identical, as what occurs in Fig. 5 with the SRR with a length $\lambda_g/10$. It should be taken into account that for these resonators with two metal layers, three metal layers are necessary: two for the resonator and one more for the coplanar line. The circuit parameters are shown in Table II. With these results, it is demonstrated that the parameter extraction method is suitable for any subwavelength resonator.

Thus, we can conclude that the circuit model shown in Fig. 2 appropriately describes the behavior of left handed lines loaded with magnetically coupled resonators. The proposed parameter extraction method is simple and useful for the characterization of resonant type metamaterial transmission lines.

III. DISCUSSION

The accurate determination of the parameters of the isolated particles (that is, without the presence of the CPW metal level) from the simulation of the structure proposed in Fig. 1(b) and applying the parameter extraction proposed in Sec. II is not possible since, to this end, the value of the mutual coupling M is required, and the assumption $M=Lf$ (f being the fraction of the slots occupied by the resonators) is

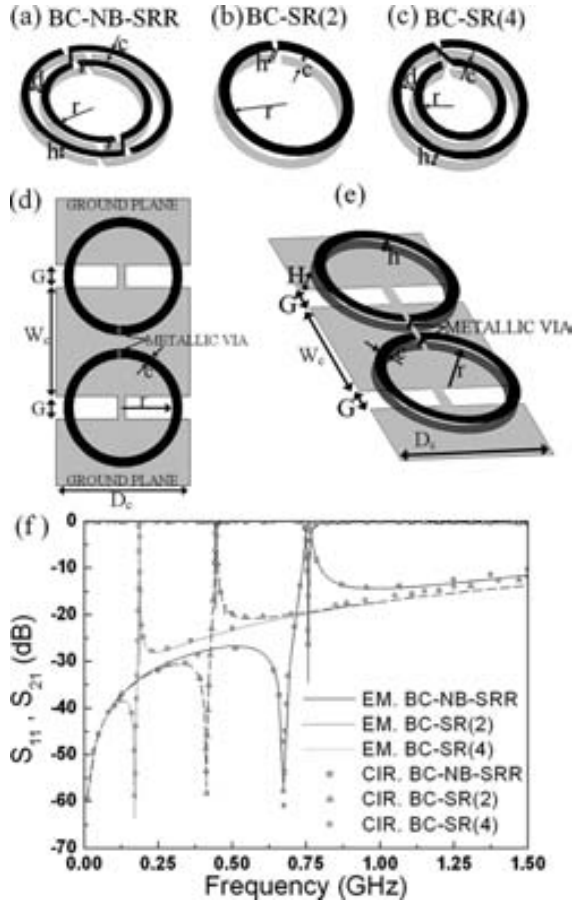


FIG. 7. Topologies of the different resonators considered: BC-NB-SRR (a), BC-SR(2) (b), and BC-SR(4) (c). Layout of a three metal layer configuration of a CPW loaded with a pair of metamaterial resonators and shunt strips: top (d) and three dimensional (e) views. Frequency responses (f) for three different resonators: BC-NB-SRR, BC-SR(2), and BC-SR(4) coupled to CPW structure with shunt strips. The considered substrate is Rogers RO3010 with dielectric constant $\epsilon_r=10.2$. The relevant dimensions for the resonators are ring width $c=0.6$ mm; distance between the rings, $d=0.2$ mm; and internal radius $r=2.4$ mm for the BC-NB-SRR and the BC-SR(4) and $r=3.2$ mm for the BC-SR(2). The substrate thickness between the resonator metallic layers is $h=0.635$ mm for the BC-NB-SRR and $h=0.127$ mm for the BC-SR(2) and the BC-SR(4). For the CPW structure the central strip width is $W_c=7$ mm, the width of the slots is $G=1.35$ mm, and the length is $D_c=8.6$ mm. The shunt strip width is $w_s=0.4$ mm; the substrate thickness between the CPW and the resonators is $H=0.635$ mm for the BC-NB-SRR and $H=1.143$ mm for the BC-SR(2) and the BC-SR(4). The results of the electrical simulation with extracted parameters are depicted using symbols.

only an approximation. Indeed, this assumption is valid as long as the following conditions are fulfilled: (i) the resonant elements can be described through a quasistatic analysis and (ii) the magnetic flux lines generated by the line in the region of the resonators penetrate the whole resonator area. As long as the resonators are electrically small, the first condition is reasonable, but it is not likely that the second condition applies for relatively thick dielectric layers (as those of many low loss commercial microwave substrates).

Let us assume however that M is known. In that case, the parameters of the model in Fig. 3 can be determined

TABLE II. Extracted parameters for the circuit shown in Fig. 2 with a CPW structure using different pairs of coupled resonators.

	C (pF)	L' (nH)	L'_p (nH)	C'_s (pF)	L'_s (nH)
BC-NB-SRR	3.37	4.01	2.08	25.80	2.16
BC-SR(2)	2.00	4.71	2.21	102.03	1.46
BC-SR(4)	2.20	4.39	2.20	545.11	1.59

through transformations (5)–(8). However, notice that L_s and C_s are the resonator parameters influenced by the presence of the metallic CPW structure on the opposite substrate side, which are different from the parameters of the isolated particles, L_l and C_l . L_l can be inferred from L_s by means of the method of images according to

$$L_l = L_s + M', \quad (14)$$

where M' is the mutual coupling between the resonator and its image, and it has been considered that the effect of the slots, in the case of the CPW, is negligible. M' is obtained by means of the well known Neumann's formula, where infinitely thin metal loops have been considered, with the radius corresponding to the average radius of the used resonator.

The determination of C_l from C_s is more complicated, but for relatively thick substrates (as compared to the distance between concentric rings, c) it is expected that $C_l \approx C_s$. A key point is if there exists a range of substrate thicknesses for which $C_l \approx C_s$ and the assumption $M=Lf$ is valid. To this end, we have calculated for four different particles the value of mutual coupling M_l that makes $C_l \approx C_s$ (C_l and L_l can be calculated analytically for the considered particles), and we have calculated the ratio M_l/M (normalized coupling). Notice that the normalized coupling coincides with f_l/f , where $f_l=M_l/L$. We may call f_l/f the normalized slot fraction. Such normalized slot fraction is depicted in Fig. 8 for different substrate thicknesses. The considered particles

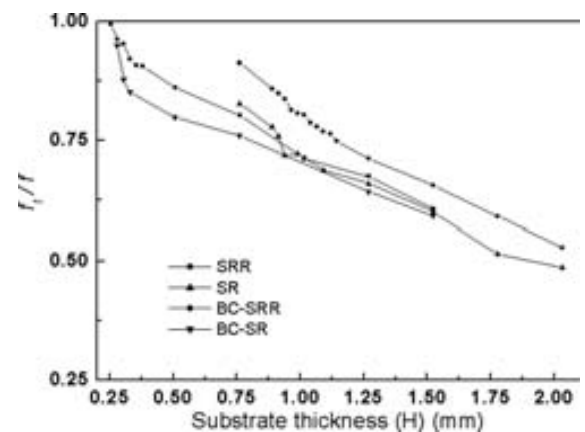


FIG. 8. Normalized slot fractions (f_l/f) that make $C_l=C_s$ for four different resonant particles and different substrate thicknesses H . The relevant dimensions for the resonators are ring width $c=0.6$ mm, distance $d=0.2$ mm between the rings, and internal radius $r=2.4$ mm for the SRR and the SR; and $c=0.6$ mm, $r=3.2$ mm, and substrate thickness $h=0.127$ mm between layers for the BC-SRR and the BC-SR(2).

TABLE III. Extracted parameters for the SRR-loaded microstrip and CPW unit cells with and without shunt strips (second and third columns) and isolated parameters (fifth and sixth columns). Modified mutual inductances (first column) and mutual couplings between the resonator and its image (fourth column).

Structure	M_I (nH)	$C_s, C_I, C_{I\text{mod}}$ (pF)	L_s (nH)	M' (nH)	$L_I=L_s+M'$ (nH)	$L_{I\text{mod}}$ (nH)
With the shunt metallic elements						
Microstrip	0.70	0.50	9.58	2.1	11.68	12.55
CPW	1.13	0.50	10.43	2.1	12.53	12.55
Removing the shunt metallic elements						
Microstrip	0.71	0.50	9.69	2.1	11.79	12.55
CPW	1.17	0.50	10.41	2.1	12.51	12.55

have the same external radius. Such particles are SRRs, SRs, BC-SRR, and BC-SR(2). As can be seen, the normalized slot fraction decreases with substrate thickness and it is the same for all the considered resonators (that corresponding to the SRR is slightly above the others for unknown reasons). That is, as the thickness increases, the fraction of the magnetic flux lines generated by the line that penetrates the resonator area decreases and $M_I < M$ ($f_I < f$). Except for two metal layer resonators with significant intermetallic distance, the normalized slot fraction is expected to depend essentially on the substrate thickness, as Fig. 8 demonstrates. However, notice that as substrate thickness is reduced, rather than a saturation of the normalized slot fraction (that is, $f_I/f \rightarrow 1$), this normalized fraction rapidly increases. This occurs for substrate thicknesses below approximately 0.50 mm, where the assumption that $C_I \approx C_s$ fails, so that neither M_I nor f_I have a physical meaning. Thus, the presented results show that there is not a region where the actual coupling M_I coincides with $M=Lf$ and, simultaneously, $C_I \approx C_s$. This precludes the accurate determination of the parameters of the isolated particles C_I and L_I . However, we can estimate these parameters by considering a test structure consisting on a CPW with thickness above 1.5 mm and use the normalized slot fraction derived from Fig. 8 (the average value). With these values, the mutual coupling M_I can be inferred and by introducing it in Eqs. (5)–(8), we can infer L_s and C_s and finally the parameters of the isolated particle, given by Eq. (14) and $C_I = C_s$.

To confirm the coherence between the extracted parameters (circuit in Fig. 2) and those of the isolated particles, we have considered the CPW and microstrip structures in Fig. 1, where we have extracted the parameters and we have calculated the mutual coupling M_I necessary to obtain $C_I = C_s$ (the substrate thickness is large enough). Actually this has been done also by removing the shunt connected inductances (strips or vias). The values of M_I , C_s , L_s , M' , C_I , and L_I are given in Table III, together with the analytical values of C_I and L_I (obtained from reported models^{8,19,30} and termed as $C_{I\text{mod}}$ and $L_{I\text{mod}}$ in Table III). As can be seen the analytical ($L_{I\text{mod}}$) and calculated (L_I) resonator inductances are in reasonable agreement (obviously $C_s = C_I$ since we have forced this). To further support the previous analysis, we have considered the four resonators in Fig. 8, each one loading a

TABLE IV. Extracted parameters for the SRR-, SR-, BC-SRR-, and BC-SR-loaded CPW unit cells without shunt strips for two substrate thicknesses.

	C (pF)	L (nH)	M (nH)	$C_s, C_I, C_{I\text{mod}}$ (pF)	L_s (nH)	M' (nH)	$L_I=L_s+M'$ (nH)	$L_{I\text{mod}}$ (nH)
Substrate thickness $H=1.27$ mm								
SRR	2.38	2.27	1.17	0.50	10.41	2.1	12.51	12.55
SR	2.03	2.49	1.19	2.00	9.96	2.1	12.06	12.55
BC-SRR	2.00	2.71	1.38	2.62	13.33	2.5	15.83	14.70
BC-SR	1.95	2.68	1.36	10.48	13.14	2.5	15.64	14.70
Substrate thickness $H=1.78$ mm								
SRR	2.05	2.32	1.00	0.50	10.53	1.3	11.83	12.55
SR	1.80	2.76	1.03	2.00	9.83	1.3	11.13	12.55
BC-SRR	1.85	2.51	1.15	2.62	13.49	1.7	15.19	14.70
BC-SR	1.78	3.50	1.34	10.48	13.50	1.7	15.20	14.70

CPW structure without strips, and we have obtained the same parameters as in Table III. We have done this for two different thicknesses. The results are shown in Table IV. Again, there is reasonable agreement between $L_{I\text{mod}}$ and L_I . (Indeed, taking into account that the width of the resonators is relatively wide— $c=0.6$ mm—some deviation between the calculation and the analytical model can be expected.)

The reported parameter extraction technique for the circuit in Fig. 2 provides results that are coherent with the parameters of the resonators analytically obtained. The parameter extraction technique is very valuable and valid for complex resonators which cannot be easily modeled, such as the ones presented in Ref. 29. We would like to mention that the resonator parameters can also be extracted from experimental data. However, losses may obscure the results and the comparison to the analytical models. For this reason this study has been restricted to frequency responses inferred from electromagnetic simulation. Nevertheless, it is well known that the electromagnetic software that has been used (AGILENT MOMENTUM) predicts the behavior of the resonators with good accuracy, as has been previously demonstrated, thus supporting the validity of this study.

IV. CONCLUSIONS

In conclusion, we have proposed a technique for the determination of resonator parameters when they are coupled to a planar transmission line. The technique has been applied to several resonant particles with different electrical lengths and different structures in microstrip and CPW configurations. We have discussed the possibility of obtaining the parameters of the isolated particles from the extracted parameters, and we have concluded that this cannot be done accurately, although we can obtain a reasonable estimation. Nevertheless, we have corroborated that the extracted parameters are coherent with the parameters of the isolated particles, and this has been done for different host lines, particles, and substrate thicknesses.

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Artículo G

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Resumen del artículo:

Este artículo de revisión se dedica a la discusión y la comparación de los modelos circuitales equivalentes de los diferentes tipos de líneas de transmisión metamaterial que han sido propuestos hasta la fecha, es decir, la conocida línea cargada con capacidades en serie e inductancias en derivación, y las líneas basadas en el modelo resonante. Esta última categoría comprende las líneas artificiales cargadas con *split ring resonators* (SRRs), o topologías relacionadas, y líneas de transmisión metamaterial basadas en *complementary split ring resonators* (CSRRs). Será objetivo principal de este artículo el justificar claramente los elementos circuitales de los modelos (y vincular estos elementos a la línea física y la topología), con el objetivo de comparar las diferentes líneas a la luz de estos modelos, así como señalar las ventajas y los inconvenientes de las distintas líneas de transmisión metamaterial. Dado que las líneas de transmisión metamaterial son ampliamente utilizadas para el diseño de componentes compactos de microondas y ondas milimétricas con unas mejores prestaciones y/o basados en nuevas funcionalidades y su síntesis se basa en el equivalente circuitual de elementos discretos, este artículo es de verdadero interés para los ingenieros de RF/microondas y, en general, para los lectores que participan en la investigación y las aplicaciones de los metamateriales.

Modelling metamaterial transmission lines: a review and recent developments

INVITED PAPER

F. AZNAR, M. GIL, J. BONACHE, and F. MARTÍN*

CIMITEC, Departament d'Enginyeria Electrònica, Universitat Autònoma de Barcelona,
08193 Bellaterra (Barcelona), Spain

This review paper is devoted to the discussion and comparison of the lumped element equivalent circuit models of the different types of metamaterial transmission lines that have been proposed so far, namely the CL-loaded lines, and those lines based on the resonant type approach. The latter category comprises both artificial lines loaded with split ring resonators (SRRs), or related topologies, and metamaterial transmission lines based on complementary split ring resonators (CSRRs). It will be the main aim of this paper to clearly justify the circuit elements of the models (and link such elements to the line physics and topology), to compare the different lines to the light of these models, and to point out the advantages and drawbacks of the different metamaterial transmission lines. As long as metamaterial transmission lines are exhaustively used for the design of compact microwave and millimeter wave components with improved performance and/or based on new functionalities, and their synthesis is based on the lumped element equivalent circuits, this paper is of actual interest for RF/microwave engineers and in general to those readers involved in metamaterial research and applications.

Keywords: metamaterials, transmission line, microstrip technology, coplanar wave guide technology, split ring resonator, complementary split ring resonator.

1. Introduction

The concept of metamaterial transmission line was introduced for the first time in 2002 [1–3] to make reference to one-dimensional metamaterials consisting on a host transmission line loaded with reactive elements, and exhibiting a negative effective permeability and permittivity in a certain frequency band. The properties of substances with negative permeability and permittivity were already predicted by Veselago at the end of the sixties [4]. However, it was not until 2000 that such media (called left handed media, negative refractive index media or double negative media) were synthesized [5]. It was achieved by Smith and co-workers by combining metallic posts and split ring resonators (SRRs) in a periodic array of constitutive unit cells or “atoms”. The posts provide a negative effective permittivity up to a cut-off (plasma) frequency (which depends on the post radius and the distance to the adjacent posts). The SRRs (proposed by Pendry in 1999 [6]) are responsible for the negative permeability (in a certain band above resonance), which is achieved by illuminating the structure with a magnetic field parallel to the SRRs axis (the electric field must be parallel to the posts). Actually, the first artificial left handed medium [5], although bulky (volumetric), it was neither three dimensional nor isotropic (a two dimensional implementation was subsequently provided by

Smith *et al.* [7]). These former media are thus artificial materials with controllable electromagnetic properties (permittivity and permeability), that is, metamaterials, although they are highly anisotropic.

Conversely, metamaterial transmission lines are not media (in the usual sense). These artificial lines have been termed this way as long as they can be designed to exhibit backward wave propagation (as synonymous of left handed wave propagation), thus mimicking wave propagation in left handed media. In spite of the fact that metamaterial transmission lines are not media, we can also consider that such lines exhibit a certain (controllable) effective permeability and permittivity. This can be done through a mapping between the line parameters (that is the parameters of the equivalent circuit model of the artificial lines) and the constitutive parameters (permeability and permittivity) of media, since plane wave propagation in isotropic media is described by the same equation that TEM wave propagation in transmission lines (telegraphist’s equation) [8].

Another important aspect of metamaterials is that, in order to define an effective permeability and permittivity, it is necessary that the structure is “seen” as a homogeneous (continuous) media. This means that the dimensions of the unit cells must be small as compared to the wavelength at the frequencies of interest. According to this, to talk in terms of effective permeability or permittivity in metamaterial transmission lines, has only sense as long as the homogeneity conditions are fulfilled. However, in metama-

*e-mail: Ferran.Martin@uab.es

terial transmission lines the relevant parameters are the characteristic impedance and the electrical length (phase shift) of the lines, rather than the effective permeability and permittivity. Thus, the homogeneity condition is not actually necessary to control the impedance and the phase. Dispersion is inherently present in metamaterial transmission lines, and the characteristic impedance is dependent on frequency. Although the design of homogeneous metamaterial transmission lines usually provides roughly constant line impedance in the vicinity of the operating frequency, where the line can be considered to be homogeneous, the impedance varies considerably at frequencies where the guided wavelength is comparable to the cell dimensions (i.e., in that region where the structure cannot be considered to be homogeneous). Thus, homogeneity in metamaterial transmission lines (and in metamaterials in general) means achievement of the homogeneity conditions in the vicinity of the operating (or design) frequency. For the design of metamaterial-based circuits and components, homogeneity is irrelevant. Indeed, the required impedance and phase shift at the design frequency can be usually achieved through a single unit cell artificial line. This will normally provide a narrower bandwidth for impedance, but final dimensions can also be reduced. According to this comments, we will not restrict the definition of metamaterial transmission lines to those lines exhibiting homogeneity. In the present work, metamaterial transmission lines are considered to be those artificial lines loaded with reactive elements, whose phase and impedance can be controlled. Notice that according to this definition, metamaterial transmission lines can exhibit left handed wave propagation (equivalent to negative effective permeability and permittivity), or right handed wave propagation (equivalent to positive effective permeability and permittivity). It is also possible to implement artificial lines exhibiting stop band behaviour in a certain frequency band, related to the arrangement of unit cells providing a negative effective permeability or permittivity in that band. Such lines will be called negative permeability or negative permittivity lines.

2. CL-loaded and resonant type metamaterial transmission lines: topologies and circuit models

2.1. CL-loaded metamaterial transmission lines

CL-loaded metamaterial transmission lines consist of transmission lines loaded with series capacitances and shunt inductances. Such lines can be implemented through lumped circuit elements (smd inductances and capacitances) or by means of semi-lumped components. Semi-lumped components mean, in this context, electrically small planar components. Through semi-lumped components, fully planar configurations can be obtained, although the values of capacitances and inductances that can be implemented are limited. Typical topologies of fully planar CL-loaded lines

are depicted in Fig. 1 [9,10]. In Fig. 1(a), a coplanar waveguide (CPW) transmission line is periodically loaded with shunt connected strips (emulating the shunt inductances) and gaps (accounting for the series capacitances). In Fig. 1(b), where a microstrip line is considered, the shunt strips are replaced with vias. Finally, in Fig. 1(c), a microstrip line is loaded with interdigital capacitances series connected and grounded stubs (acting as shunt inductors). In all cases, the lumped element equivalent circuit model of the unit cell is that depicted in Fig. 2 (this model is valid as long as the distance between the semi-lumped elements is small). Losses are considered to be negligible and hence they are not included in the circuit model. The elements of the model are the line parameters (capacitance C_R and inductance L_R), the series capacitance C_L , and the shunt inductance, L_L . As usual, the sub-index denote the elements responsible for the left handed (L) and right handed (R) bands of these artificial lines. Namely, at low frequencies, the loading elements are dominant and left handed wave propagation arises in a certain frequency band. At higher frequencies, the loading elements are no longer dominant and the wave propagation is forward. The typical dispersion diagram (ω - β) of such lines can be inferred from the following expression

$$\cos \beta l = 1 + \frac{Z_s(\omega)}{Z_p(\omega)}, \quad (1)$$

where β is the phase constant, l is the unit cell length, and $Z_s(\omega)$ and $Z_p(\omega)$ are the series and shunt impedances of the T-circuit model. Inspection of this dispersion diagram (Fig. 3) reveals that the group velocity is positive in the al-

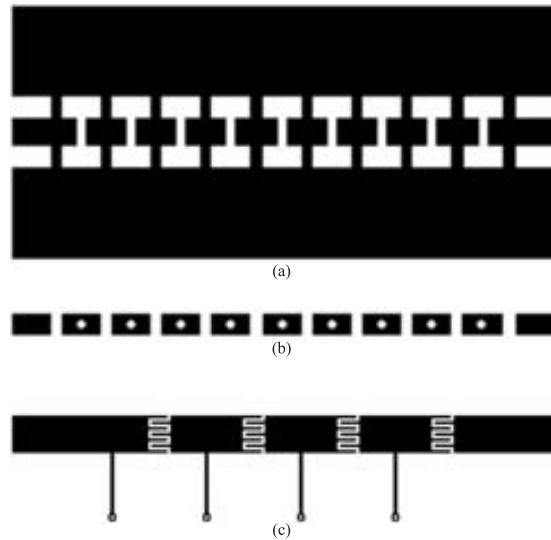


Fig. 1. Typical topologies of CL-loaded metamaterial transmission lines: (a) CPW structure loaded with shunt strips and series gaps, (b) microstrip structure loaded with vias and series gaps, and (c) microstrip structure loaded with grounded stubs and interdigital capacitors.

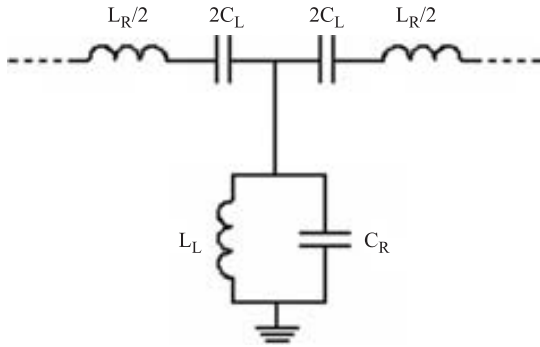


Fig. 2. Lumped element equivalent T-circuit model of the unit cell of CL-loaded metamaterial transmission lines.

lowed bands, whereas the phase velocity is negative (backward waves) in the left handed band, and positive (forward waves) in the right handed band. Thus, CL-loaded lines do actually exhibit a composite right/left handed (CRLH) behaviour [11]. To obtain a purely left handed line, we would need a cascade of series capacitances alternating with shunt inductances. This corresponds to the dual model of a conventional transmission line, which is well known to exhibit backward waves above a certain cut-off frequency [12]. However, such line cannot be implemented in practice since a host line is required. The frequency gap present between the left handed and the right handed bands is delimited by the following frequencies (Fig. 3)

$$\omega_{G1} = \min(\omega_s, \omega_p), \quad (2)$$

$$\omega_{G2} = \max(\omega_s, \omega_p), \quad (3)$$

with

$$\omega_s = \frac{1}{\sqrt{L_R C_L}}, \quad (4)$$

$$\omega_p = \frac{1}{\sqrt{L_L C_R}}, \quad (5)$$

and the lower (ω_-) and upper (ω_+) limits of the left handed and right handed bands, respectively, are

$$\omega_{\pm} = \sqrt{\frac{4 + L_R C_R (\omega_s^2 + \omega_p^2)}{2 L_R C_R}} \pm \frac{1}{2 L_R C_R} \sqrt{L_R^2 C_R^2 (\omega_s^2 - \omega_p^2)^2 + 8 L_R C_R (\omega_s^2 + \omega_p^2) + 16}. \quad (6)$$

By designing the structure with identical series and shunt resonance frequencies, the gap disappears and there is a continuous transition between the left handed and right handed bands (balance condition). At the transition frequency, the phase velocity is infinity, whereas the group velocity is finite (Fig. 4). The implications of this (out of the scope of this work) have been discussed in detail in Refs. 13 and 14. Nevertheless, it is interesting to mention that the characteristic impedance (or image impedance) of these lines, given by the following general expression

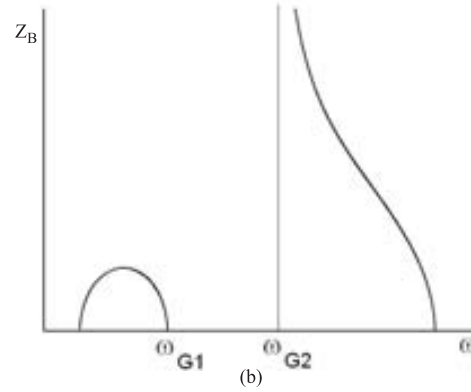
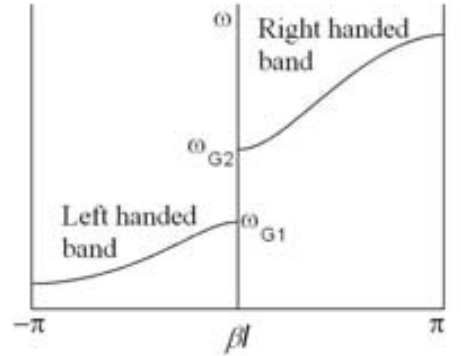


Fig. 3. Typical dispersion diagram (a) and characteristic impedance (b) of a CL-loaded metamaterial transmission line. The line exhibits a composite right/left handed (CRLH) behaviour.

$$Z_B = \sqrt{Z_s(\omega)[Z_s(\omega) + 2Z_p(\omega)]}, \quad (7)$$

takes this form

$$Z_B = \sqrt{\frac{L_R \left(1 - \frac{\omega_s^2}{\omega^2}\right)}{C_R \left(1 - \frac{\omega_p^2}{\omega^2}\right)} - \frac{L_R^2 \omega^2}{4} \left(1 - \frac{\omega_s^2}{\omega^2}\right)^2}, \quad (8)$$

for CRLH lines based on the model depicted in Fig. 2 [see Fig. 3(b) for a typical representation of the dependence of Z_B with frequency], and this expression reduces to

$$Z_B = \sqrt{\frac{L_R}{C_R} - \frac{L_R^2 \omega^2}{4} \left(1 - \frac{\omega_s^2}{\omega^2}\right)^2}, \quad (9)$$

for balanced lines. In this latter case (balanced lines) the impedance is maximum and roughly constant in the vicinity of the transition frequency [13,14] (see Fig. 4).

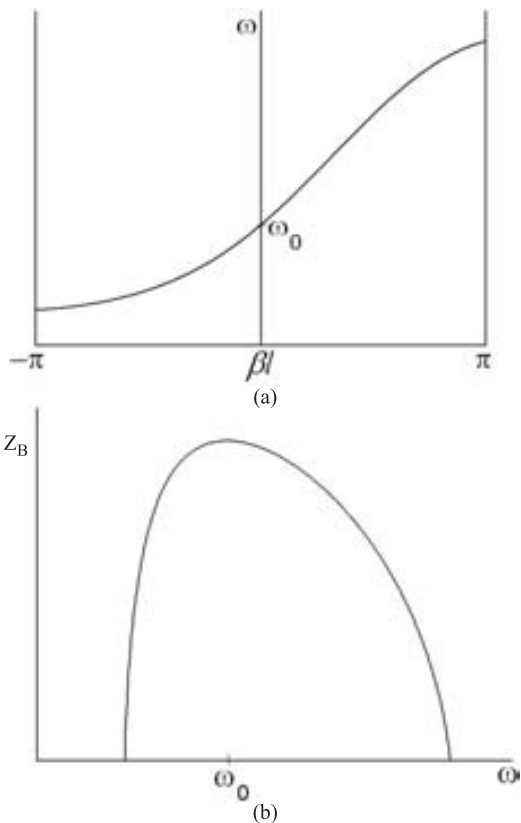


Fig. 4. Typical dispersion diagram (a) and characteristic impedance (b) of a balanced CRLH CL-loaded metamaterial transmission line.

2.2. Resonant-type metamaterial transmission lines

Let us now consider the resonant type approach of metamaterial transmission lines. These lines can be implemented either by means of SRRs [15] or through their complementary counterparts (CSRRs) [16,17]. In the former case, the typical topology (unit cell) is depicted in Fig. 5. It consists of a CPW with SRRs etched on the bottom side of the substrate, below the slots of the line, and shunt connected strips. As in Fig. 1(a), the shunt strips provide the negative effective permittivity, whereas the SRRs are responsible for the negative effective permeability in a narrow band above their resonance frequency. As long as the plasma frequency of the CPW/strips is substantially higher than the resonance frequency of the SRRs, left handed wave propagation arises in a certain band. Alternatively, left handed wave propagation can be interpreted to the light of the lumped element equivalent circuit model of the structure. The first proposed model for the unit cell of these lines was reported in Ref. 15, Fig. 6(a). According to this model, the SRRs are described by the resonant tank, L_s - C_s , the magnetic coupling between the line and the SRRs is accounted for through the mutual inductance M , L , and C are the inductance and capacitance of the line, respectively,

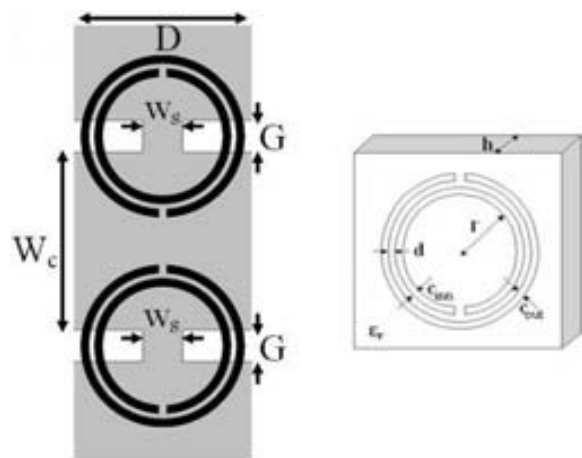


Fig. 5. Typical unit cell and relevant dimensions of a CPW metamaterial transmission line loaded with SRRs and shunt connected strips. SRRs, depicted in black, are etched in the back substrate side.

and, finally, L_p is the inductance of the shunt connected strips. This model is also valid for other SRR-based configurations implemented in microstrip technology [18]. As reported in Refs. 15 and 19, the model can be easily transformed to that depicted in Fig. 6(c). One of the objectives of the present paper is to discuss the validity of this model and compare it to the model of CL loaded CRLH lines (the models of the structure without shunt connected strips are depicted in Figs. 6(b) and 6(d)). However, before that, let us consider the resonant type metamaterial transmission lines based on CSRRs.

The typical topology of CSRR-based metamaterial transmission lines (unit cell) is depicted in Fig. 7. It is a

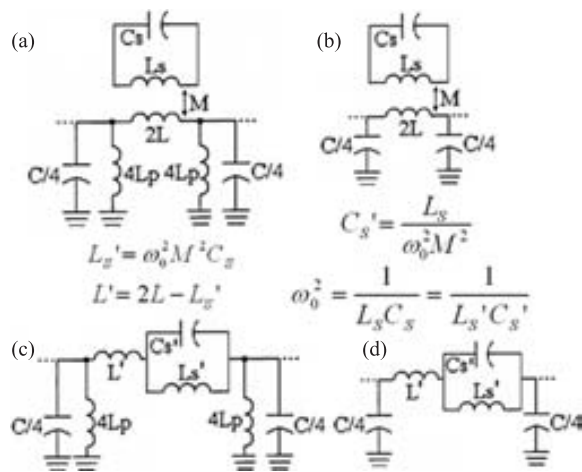


Fig. 6. Formerly proposed lumped element equivalent circuit for the basic cell of the left handed (a) and negative permeability (b) CPW structures loaded with SRRs. These circuits can be transformed to those depicted in (c) and (d), according to the indicated transformations (after Ref. 15).

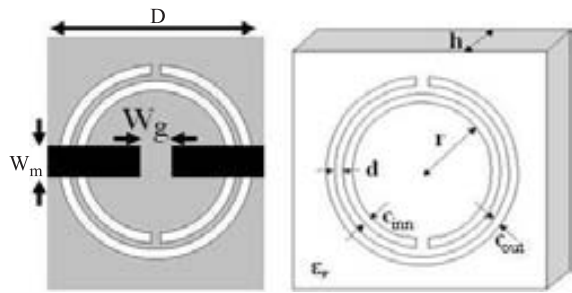


Fig. 7. Typical layout (unit cell) and relevant dimensions of a microstrip metamaterial transmission line loaded with CSRRs and series gaps. Ground plane metal is depicted in grey.

microstrip line with a series gap and a CSRR etched in the ground plane, below the position of the gap. The CSRRs, provide the negative effective permittivity in a certain band in the vicinity of their resonance frequency, provided they are excited by means of an axial electric field [16,17]. The negative effective permeability is provided by the series gaps. The lumped element equivalent circuit model of the unit cell of such lines was reported in Ref. 19 (see Fig. 8). In this T-circuit model, the CSRRs are modelled by the resonant tank L_c-C_c , the line inductance is L , the series gap is modelled by C_g and C account for the electric coupling between the line and the CSRRs.

Let us now focus on the validity of the circuit models for the description of SRR- and CSRR-based metamaterial transmission lines. These circuit models are valid as long as they are able to properly describe the structures and the element parameters are representative of the different components (i.e., they have the given interpretation). To this end, we have developed a parameter extraction method for SRR- and for CSRR-based lines. The relevant ideas of this technique have been published in Ref. 20 for CSRR-loaded lines (for lines loaded with SRRs, the technique can be easily extended). Figure 9 shows the frequency responses of 4 different structures (unit cells) that have been obtained by means of electromagnetic simulation (using the Agilent Momentum commercial software).

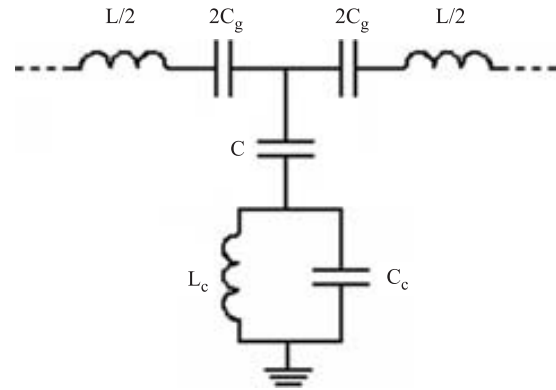


Fig. 8. Formerly proposed lumped element equivalent T-circuit model of the unit cell of left handed lines loaded with CSRRs and series gaps (after Figs. 19 and 20).

The structures are four artificial lines. The first one [Fig. 9(a)] is a microstrip line loaded with CSRRs and a series gap; the second one [Fig. 9(b)] is the same structure, but without the presence of the series gap; the third one [Fig. 9(c)] is a CPW loaded with a SRR pair and two shunt strips; finally, Fig. 9(d) depicts the CPW without the presence of the strips. Also depicted in this figure are the frequency responses that have been obtained through circuit simulation of the lumped element models with the extracted parameters (represented in Table 1).

As it can be appreciated, the fit is good. However, there is a noticeable aspect that requires some discussion. Namely, for the two CSRR-loaded structures (with and without series gap), the transmission zero

$$f_z = \frac{1}{2\pi\sqrt{L_c(C+C_c)}}, \quad (10)$$

varies substantially. The effect is due to a variation of the coupling capacitance, C (as revealed in Table 1). In the SRR-loaded structures (with and without shunt connected strips) a similar behaviour results, that is, the transmission zero is substantially modified, this time being due to a variation of the parallel resonant tank series connected in the

Table 1. Extracted element parameters for the structures used in Fig. 9.

Microstrip line loaded with CSRRs					
	C (pF)	L (nH)	C_g (pF)	C_c (pF)	L_c (nH)
With series gap	6.43	5.78	0.31	3.10	1.49
Without series gap	0.95	5.78	–	2.81	1.65
CPW loaded with SRRs					
	C (pF)	L' (nH)	L_p' (nH)	C_s' (pF)	L_s' (nH)
With shunt strips	2.07	4.91	1.38	3.90	1.68
Without shunt strips	2.07	1.91	–	23.01	0.24

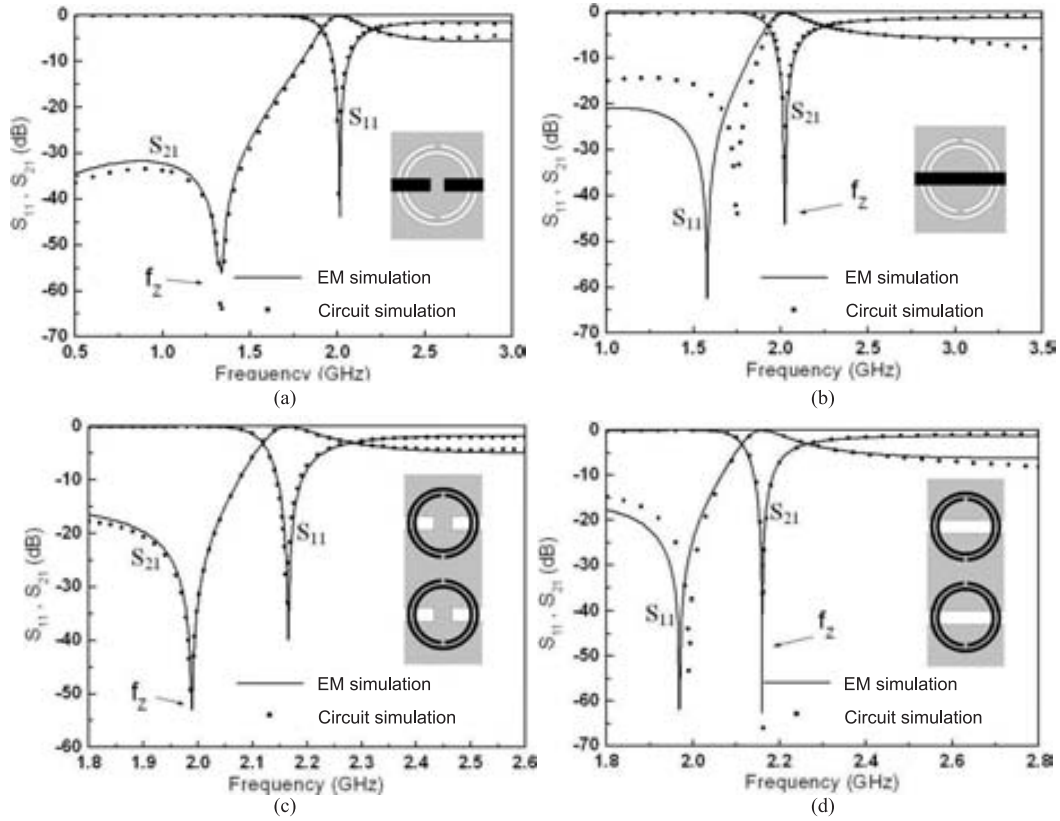


Fig. 9. Simulated (through the Agilent Momentum commercial software) frequency responses of the unit cell structures shown in the insets: (a) microstrip line loaded with CSRRs and series gaps, (b) microstrip line only loaded with CSRRs, (c) CPW loaded with SRRs and shunt connected strips, and (d) CPW only loaded with SRRs. The response that has been obtained from circuit simulation of the equivalent model with extracted parameters is also included. For the structures (a) and (b), the dimensions are: the strip line width $W_m = 1.15$ mm, the length $D = 8$ mm, and the gap width $w_g = 0.16$ mm. For the structures of (c) and (d), the dimensions are: the central strip width $W_c = 8$ mm, the width of the slots $G = 1.43$ mm, length of the line $D = 8$ mm, and the shunt strip width $w_s = 1.8$ mm. In all cases, the dimensions for the resonators (CSRRs and SRRs) are: the outer ring width $c_{out} = 0.364$ mm, the inner ring width $c_{inn} = 0.366$ mm, distance between the rings $d = 0.24$ mm, and the internal radius $r = 2.691$ mm. The considered substrate is Rogers RO3010 with the dielectric constant $\epsilon_r = 10.2$ and the thickness $h = 1.27$ mm.

π -circuit model. This is obvious since the transmission zero is now given by

$$f_z = \frac{1}{2\pi\sqrt{L_s C_s}}. \quad (11)$$

In the CSRR-loaded lines, the enhancement of the coupling capacitance in the structure with series gap has been attributed to the fringing capacitance [20]. However, the variation seems to be excessive to be attributed to this fact, and this aspect has been investigated by the authors of Refs. 21 and 22. The solution is simply that the circuit model of Fig. 8 is valid (it provides an excellent fit to the electromagnetic simulation or experiment), but not all the parameters of this circuit model have the given physical interpretation. Namely, neither C_g is actually the capacitance of the series gap, nor C is its fringing capacitance. A more realistic model of the structure is that depicted in Fig. 10, where C_L is the line capacitance, C_f is the fringing capaci-

tance of the gap and C_s is the series capacitance of the gap. Obviously, from π -T transformation, the circuit model of Fig. 8, which is the previously reported model of microstrip lines loaded with CSRRs and series gaps, is obtained, but the values of C_g and C do not actually have a physical interpretation. Indeed, C_g and C can be expressed in terms of C_s and $C_{par} = C_f + C_L$ according to

$$C_g = 2C_s + C_{par}, \quad (12)$$

$$C = \frac{C_{par}(2C_s + C_{par})}{C_s}. \quad (13)$$

Inspection of Eq. (13) indicates that as C_s decreases (gap distance increases), C increases. Indeed C can be made very large if C_s is sufficiently small. It is also apparent from Eq. (12) that in spite of the fact that the gap distance is increased, it is not expected that C_g experiences a significant reduction. The reason is that for small values of C_s , C_g is dominated by the line capacitance and the fring-

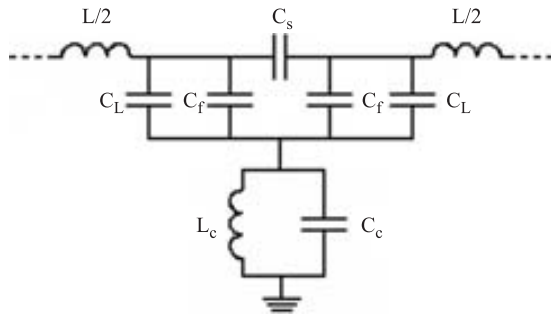


Fig. 10. Improved lumped element equivalent circuit model (unit cell) for the left handed lines loaded with CSRRs.

ing capacitance (C_{par}). Such behaviour of C and C_g has been recently corroborated by the authors by considering several CSRR-loaded structures with identical geometry except in the gap distance [21]. This analysis reveals that C can be enhanced by merely increasing the gap distance, without the penalty of a small C_g value (since C_f and C_L do also contribute to this capacitance, see Eq. (12)). These results are of interest because they reveal that it is possible to

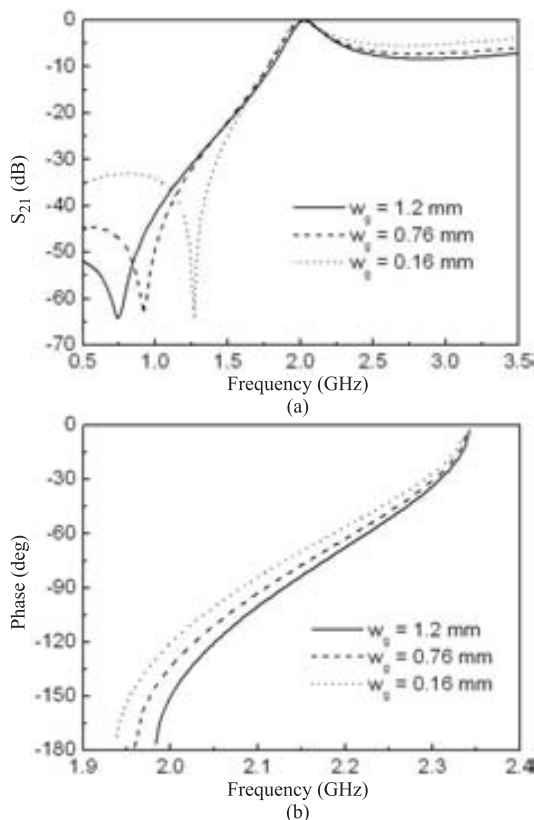


Fig. 11. Transmission coefficient (a) and phase (b) of different CSRR-loaded structures with different gap separation (indicated in the figure). The phase nulls at identical frequency in all the considered cases. Dimensions are the same that in Fig. 9.

obtain high values of C (by decreasing C_s), regardless of the substrate thickness. These high values of C are typically necessary to enhance bandwidth and to drive the transmission zero frequency to small values.

We would also like to highlight that in these CSRR-based structures, the frequency where the phase shift is null ($\phi = \beta l = 0$), ω_s does not depend on the characteristics of the gap, as has been demonstrated from full wave electromagnetic simulation of several structures with different gaps. The results of the electrical length are depicted in Fig. 11. Indeed, this behaviour is consistent with the improved circuit model of CSRR-loaded lines (Fig. 10), and hence it supports the validity of this model. Namely, the phase shift nulls at that frequency that opens the shunt impedance (see Eq. 1), and this occurs at the resonance frequency of CSRRs

$$f_z = \frac{1}{2\pi\sqrt{L_c C_c}} \quad (14)$$

Let us now focus on the SRR-based structures. It is obvious that the circuits of Figs. 6(a) and 6(b), reported for the first time in Ref. 15 by Martin *et al.*, cannot be simultaneously correct since they predict identical transmission zero frequency (provided identical SRRs are used). In Ref. 15, it was argued that the left handed band in a CPW loaded with 4 SRR and 4 shunt strip pairs was slightly situated to the right of the stop band measured on identical structure without the presence of the strips. But the greater number of stages obscures the position of the transmission zeros, as has been recently discussed by some of the authors [21,23]. The problem is that the CPW structure of Fig. 5 is not properly modelled by the circuit of Fig. 6(a). The shunt strips are not actually allocated at both sides of the SRRs, but just on the top of them. Thus, the circuit model (unit cell) of the CPW loaded with SRRs and shunt strips must be modelled as depicted in Fig. 12 [21,23,24]. However, this circuit model can be transformed to a model formally identical to that of Fig. 6(c), but with modified parameters [21,23]. That is the reason why the model of Fig. 6(c) perfectly fits the simulated (or measured) frequency responses of SRR/strip loaded CPWs. From this transformation, which involves a tedious calculation, the shift of the transmission zero towards lower frequencies as compared to the structure without shunt strips can perfectly be explained.

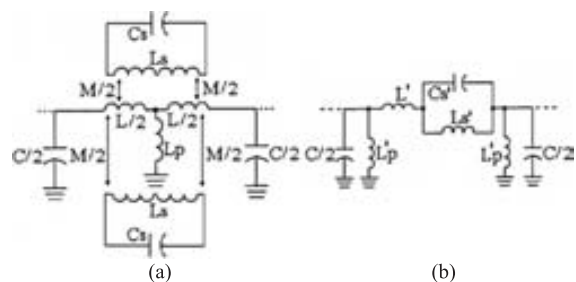


Fig. 12. Improved lumped element equivalent circuit model (unit cell) for the left handed lines loaded with SRRs.

Due to symmetry considerations and reciprocity, the admittance matrix of the circuit of Fig. 12(a) (which is a bi-port) must satisfy $Y_{12} = Y_{21}$ and $Y_{11} = Y_{22}$. From these matrix elements, the equivalent π -circuit model can be obtained according to [25]

$$Z_s(\omega) = -(Y_{21})^{-1}, \quad (15a)$$

$$Z_p(\omega) = (Y_{11} + Y_{21})^{-1}, \quad (15b)$$

Y_{21} is inferred by grounding the port 1 and obtaining the ratio between the current at port 1 and the applied voltage at port 2. Y_{11} is simply the input admittance of the bi-port, seen from port 1, with a short-circuit at port 2. After a straightforward but tedious calculation, the elements of the admittance matrix are obtained, and by applying Eq. (15), we finally obtain

$$Z_s(\omega) = \left(2 + \frac{L}{2L_p}\right) j\omega \left[\frac{L}{2} + M^2 \frac{1 + \frac{L}{4L_p}}{L_s \left(\frac{\omega_o^2}{\omega^2} - 1 \right) - \frac{M^2}{2L_p}} \right], \quad (16a)$$

$$Z_p(\omega) = j\omega \left(2L_p + \frac{L}{2}\right), \quad (16b)$$

with $\omega_o = (L_s C_s)^{-1/2}$. Equation (16a) can be re-written as

$$Z_s(\omega) = \left(2 + \frac{L}{2L_p}\right) j\omega \left[\frac{L}{2} - L'_s + \frac{L'_s}{1 - L'_s C'_s \omega^2} \right], \quad (17)$$

with

$$L'_s = 2M^2 C_s \omega_o^2 \frac{\left(1 + \frac{L}{4L_p}\right)^2}{1 + \frac{M^2}{2L_p L_s}}, \quad (18)$$

$$C'_s = \frac{L_s}{2M^2 \omega_o^2} \left(\frac{1 + \frac{M^2}{2L_p L_s}}{1 + \frac{L}{4L_p}} \right)^2. \quad (19)$$

These results indicate that the improved lumped element circuit model of the unit cell of the left handed lines loaded with SRRs and shunt inductors (Fig. 12) can be formally expressed as the π -circuit model of Fig. 6(c), but with modified parameters. These modified parameters are related to the parameters of the circuit of Fig. 12, according to Eqs. (18) and (19)

$$L' = \left(2 + \frac{L}{2L_p}\right) \frac{L}{2} - L'_s, \quad (20)$$

$$L'_p = 2L_p + \frac{L}{2}. \quad (21)$$

Inspection of these results reveals that the transmission zero frequency ω_z , for the circuit of Fig. 12 is no longer given by the resonance frequency of the SRRs ω_o , but it is smaller, that is, $\omega_z < \omega_o$.

It is also interesting to mention that for the circuit of Fig. 12, the frequency ω_s where $\phi = \beta l = 0$, which is obtained by forcing $Z_s(\omega) = 0$ (see Eq. 1)

$$\omega_s = \frac{1}{\sqrt{C_s \left(L_s - 2 \frac{M^2}{L} \right)}}, \quad (22)$$

does not depend on the shunt inductance, in spite that $Z_s(\omega)$ is a function of L_p . This has been corroborated through full wave electromagnetic simulation of identical left handed and negative permeability structures (with the exception of the absence of shunt inductive strips in the latter) [23] (see Fig. 13).

According to this analysis, the previous reported circuit model of left handed lines loaded with SRRs and shunt inductive elements [Fig. 6(c)] [15] is formally correct. The weakness relies on the physical interpretation of the elements of that model. These elements do not have any phys-

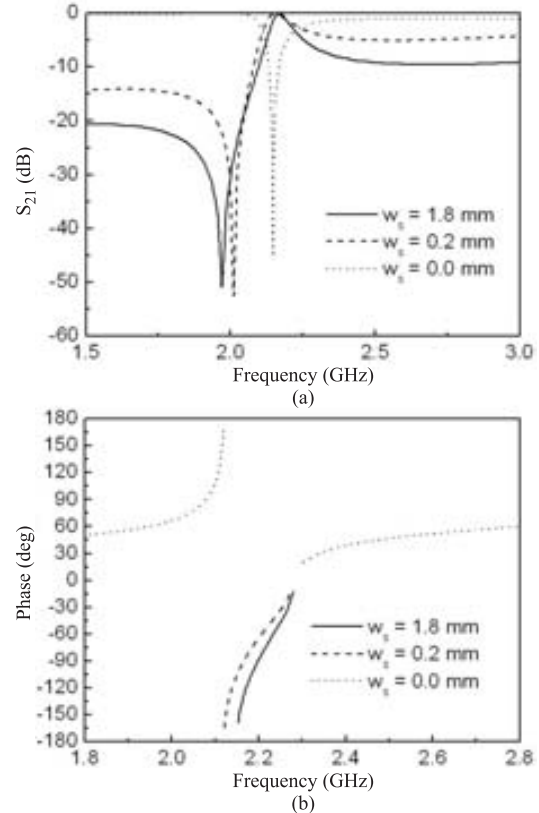


Fig. 13. Transmission coefficient (a) and phase (b) of different SRR-loaded structures with different shunt strip width (indicated in the figure). The phase nulls at identical frequency in all the considered cases. Dimensions are the same that in Fig. 9.

Table 2. Parameters for the circuit shown in Fig. 12(a), obtained from Table 1.

	C (pF)	L (nH)	M (nH)	L_p (nH)	C_s (pF)	L_s (nH)
With shunt strips	2.07	1.94	1.23	0.20	0.31	17.16
Without shunt strips	2.07	2.15	1.45	–	0.31	17.57

ical meaning. However, they are related to the elements of the circuit model of Fig. 12, which describe the different components of the left handed unit cell.

To further confirm the validity of the proposed circuit model of left handed lines, based on SRRs, we have inverted Eqs. (18)–(21) in order to obtain the parameters of the model of Fig. 12(a) (from the results shown in Table 1), and these coincide in very good agreement with those parameters inferred in the structure without shunt strips (see Table 2), and obtained through the transformation indicated in Fig. 6. This means that the presence of the shunt strips does not affect the parameters of the structure, hence having a clear physical interpretation.

2.3. Comparison between CL-loaded and resonant type metamaterial transmission lines

Inspection of the circuit models of CL-loaded and resonant type metamaterial transmission lines [Figs. 2, 8, and 6(c)] reveals that these lines must exhibit similar characteristics. Let us first compare the SRR- and CSRR loaded lines. It has been already shown in the last section that the phase and transmission characteristics are very similar. Indeed, SRR- and CSRR-loaded lines are roughly dual structures. They are not strictly electromagnetic duals, but the circuit models of these lines are formally circuit duals [26]. That is, the series/shunt impedance of any of these circuit models can be expressed as proportional to the admittance of the shunt/series branch of the other circuit. Thus, the similar behaviour is not actually surprising.

Let us now compare the CL-loaded lines with CSRR-based metamaterial transmission lines. The corresponding circuit models are identical with the exception of the presence of the coupling capacitance C in the model of the CSRR-loaded lines. This coupling capacitance is the responsible for a transmission zero at a finite frequency in CSRR-loaded lines (in CL-loaded lines such transmission zero is at the origin). But apart from this, the behaviour of both structures is very similar. CSRR-loaded lines do also exhibit CRLH behaviour [27]. The frequency gap present between the left handed and right handed bands (unbalanced line) is delimited by the resonance frequency of the series and shunt branches (as occurs in CL-loaded lines). In the balanced CSRR-based structures, the characteristic impedance is given by the following expression

$$Z_B = \sqrt{\frac{L}{C_c} - \frac{L^2 \omega^2}{4} \left(1 - \frac{\omega_o^2}{\omega^2}\right)^2} + \frac{L}{C} \left(1 - \frac{\omega_o^2}{\omega^2}\right), \quad (23)$$

which corresponds to an impedance dependence on frequency very similar to that depicted in Fig. 4. However, due to the presence of C , the maximum of the impedance is slightly displaced to the right of the transition frequency. A balanced CSRR-loaded line is depicted in Fig. 14, together

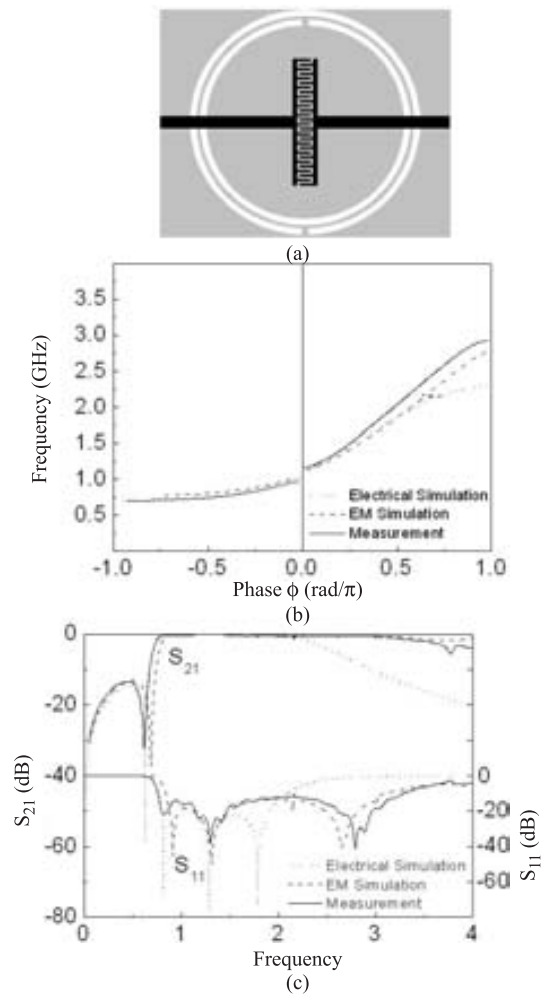


Fig. 14. Balanced CRLH cell based on a microstrip line loaded with CSRRs (a), dispersion diagram (b), and frequency response (c). The structure has been implemented in Rogers RO3010 substrate with the dielectric constant $\epsilon_r = 10.2$ and the thickness $h = 1.27$ mm. The dimensions are: the line width $W_m = 0.8$ mm, the internal radius $r = 6.3$ mm, the ring width $c = 0.4$ mm, and the ring separation $d = 0.2$ mm; the interdigital capacitor, formed by 28 fingers separated 0.16 mm, has been used to achieve the required capacitance value.

with the dispersion diagram and frequency response. The transmission zero and the broad band, consequence of the continuous transition between the backward and forward bands, are visible (these balanced structures have been successfully applied to the design of high pass filters [28]).

According to these comments, CL-loaded, SRR-loaded and CSRR-loaded metamaterial transmission lines are conceptually very similar structures that exhibit comparable behaviour, including the possibility of balancing the lines. In this later case, the characteristic impedance is scarcely dependent on frequency in the vicinity of the transition frequency, and the allowed band can be made very wide. CL-loaded lines are more easily synthesized and usually exhibit a wider left handed band. However, to extend the left handed band significantly beyond those widths achievable through the resonant type approach, it is usually required the use of lumped components. The presence of the transmission zero in resonant type metamaterial transmission lines is very interesting for the application of these structures to microwave filters, as has been recently demonstrated [28]. However, it has also been demonstrated by the authors that resonant type metamaterial transmission lines can be useful for the design of enhanced bandwidth components [29,30], dual-band components [31], and tunable components [32–34].

3. Conclusions

In conclusion, we have reviewed the main metamaterial transmission line types, we have discussed their lumped element circuit models, and we have compared these lines to the light of such circuit models. Although apparently very different, we have found that there are actually many similarities between CL-loaded, SRR-loaded and CSRR-loaded artificial lines. We have seen that CL-loaded and resonant type metamaterial transmission lines can be modelled by means of circuit models that only differ in a single element, and we have pointed out that SRR- and CSRR-loaded lines are described by formally dual circuits, this explaining the almost undistinguishable behaviour of these resonant type artificial lines, and the effects of geometry.

Acknowledgments

This work has been supported by MEC by project contract TEC2007-68013-C02-02 META-INNOVA and by a FPU Grant (0403/2005) awarded to Marta Gil. Thanks are also given to the European Union for funding the Network of Excellence NoE METAMORPHOSE, and to the Catalan Government (CIDEM) for funding CIMITEC and for giving support through the action SGR-2005-00624.

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Artículo H

Referencia:

F. Aznar, J. Bonache and F. Martín, “Analysis of resonant-type metamaterial transmission lines on the basis of their equivalent circuit models”, *Asia Pacific Microwave Conference (APMC)*, Hong Kong (China). December 2008.

Resumen del artículo:

En este artículo, se lleva a cabo un análisis exhaustivo de líneas de transmisión metamaterial de tipo resonante. En concreto, los efectos de las dimensiones y características del resonador sobre la respuesta frecuencial es estudiada e interpretada a partir del modelo circuital equivalente de las líneas. Este estudio es importante para la aplicación de estas líneas artificiales para el diseño de circuitos de microondas, ya que se pueden deducir importantes directrices relativas al ancho de banda de los dispositivos.

Analysis of resonant-type metamaterial transmission lines on the basis of their equivalent circuit models

F. Aznar, J. Bonache and F. Martín

CIMITEC, Departament d'Enginyeria Electrònica,
Universitat Autònoma de Barcelona
08193 BELLATERRA (Barcelona), Spain

Abstract

In this paper, an exhaustive analysis of resonant type metamaterial transmission lines is carried out. Specifically, the effects of the dimensions and resonator characteristics on the frequency response are studied and interpreted to the light of the equivalent circuit model of the lines. This study is important for the application of these artificial lines to microwave circuit design, since important guidelines concerning device bandwidth can be inferred.

1. Introduction

Resonant type metamaterial transmission lines based on split ring resonators (SRRs) were proposed in 2003 by Martín *et al.* [1]. Such artificial lines consist on a host coplanar waveguide (CPW) loaded with SRRs and shunt connected strips (see Fig. 1). The SRRs provide the required negative effective permeability to the structure in a narrow band above their resonance frequency, whereas the shunt strips act as shunt connected inductors, and give the negative value of the permittivity up to a frequency that depends on strips width and length. In this way, left handed wave propagation is achieved in a certain band, as has been demonstrated [1]. The circuit model (unit cell) of these artificial lines is depicted in Fig. 2(a) [2] (such circuit is an improved version of that published in [1]). L is the line inductance, L_p is the inductance of the pair of shunt strips, C is the line capacitance, L_s and C_s model the SRR, and M is the mutual inductance between the host line and the SRRs. This circuit can be transformed into that circuit depicted in Fig. 2(b), where the following correspondence between parameters applies:

$$L_s' = 2M^2 C_s \omega_o^2 \frac{\left(1 + \frac{L}{4L_p}\right)^2}{1 + \frac{M^2}{2L_p L_s}} \quad (1)$$

$$C_s' = \frac{L_s}{2M^2 \omega_o^2} \left(\frac{1 + \frac{M^2}{2L_p L_s}}{1 + \frac{L}{4L_p}} \right)^2 \quad (2)$$

$$L' = \left(2 + \frac{L}{2L_p}\right) \frac{L}{2} - L_s' \quad (3)$$

$$L_p' = 2L_p + \frac{L}{2} \quad (4)$$

with $\omega_0=(L_s C_s)^{-1/2}$. The typical frequency response of the unit cell of these structures is depicted in Fig. 3. A transmission zero to the left of the left handed pass band arises. This transmission zero is due to the presence of the SRR and is given by the following expression:

$$\omega_z = \omega_0 \left(1 + \frac{M^2}{2L_p L_s} \right)^{-1/2} \quad (5)$$

If the shunt inductance is absent, the structure behaves as a one-dimensional negative permeability medium and a stop band arises in the vicinity of the resonance frequency (a typical response is also depicted in Fig. 3).

The purpose of this paper is to analyze the effects of the parameters of the circuit in Fig. 2, with the final aim of analyzing the influence of using different resonant particles, apart from the SRR, on the frequency response. Specifically it is our aim to explain the reason of bandwidth reduction when sub-wavelength particles are used.

2. Analysis and discussion

Let us consider for simplicity the circuit of Fig. 2(a) without the presence of the shunt inductance ($L_p \rightarrow \infty$). The analysis of the circuit [Fig. 2(b)] reveals that, in order to enhance bandwidth, i.e. decrease the susceptance slope ($b=\omega_0' C_s'$) of the lumped $L_s' C_s'$ parallel resonator [3], we need a large value of M and C_s and a small value of L_s (this provides a large value of L_s' [Eqs. (1)] and a small value of C_s' [Eqs. (2)] in the circuit of Fig. 2b). In CPW configuration, M is related to the proximity of the particle to the CPW metal level and to particle dimensions (in a first order approximation, $M=Lf$, where f is the slot fraction occupied by the particles [1]). M is a key parameter since the dependence of L_s' and C_s' on it is quadratic, and bandwidth is indeed given by the ratio of L_s' and C_s' . It has been experimentally shown that by decreasing the electrical size of the particles, the fractional bandwidth of the stop bands (in negative permeability structures) also decreases. For a certain given resonance frequency, the smaller the electrical size of the particle is, the smaller M becomes. This explains bandwidth reduction, but the key question is if we can compensate this inevitable reduction in M with an enhancement in C_s and a reduction in L_s , and thus maintain or even increase the bandwidth. To this end we have considered, apart from a SRR, a broad side coupled (BC-SRR) [4] and a spiral resonator (SR) [5], and we have driven particle dimensions to the limit to enhance C_s and reduce L_s (by using very wide particle and close strips). The layouts of the particles coupled to the host CPW, with and without shunt strips, are given in Fig. 4, where transmission coefficients is also depicted. As can be seen we can not enhance bandwidth by using SRs or BC-SRR. The reason is that M is the key parameters since it has the greater influence in L_s' and C_s' . Thus, from this analysis we derive the conclusion that M is the relevant parameter determining bandwidth, and the fractional bandwidth decreases by using electrically smaller particles.

3. Conclusion

In conclusion, it has been shown that by reducing the electrical size of metamaterial resonators, the bandwidth of metamaterial transmission lines based on them (stop band in negative permeability lines and pass band in left handed lines) is generally degraded, and an explanation, based on the effects of mutual coupling between the line and the particles, has been given.

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Figures

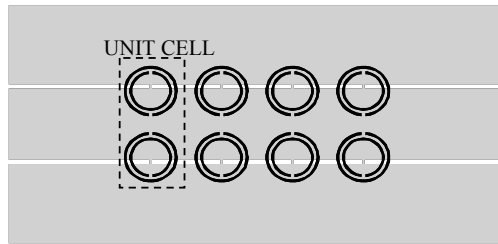


Figure 1. Typical SRR loaded CPW metamaterial transmission line [1].

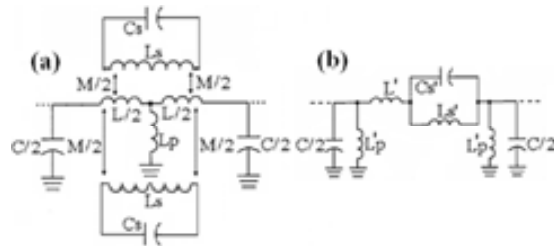


Figure 2. Equivalent circuit model of the unit cell of CPW transmission lines loaded with SRR (or other resonant elements) and shunt connected strips (a) and transformed model (b) [2].

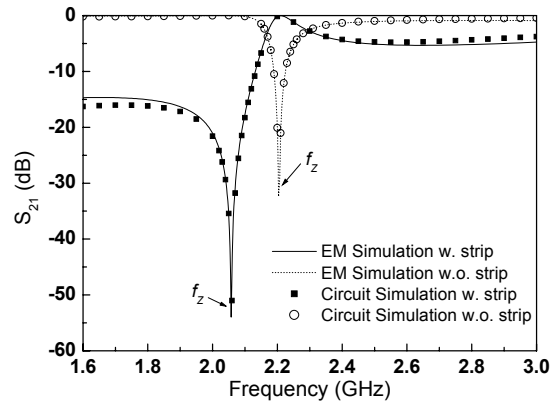


Figure 3. Frequency response of the unit cell of a CPW loaded with SRRs and shunt strips, and with SRRs only (without shunt strips).

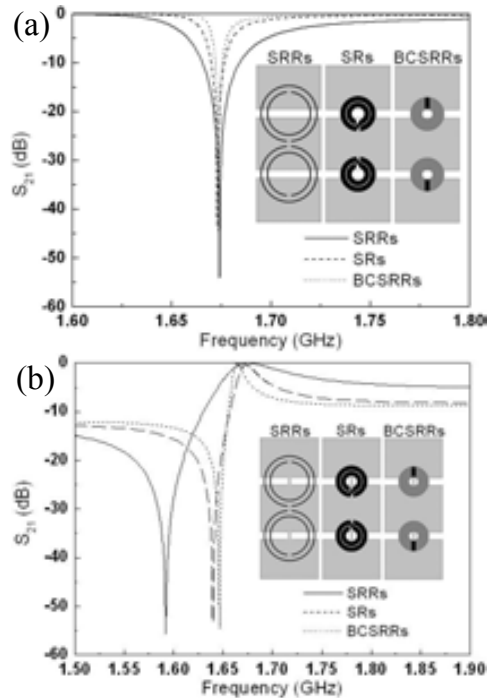


Figure 4. Layouts and transmission coefficients of several CPWs loaded with pairs of SRRs, BC-SRRs and SRs (a) and without shunt strips (b). Relevant dimensions are: ring width $c=0.2\text{mm}$, distance between the rings $d=1\text{mm}$, external radius $r_{ext}=5.4\text{mm}$ for the SRRs; $c=0.961\text{mm}$, $d=0.19\text{mm}$, $r_{ext}=3.36\text{mm}$ for the SRs; and $c=2.12\text{mm}$, $d=0.2\text{mm}$, $r_{ext}=3.17\text{mm}$ for the BCSRRs. For the CPW the central strip width is $W=10\text{mm}$, the width of the slots is $G=1.59\text{mm}$ and the length $D=12.2\text{mm}$, shunt strip width $l_v=0.2\text{mm}$. The considered substrate characteristics are: dielectric constant $\epsilon_r=10.2$ and thickness $h=2.54\text{mm}$.

Artículo I

Referencia:

F. Aznar, M. Gil, J. Bonache and F. Martín, “SRR- and CSRR-loaded metamaterial transmission lines: a comparison to the light of duality”, *International Congress on Advanced Electromagnetic Materials in Microwaves and Optics* (Metamaterials), Pamplona (Spain). September 2008. **Invited**.

Resumen del artículo:

En este artículo, se señalan y comparan las principales características de líneas de transmisión metamaterial cargadas con *split ring resonators* (SRR) y con *complementary split ring resonators* (CSRR). Para obtener una línea zurda paso banda, las líneas cargadas con SRR (normalmente implementado en tecnología CPW) se cargan también con tiras metálicas en derivación (que actúan como inductancias en derivación). En líneas zurdas basadas en CSRRs, el elemento reactivo adicional es un corte en la línea microstrip (capacidad en serie). Se demuestra que la frecuencia del típico cero de transmisión de estas estructuras varía de una forma similar (disminuye) cuando el corte serie o la tira en derivación se amplían (para las líneas basadas en CSRR y SRR, respectivamente). Sin embargo, se ha constatado que la frecuencia de fase igual a cero, que es la frecuencia donde la fase de la celda unidad se anula, no depende de la geometría del corte serie o de la tira en derivación. Estos comportamientos similares se interpretan a partir de los modelos circuitales mejorados de las líneas de transmisión metamaterial (recientemente presentados por los autores) y una discusión de la dualidad. Aunque las líneas de transmisión metamaterial comparadas no son estrictamente estructuras electromagnéticamente duales, sus modelos circuitales son formalmente duales circuitales.

SRR- and CSRR-loaded metamaterial transmission lines: a comparison to the light of duality

Francisco Aznar, Marta Gil, Jordi Bonache and Ferran Martín

CIMITEC, Departament d'Enginyeria Electrònica, Universitat Autònoma de Barcelona
 08193 BELLATERRA (Barcelona), Spain
 Fax: 34 93 581 26 00; E-mail: Ferran.Martin@uab.es

Abstract

In this paper, the main relevant characteristics of split ring resonator (SRR) and complementary split ring resonator (CSRR) loaded metamaterial transmission lines are pointed out and compared. To obtain a left handed pass band, the SRR-loaded lines (typically implemented in CPW technology) are also loaded with shunt connected strips (acting as shunt inductances). In CSRR-based left handed lines, the additional reactive element is a series gap (series capacitor). It is shown that the typical transmission zero frequency of these structures varies in a similar form (decreases) when the series gap and the shunt strip are widened (for the CSRR- and SRR-based lines, respectively). However, it has been found that the zero-phase shift frequency, that is, the frequency where the phase shift of the unit cell is null, does not depend on the geometry of the series gap or shunt strip. This similar behaviour is interpreted to the light of the improved circuit models of the considered metamaterial transmission lines (recently reported by the authors), and duality arguments. Although the compared metamaterial transmission lines are not exactly electromagnetic dual structures, their models are formally circuit duals.

1. Introduction

Metamaterial transmission lines can be implemented through the so-called CL-loaded approach (see [1]) or by means of the resonant type approach. In the latter case, a host line is either loaded with split ring resonators (SRRs) and shunt connected inductors (strips or vias) [2] or with complementary split ring resonators (CSRRs) and series capacitances (series gaps or interdigital capacitors) [3]. The typical layout of these structures is depicted in Fig. 1 (usually the SRR-based lines are implemented in CPW technology and the CSRR-loaded lines in microstrip technology). These structures exhibit a left handed pass band in the vicinity of the resonance frequency of the SRRs or CSRRs, as has been exhaustively discussed in the recent literature [2,3], and a forward wave transmission band at higher frequencies [4] (for this reason these structures are called composite right/left handed –CRLH– transmission lines). They also exhibit a transmission zero to the left of the left handed band, which has been applied for instance to the design of miniature high pass filters with sharp transition band.

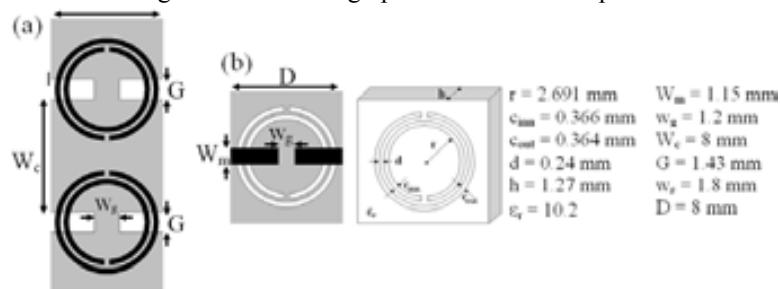


Fig. 1: Typical layout (unit cell) of a SRR-loaded (a) and a CSRR-loaded (b) metamaterial transmission line implemented in CPW and microstrip technology, respectively. In (a), the pair of SRRs (depicted in black) are etched in the back side of the substrate; in (b), the CSRR is etched in the ground plane.

This paper is focused on investigating the similarities of both SRR-loaded and CSRR-loaded metamaterial transmission lines. As will be shown, they exhibit a very similar behaviour. This behaviour can

be explained from the analysis of the recently reported circuit models of the lines. As will be shown, these models are circuit duals, and this provides a qualitatively interpretation of the results. Special emphasis is put on this aspect.

2. Circuit models of the SRR- and CSRR-based metamaterial transmission lines

The lumped element T-circuit model for the CSRR-loaded CRLH metamaterial transmission line (unit cell) is depicted in Fig. 2(a) [5]. In this model, L is the line inductance, C_s is the gap capacitance, C_f is the fringing capacitance of the gap, C_L is the line capacitance, and the CSRR is modelled by the resonant tank L_c - C_c . This model can be transformed into that depicted in Fig. 2(b) which was the first model of the structure reported by the authors [3]. The following equations apply in this transformation:

$$C_g = 2C_s + C_{par} \qquad C = \frac{C_{par}(2C_s + C_{par})}{C_s} \qquad (1)$$

The model in Fig 2(b) is formally correct, but the interpretation given to some of the parameters in the recent papers was not [6].

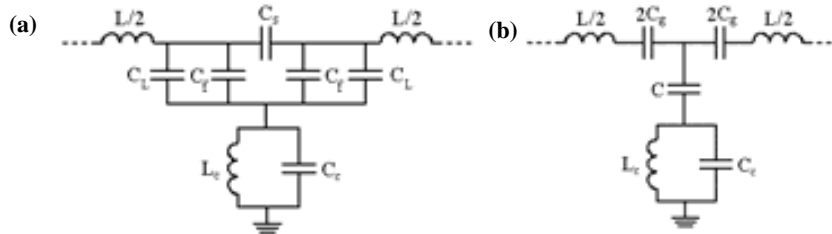


Fig. 2: Lumped element equivalent T-circuit model of the CSRR-loaded line (a) and transformed model (b)

With regard to the model of the SRR-loaded metamaterial transmission lines, it has been recently reported by the authors [6], as an improved circuit model of that formerly proposed in 2003 [2]. It is depicted in Fig. 3(a). L and C are the line inductance and capacitance, respectively, L_p is the inductance of the shunt strips, the SRRs are modelled by the resonant tanks L_s - C_s and M is the mutual inductance, accounting for magnetic coupling between the line and the SRRs. This circuit can be transformed into that depicted in Fig. 3(b), which is formally identical to the transformed circuit reported in [2]. The following equations apply in this transformation [6]:

$$L_s' = 2M^2 C_s \omega_o^2 \frac{\left(1 + \frac{L}{4L_p}\right)^2}{1 + \frac{M^2}{2L_p L_s}}; \quad C_s' = \frac{L_s}{2M^2 \omega_o^2} \left(\frac{1 + \frac{M^2}{2L_p L_s}}{1 + \frac{L}{4L_p}} \right)^2; \quad L' = \left(2 + \frac{L}{2L_p}\right) \frac{L}{2} - L_s'; \quad L_p' = 2L_p + \frac{L}{2} \quad (2)$$

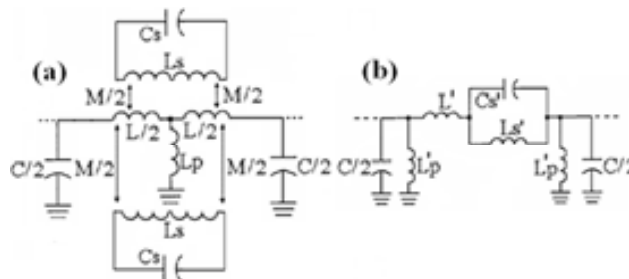


Fig. 3: Lumped element equivalent π -circuit model of the CSRR-loaded line (a) and transformed model (b).

3. Comparison of SRR- and CSRR-loaded metamaterial transmission lines

The simulated (by means of *Agilent Momentum*) frequency responses of the structures of Fig. 1, for different values of the strip width (SRR-loaded line) and gap separation (CSRR-loaded line), are depicted in Fig. 4. As the strip width and gap separation increase, the transmission zero decreases. In CSRR-loaded lines this is due to the fact that C_s decreases and hence C increases (and it lowers the transmission zero). In SRR-loaded lines, the resonance frequency of the tank formed by L_s ' and C_s ' also decreases as the strip width increases (this decreases L_p and hence the resonance frequency, as is derived from 2). It is also interesting to mention that the zero phase shift frequency does not depend on gap separation (for CSRR-loaded lines) or strip width (for SRR-loaded lines). In SRR-loaded lines this frequency nulls the impedance of the series branch of the circuit model (Fig. 3b); it is given by:

$$\omega_s = \frac{1}{\sqrt{C_s \left(L_s - 2 \frac{M^2}{L} \right)}} \quad (3)$$

and it is inferred analytically that the zero phase shift frequency is independent of L_p . In CSRR-loaded lines, ω_s is given by the resonance frequency of the tank L_c - C_c , since it nulls the shunt admittance of the T-circuit model.

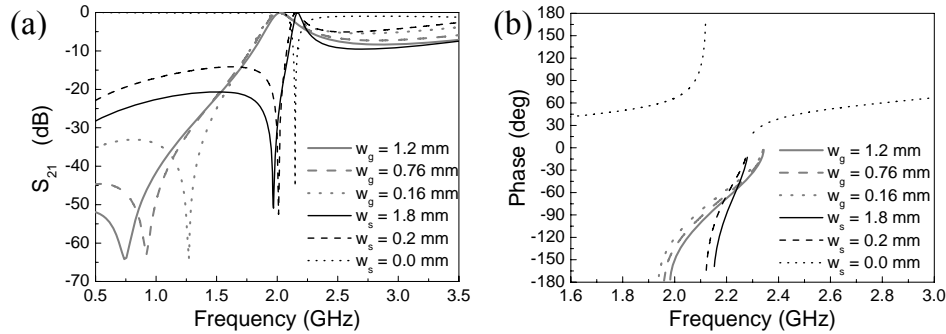


Fig. 4: Simulated frequency responses (a) and the dispersion relation (b) for the structures in Fig. 1 for different gap separations (w_g) and strip widths (w_s). Notice that $w_s=0.0$ mm corresponds to the particular case when $L_p=\infty$.

4. Interpretation of the results and conclusion

Both structures exhibit similar behaviour. Indeed, this is not surprising since these structures are roughly dual. They are not strictly dual (electromagnetically), but their lumped element models are formally circuit duals. According to this, the effects of the series gap (CSRR-based lines) and shunt strips (SRR-loaded lines) on the frequency response must be qualitatively very similar, as it occurs.

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Artículo J

Referencia:

F. Aznar, M. Gil, J. Bonache, J. García-García and F. Martín, “Metamaterial transmission lines based on broad-side coupled spiral resonators”, *Electronics Letters*, Vol. 43, No. 9, pp. 530-532. April 2007.

Resumen del artículo:

Se han diseñado, fabricado y caracterizado por primera vez líneas microstrip periódicas cargadas con *broad-side coupled spiral resonators* (BC-SRs). Estas estructuras inhiben la propagación de la señal en las proximidades de la frecuencia de resonancia de las espirales, f_0 . Este comportamiento de rechazo de banda es debido al acoplamiento magnético entre la línea y los BC-SRs en la resonancia. Como el tamaño eléctrico de los BC-SRs es pequeño, las estructuras pueden ser consideradas como líneas de transmisión metamaterial con permeabilidad efectiva negativa, μ_{eff} . Este valor negativo de μ_{eff} tiene lugar en una estrecha banda por encima de f_0 , como ocurre en líneas cargadas con *split ring resonators* (SRR), o en líneas de transmisión acopladas a espirales grabadas en una única capa de metal (*edge-coupled spiral resonators*, EC-SRs). Sin embargo, debido a que cada anillo del BC-SR está grabado en diferentes capas de metal (anillos cara a cara), la capacidad equivalente del tanque resonante puede ser aumentada sustancialmente. Así, mediante el uso de estrechos subs-tratos de microondas, las líneas cargadas con BC-SRs pueden hacerse mucho más pequeñas que las líneas cargadas con SRR o EC-SR.

Metamaterial transmission lines based on broad-side coupled spiral resonators

F. Aznar, M. Gil, J. Bonache, J. García-García and F. Martín

Microstrip lines periodically loaded with broad-side coupled spiral resonators (BC-SRs) are designed, fabricated and characterised for the first time. These structures inhibit signal propagation in the vicinity of the resonance frequency of the spirals, f_o . This stopband behaviour is due to magnetic coupling between the line and the BC-SRs at resonance. As long as the electrical size of the BC-SRs is small, the structures can also be considered as metamaterial transmission lines with negative effective permeability, μ_{eff} . This negative value of μ_{eff} takes place in a narrow band above f_o , as occurs in split ring resonator (SRR) loaded lines, or in transmission lines coupled to spiral resonators etched in a single metal level (edge-coupled spiral resonators (EC-SRs)). However, because each loop of the BC-SR is etched in a different metal level (face-to-face loops), the equivalent capacitance of the resonant tank can be substantially enhanced. Thus, by using narrow microwave substrates, the BC-SRs loaded lines can be made much smaller than SRR or EC-SR loaded lines.

Introduction: It has been previously demonstrated that split ring resonators (SRRs) [1] or spiral resonators (SRs) [2] coupled to microstrip lines or coplanar waveguides (CPWs) are able to inhibit signal propagation in the vicinity of resonance, f_o [3–5]. This stopband behaviour is due to the inductive coupling between the line and the particles. Namely, the magnetic field of the line induces current loops in the particles at resonance, and the incident signal is returned back to the source. Alternatively, this frequency gap has been interpreted by considering the structures as one-dimensional effective media [6]. In other words, since the electrical size of the particles is small (diameter much smaller than signal wavelength at resonance, λ_g), these structures are artificial transmission lines with effective parameters, permeability (μ_{eff}) and permittivity (ϵ_{eff}). For transmission lines loaded with SRRs or SRs, μ_{eff} is negative/positive in a narrow band above/below resonance, and it takes extreme values near f_o . For this reason, the typical stopband of these structures extends not only above f_o (where $\mu_{eff} < 0$), but also in a narrow band below it [7].

It was previously demonstrated from theory that the resonance frequency of a two-turn SR is approximately half the resonance frequency of a SRR of identical dimensions [2], and it was experimentally corroborated later [5]. This represents a substantial miniaturisation of the resonator size for the SR compared to the SRR. It was also demonstrated that by etching either ring of the SRR face-to-face at both sides of a narrow dielectric slab, the distributed capacitance between the rings can be significantly enhanced owing to broad-side coupling [8]. The resulting structure was labelled as broad-side coupled SRRs (BC-SRRs) to distinguish it from the conventional SRR (also called edge-coupled SRR-EC-SRR). Dimensions (diameter) as small as $\lambda_g/20$ or smaller were demonstrated for BC-SRRs (in a 100 μm -thick substrate with $\epsilon_r = 10$) [8]. From these considerations, it follows that by etching the two turns of a spiral resonator at both sides of a thin dielectric, face-to-face and connected through a metallic via (see Fig. 1), high levels of miniaturisation for these particles are expected. Namely, the benefits of the spiral topology and the broad-side coupling are combined. For obvious reasons, and following the nomenclature previously used, this particle can be called a broad-side coupled spiral resonator (BC-SR). It will be shown in the following Sections that the size of negative permeability lines loaded with BC-SRs can be made actually very small compared to previous implementations based on SRRs or SRs.

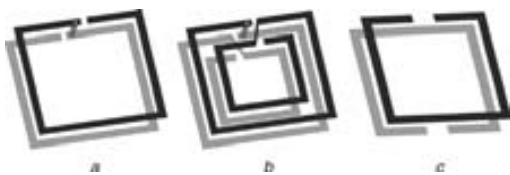


Fig. 1 Structure of BC-SR with two turns (each etched on opposite substrate sides) connected through a metallic via (Fig. 1a), and topology of four-turn BC-SR (Fig. 1b) (in latter structure, two turns are etched on both top and bottom metal levels); for comparison, structure of BC-SRR depicted in (Fig. 1c)

Design of microstrip lines loaded with BC-SRs: The design of the microstrip lines loaded with BC-SRs follows the procedure described in previous papers, where such lines were coupled to conventional SRRs [3, 5]. However, since BC-SRs need two metal levels, it is imperative to open windows in the ground plane in order to etch the lower ring of the spiral. Fig. 2 shows the layout of the designed device. It consists of a 50 Ω microstrip line with ground plane windows and three pairs of square shaped BC-SRs. To preserve the characteristic impedance of the line in the region where the spirals are present, the strip width has been modified. The considered substrate is the Rogers RO3010 with thickness $h = 254 \mu\text{m}$ and dielectric constant $\epsilon_r = 10.2$. The dimensions of the BC-SRs are given in the caption of Fig. 2. The period of the structure is 7.82 mm, the separation between the signal strip and the BC-SRs is 0.2 mm, and the distance between adjacent BC-SRs is 0.9 mm.

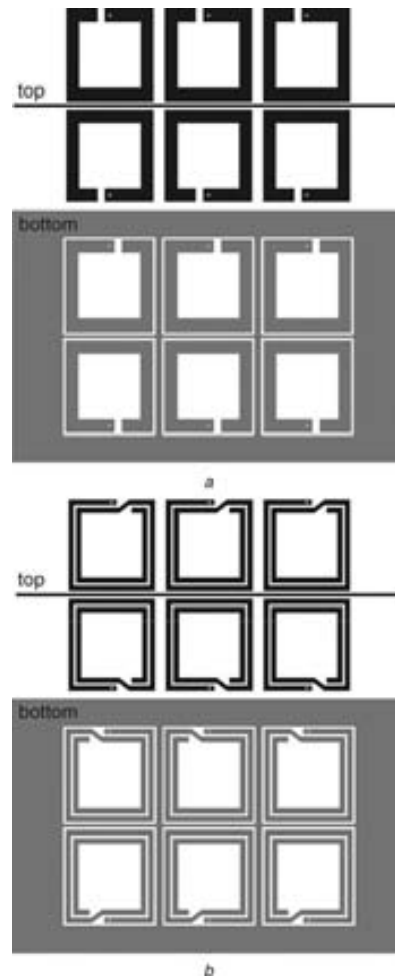


Fig. 2 Layouts of two-turn (Fig. 2a) and four-turn (Fig. 2b) BC-SR loaded microstrip lines

In Fig. 2a, strip width of resonators is 1 mm, area $6.92 \times 6.92 \text{ mm}$ and metallic vias consist of metalised holes with diameter 0.2 mm. In Fig. 2b, resonators occupy identical area, strip width is 0.4 mm in whole structure, and separation between different strips is 0.2 mm. Except that spiral is replaced by a SRR, the layout of the BC-SRR loaded line (which has been also fabricated) is identical to that of Fig. 2a and hence is not depicted

Results: The simulated (through the electromagnetic software Agilent Momentum 2005A) frequency response of the structure shown in Fig. 2 is depicted in Fig. 3. For comparison, Fig. 3 also includes the frequency responses that have been obtained from electromagnetic simulation of the same structure, but replacing the BC-SRs by BC-SRRs and by four-turn (two turns on both substrate sides) BC-SRRs. The BC-SRRs have identical dimensions to those of the BC-SRs of Fig. 2. The four-turn BC-SRs (see Fig. 1) have narrower strips (0.4 mm), but they occupy the same area.

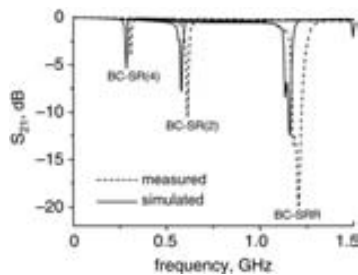


Fig. 3 Measured and simulated insertion losses for three loaded microstrip lines designed and fabricated (number in parentheses refers to number of turns of BC-SRs)

As can be appreciated, the resonance frequency of the structures (revealed by the notch in the transmission coefficient) is lower for the four-turn BC-SR, and higher for the BC-SRR. The three structures have been fabricated by means of a milling machine (LPKF HF100) and have been characterised through the Agilent 8720ES vector network analyser (the measured transmission coefficients are also depicted in Fig. 3). Good agreement between simulations and experiments has been obtained. In terms of guided wavelength at resonance, the size (side length) of the different resonators (extracted from experimental data) is $\lambda_g/14$, $\lambda_g/28$ and $\lambda_g/56$ for the BC-SRR, BC-SR and four-turn BC-SR, respectively. Thus, it is possible to achieve very small electrical size for the resonant elements by providing broad-side coupling to spiral topologies. Obviously, these dimensions can be further optimised using thinner (5mils or below) substrates. In that case, rather than mechanical systems (milling machine), high resolution photo/mask etching techniques are more convenient owing to the lack of rigidity of such narrow substrates.

Conclusions: It has been demonstrated that the electrical size of metamaterial resonant elements can be substantially reduced by combining spiral topologies with broad-side coupling. Specifically, it has been shown that the resonance frequency of two- and four-turn broad-side coupled spiral resonators is decreased by a factor of 2 and 4, respectively, compared to that of a broad-side coupled split rings resonator. For the optimum structure (four-turn spiral) the achieved area is as small as $3.2 \times 10^{-4} \lambda_g^2$ (λ_g being signal wavelength at resonance). With such small dimensions, these resonant elements can be of interest for the synthesis of effective media with negative permeability. These particles are also of interest for the suppression of narrow band interfering signals in applications where size is critical.

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F. Aznar, M. Gil, J. Bonache, J. García-García and F. Martín (CIMITEC, Departament d'Enginyeria Electrònica, Universitat Autònoma de Barcelona, Barcelona, Bellaterra 08193, Spain)

E-mail: joan.garcia@uab.es

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Artículo K

Referencia:

F. Aznar, J. Bonache, A. Valcarcel and F. Martín, “Miniaturization of narrow-band power dividers by using CPW metamaterial transmission lines”, *Microwave and Optical Technology Letters*, Vol. 51, No. 4, pp. 926-929. April 2009.

Resumen del artículo:

En este artículo, se demuestra que el tamaño de los circuitos y componentes de microondas planares puede ser reducido usando líneas de transmisión metamaterial en tecnología coplanar (CPW). El pequeño tamaño y la posibilidad de controlar las características eléctricas (fase e impedancia característica) de estas líneas, las hace apropiadas para el diseño de dispositivos planares compactos. Para ilustrar las posibilidades del enfoque, se han diseñado, fabricado (en un sustrato de microondas de bajas pérdidas convencional) y medido varios divisores de potencia, que consisten en CPWs cargadas con *split ring resonators* (SRR) y *spiral resonators* (SRs). Se ha logrado un sesenta por ciento de reducción del tamaño, respecto a dispositivos convencionales. Esta reducción de tamaño se puede mejorar mediante el uso de menores resonadores basados en diferentes topologías.

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MINIATURIZATION OF NARROW-BAND POWER DIVIDERS BY USING CPW METAMATERIAL TRANSMISSION LINES

Francisco Aznar, Jordi Bonache, Alexandre Valcarcel, and Ferran Martín

GEMMA/CIMITEC, Departament d'Enginyeria Electrònica, Universitat Autònoma de Barcelona, 08193 BELLATERRA (Barcelona), Spain; Corresponding author: Ferran.Martin@uab.es

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ABSTRACT: In this article, it is demonstrated that the size of planar microwave circuits and components can be reduced by using coplanar waveguide (CPW) metamaterial transmission lines. The small size and the possibility to control the electrical characteristics (phase and characteristic impedance) of these lines, make them suitable for compact planar device design. To illustrate the possibilities of the approach, different power dividers, consisting on split-ring resonators (SRRs) and spiral resonators (SRs) loaded CPWs, have been designed, fabricated (in a conventional low loss microwave substrate) and measured. Sixty percentage size reduction, as compared to conventional devices, has been achieved. This size reduction can be enhanced by using smaller resonators based on different topologies. © 2009 Wiley Periodicals, Inc. *Microwave Opt Technol Lett* 51: 926–929, 2009; Published online in Wiley InterScience (www.interscience.wiley.com). DOI 10.1002/mop.24195

Key words: metamaterials; split-ring resonator (SRRs); transmission lines; power dividers

1. INTRODUCTION

Many techniques devoted to size reduction in planar microwave circuits and components have been proposed in the literature. This is due to the fact that planar microwave components usually exhibit large dimensions, and this can be a critical aspect in certain applications. One of these approaches consists on the implementation of microwave components by means of metamaterial transmission lines. Such lines are artificial lines consisting on a host line loaded with reactive elements, and they can be implemented by means of two main approaches: (i) the CL-loaded approach, where conventional transmission lines are loaded with series capacitances and shunt inductances [1–3] and (ii) the resonant-type approach, where the lines are loaded with sub-wavelength resonators, such as split-ring resonators (SRRs) or complementary split-ring resonators (CSRRs) combined with shunt inductances and series capac-

itances, respectively [4, 5]. These lines exhibit controllable electrical characteristics (phase and characteristic impedance), beyond what can be achieved in conventional lines, because of the fact that there are more degrees of freedom. In addition, metamaterial transmission lines can be designed to be electrically small, what makes them suitable for the synthesis of compact microwave circuits [6]. In the case of the resonant-type approach, the size of the line is determined by the size of the resonators and, for this reason, it is possible to obtain an important level of miniaturization (the use of electrically small resonators and the possibility to implement artificial lines with the required phase and impedance with a single unit cell is relevant for size reduction [7–9]).

In previous works by some of the authors, it was demonstrated that left-handed microstrip lines based on CSRRs were useful for the design of microwave components [10, 11]. In this work, it is demonstrated that left-handed lines implemented in CPW technology and based on SRRs and other related topologies, such as the spiral resonator (SR), are also useful for the synthesis of compact microwave devices. Specifically, several power dividers have been designed, fabricated, and measured. Essentially, such power dividers consist on a 35.35Ω impedance inverter and two 50Ω output lines. It will be shown that the required impedance (35.35Ω) and phase (90°) of the inverter can be achieved with a single unit cell based on such resonators.

2. TOPOLOGY AND CIRCUIT MODEL OF LEFT-HANDED LINES BASED ON SRRs

As shown in Figure 1, a CPW can be loaded with magnetically coupled SRRs (etched in the back substrate side) and metallic strips placed between the central strip and the ground plane (acting thus as shunt connected inductances). A periodic structure consisting on a cascade of such unit cells can be considered, in a certain band above SRR's resonance, as a one-dimensional effective medium exhibiting left handed wave propagation (that is, with anti-parallel phase and group velocities), and this has been interpreted as due to the negative effective permeability and permittivity of the line [4]. In this region, the phase constant is negative, as opposite to the positive phase constant of conventional lines. By using a single unit cell line, we can no longer consider the structure as an effective medium and, hence, we can not properly talk in terms of

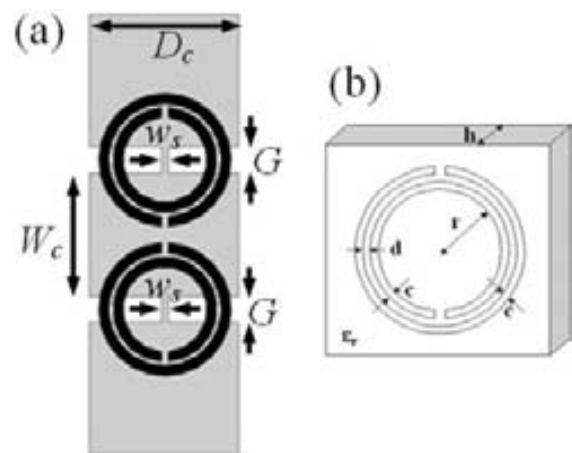


Figure 1 Layout of a left-handed cell in CPW technology (a), where the CPW (grey) is loaded with shunt inductances and magnetically coupled resonators on the other side of the substrate (black). Relevant dimensions for the SRR (b)

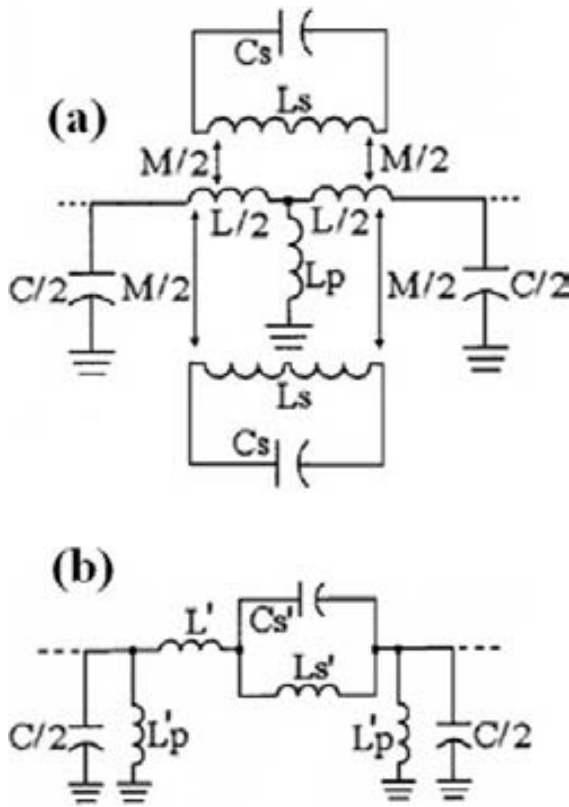


Figure 2 Circuit model for a left-handed cell based on SRRs (a) and its transformed model (b)

effective permeability or permittivity. However, the phase constant of these unit cell lines is also negative and therefore they exhibit left handed wave propagation. As long as the phase constant (or the dispersion diagram) and the characteristic impedance are controllable, such lines can be called metamaterial transmission lines, regardless of the number of cells. In these artificial transmission lines, homogeneity is not relevant, but the control of the characteristic impedance and phase. Nevertheless, we would like to mention that what is important for size reduction of the power dividers presented in this work is the small electrical size achievable with the unit cells of these lines, rather than the sign of the phase constant.

The lumped element equivalent circuit model for the unit cell of SRR-loaded left handed CPWs is shown in Figure 2(a). This circuit can be transformed into the π -circuit represented in Figure 2(b), subjected to the following expressions [12]:

$$L_s' = 2M^2 C_s \omega_0^2 \frac{\left(1 + \frac{L}{4L_p}\right)^2}{1 + \frac{M^2}{2L_p L_s}} \quad (1)$$

$$C_s' = \frac{L_s}{2M^2 \omega_0^2} \left(\frac{1 + \frac{M^2}{2L_p L_s}}{1 + \frac{L}{4L_p}} \right)^2 \quad (2)$$

$$L' = \left(2 + \frac{L}{2L_p}\right) \frac{L}{2} - L_s' \quad (3)$$

$$L_p' = 2L_p + \frac{L}{2} \quad (4)$$

In the previous expressions, L and C account for the line inductance and capacitance, respectively, C_s and L_s model the SRR, M is the mutual inductive coupling between the line and the SRRs, and L_p is the inductance of the shunt strips. The phase (βl) and the characteristic impedance (Z_B) of the unit cell are determined by the following expressions:

$$\cos(\beta l) = 1 + \frac{Z_s(\omega)}{Z_p(\omega)} \quad (5)$$

$$Z_B = \sqrt{\frac{Z_p^2 Z_s}{2Z_p + Z_s}} \quad (6)$$

where Z_s and Z_p are the series and shunt impedance of the unit cell model.

The possibility of controlling the electrical characteristics (βl and Z_B) is the key point to use this kind of cells for the design of microwave devices. For a specific operation frequency, the cell can be designed to provide a certain value of phase and impedance [10]. As previously indicated, in the design of the power dividers of this work, impedance inverters exhibiting a characteristic impedance of $Z_B = 35.35 \Omega$ and a phase of $\beta l = \pm 90^\circ$ are necessary. The simulated (through Agilent Momentum) frequency response for a cell designed to exhibit these values at $f_c = 1.5$ GHz is shown in Figure 3(a) (the impedance of the ports has been set to 35.35Ω in this simulation). As can be observed in Figure 3(b), the required values of phase and impedance are obtained at the design frequency (the phase and impedance have been inferred from the simulated S-parameters following standard procedures). The structure exhibits a relatively narrow band as compared with conventional power dividers. Thus, the proposed small size circuits are useful for narrow band applications. However, for certain devices, specifically those based on phase differences, resonant-type metamaterial transmission lines provide wide operative bandwidths, as has been recently demonstrated [13–15].

3. MINIATURIZATION OF POWER DIVIDERS BY MEANS OF SRR- AND SR-BASED LEFT-HANDED LINES

Y-junction power dividers based on CSRR-loaded microstrip lines have been already demonstrated [10]. This is the first time that resonant-type metamaterial transmission lines based on SRRs (or other related resonators) are used for the design of such compact power dividers in CPW technology. The topology of the power dividers is shown in Figure 4. Thus, the target is to design quarter wavelength artificial left handed lines (i.e. with $\beta l = -90^\circ$) exhibiting a characteristic impedance of $Z_B = 35.35 \Omega$ at the operating frequency ($f_c = 1.5$ GHz). The first power divider has been implemented by means of the impedance inverter shown in Figure 1. The considered substrate is the commercial Rogers RO3010 with dielectric constant $\epsilon_r = 10.2$ and thickness $h = 1.27$ mm. Relevant dimensions for the cell are (see Fig. 1): ring width $c = 0.29$ mm, distance between the rings $d = 0.19$ mm, internal radius $r = 4.05$ mm, the central strip width is $W_c = 8.85$ mm, the width of the slots is $G = 1.71$ mm and the length of the line $D_c = 13.38$ mm; the shunt strip width is $w_s = 0.60$ mm. The layout of this device can be seen in Figure 5(b), where it can be compared with a conventional power divider [Fig. 5(a)]. It

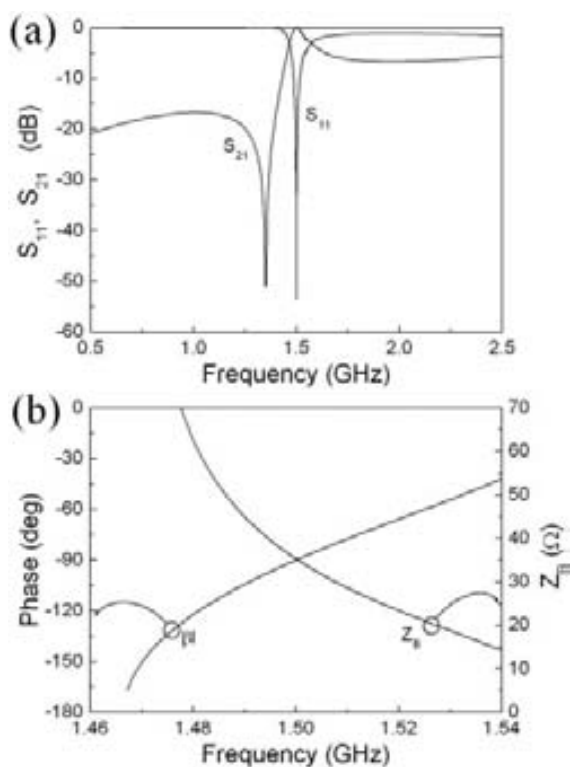


Figure 3 Frequency response (without losses) of a metamaterial cell based on SRRs designed to exhibit $Z_B = 35.35 \Omega$ y $\beta l = -90^\circ$ at $f_c = 1.5$ GHz (a). The simulation has been made using ports with impedance $Z_0 = 35.35 \Omega$. It can be seen that matching is produced at $f_c = 1.5$ GHz. Phase and characteristic impedance of the cell are also shown (b)

can be observed that, in the impedance inverter, which determines the final size of the power divider, the size reduction that can be achieved using SRRs is around 40% as compared with the conventional device. The lengths of the impedance inverters are 22.54 mm for the conventional one and 13.38 mm in the case of the SRR-based cell. This prototype has been fabricated and measured. Its transmission and reflexion coefficients are shown in Figure 6. As can be seen in the measured transmission coefficient, the insertion losses (optimum value) are -4.4 dB, which is very close to the ideal value, -3 dB. However, there is a shift in the design frequency towards lower values. This is due to the fact that the dielectric constant of the substrate has been found (from measurement) to be more than 15% higher than the nominal value (which was used for the design).

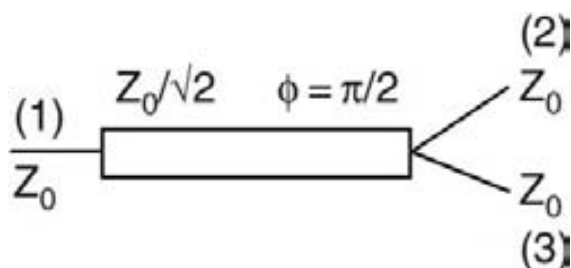


Figure 4 Model for a power divider with input impedance inverter

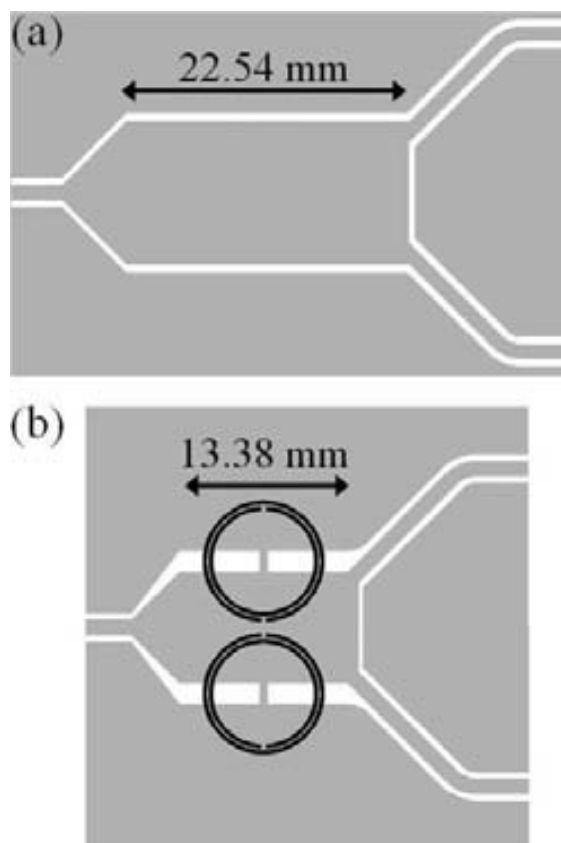


Figure 5 Conventional power divider layout (a) and designed prototype layout using a metamaterial cell based on SRRs as impedance inverter (b)

Considering the previous results, the viability of SRR-based CPW left handed lines for the miniaturization and design of microwave devices has been demonstrated. As has been previously mentioned, size reduction can be enhanced by using other smaller resonators, like the spiral resonator (SR) [8], the broadside split-

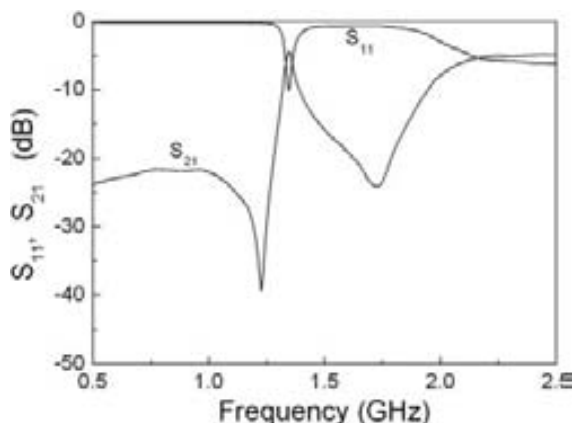


Figure 6 Measurement of the power divider prototype shown in Figure 5(b). S_{31} , very similar to S_{21} , is not represented

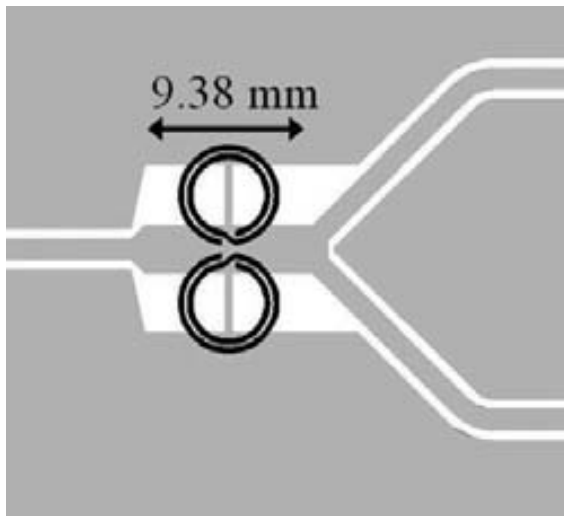


Figure 7 Layout for the power divider prototype using a metamaterial cell based on SRs as impedance inverter

ring resonator (BC-SRR) [7], and other resonators recently proposed by some of the authors in [9]. To corroborate this size reduction, a second impedance inverter with the same characteristics ($\beta l = -90^\circ$ and $Z_B = 35.35 \Omega$ at $f_c = 1.5$ GHz), but using a pair of SRs has been designed. In this case, the relevant dimensions are: ring width $c = 0.30$ mm, distance between the rings $d = 0.2$ mm, internal radius $r = 2.02$ mm, the central strip width of the line is $W_c = 2.72$ mm, the width of the slots is $G = 3.34$ mm and the length of the line $D_c = 9.38$ mm; the shunt strip width is $w_s = 0.40$ mm. The layout of the power divider based on this cell, which has been fabricated and measured, is shown in Figure 7. Its frequency response can be seen in Figure 8 (again a shift is produced due to the reasons explained before). In this device, the size reduction achieved is around 60% as compared with the conventional device.

Size reduction on the reported devices has been done at the expense of bandwidth reduction. Thus, this approach is limited to

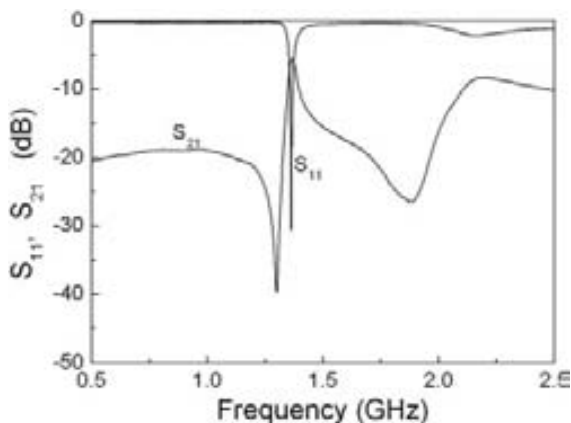


Figure 8 Measurement of the power divider prototype shown in Figure 7. S_{31} , very similar to S_{21} , is not represented.

narrow band devices. It is especially interesting in applications where power division and filtering are necessary.

4. CONCLUSIONS

In conclusion, it has been demonstrated that the artificial metamaterial transmission lines, implemented in CPW technology, where the transmission line is loaded with magnetically coupled resonators and shunt inductances, can be used in the design of compact microwave devices. As examples, several impedance inverters have been designed to be part of compact narrow band power dividers. Two different prototypes have been designed, fabricated, and characterized. A 60% size reduction level has been obtained with respect to a conventional device using metamaterial transmission lines based on SRs.

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Artículo L

Referencia:

F. Aznar, A. Vélez, J. Bonache, J. Menés and F. Martín, “Compact low pass filters with very sharp transition bands based on open complementary split ring resonators”, *Electronics Letters*, Vol. 45, No. 6, pp. 316-317. March 2009.

Resumen del artículo:

Se demuestra que los *open complementary split ring resonators* (OCSRrs) son partículas útiles para el diseño de filtros paso baja en tecnología microstrip con una banda de transición muy estrecha. Debido al pequeño tamaño eléctrico de los OCSRrs, los filtros también son muy pequeños. Estos filtros son implementados mediante varias etapas en cascada basadas en OCSRrs, sobre una línea microstrip. Las etapas de OCSRrs pueden ser modeladas como resonadores paralelo en una configuración en serie, proporcionando así un cero de transmisión en su frecuencia de resonancia. Mediante el diseño de OCSRrs con una pequeña inductancia y una alta capacidad, se consiguen cortes muy abruptos. Para ilustrar las posibilidades del modelo, se presenta como ejemplo el prototipo de un dispositivo. Consiste en un filtro paso baja periódico de cinco etapas con $3dB$ de pérdidas en inserción a $1.33GHz$ y $40dB$ de rechazo a $1.45GHz$. Para mejorar la banda de rechazo, se añade en cascada otras cuatro etapas de OCSRrs adicionales sintonizadas en las bandas de frecuencias espurias. Las dimensiones del filtro son $39mm \times 7.4mm$ (es decir, $0.46\lambda \times 0.087\lambda$, donde λ es la longitud de onda guiada a la frecuencia de corte). Estos filtros son de interés en aplicaciones donde se requieren pequeñas dimensiones y una abrupta transición.

Compact lowpass filters with very sharp transition bands based on open complementary split ring resonators

F. Aznar, A. Vélez, J. Bonache, J. Menés and F. Martín

It is shown that open complementary split ring resonators (OCSRRs) are useful particles for the design of lowpass filters with very narrow transition bands in microstrip technology. Owing to the small electrical size of OCSRRs, the filters are also very small. These filters are implemented by cascading several OCSRR stages in a microstrip line. The OCSRR stages can be modelled as parallel resonators in series configuration, thus providing a transmission zero at their resonance frequency. By designing the OCSRRs with small inductance and high capacitance, very sharp cutoffs are achieved. To illustrate the possibilities of the approach, a prototype device example is provided. It consists on a five-stage periodic microstrip lowpass filter with 3 dB insertion losses at 1.33 GHz and 40 dB rejection at 1.45 GHz. To improve out-of-band rejection, four additional OCSRR stages tuned at the spurious frequency band have been cascaded. Filter dimensions are 39×7.4 mm (i.e. $0.46\lambda \times 0.087\lambda$, where λ is the guided wavelength at the cutoff frequency). These filters are of interest in applications where small size and severe cutoff requirements are required.

Introduction: Open complementary split rings resonators (OCSRRs) have been reported recently (e.g. [1]) as useful particles for the synthesis of planar metamaterials and for the miniaturisation of microwave components in coplanar waveguide (CPW) technology. The OCSRR is the complementary counterpart of the open split ring resonator (OSRR) [2]. OCSRRs and OSRRs are open resonators, i.e. both particles can be excited by means of a voltage or current source connected to their terminals. The OCSRR (see Fig. 1) can be modelled as an open parallel resonant tank, where the inductance L_o is the inductance of the strip connecting the internal and external metallic regions of the particle, and the capacitance C_c is equivalent to the capacitance of a disk of radius $r_o - c/2$ surrounded by a metallic plane at a distance c of its edge [1]. Such element values can be roughly inferred from the geometry of the particle (following the model reported in [3] for CSRRs).

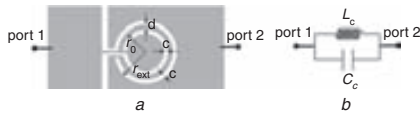


Fig. 1 Typical topology of OCSRR with relevant dimensions, and equivalent circuit model

a Typical topology of OCSRR
 b Equivalent circuit model
 Circuit only models resonator (i.e. capacitance to ground for microstrip implementations or other parasitic effects have not been considered)

In the work reported in this Letter, OCSRRs are periodically etched in the conductor strip of a microstrip line to implement lowpass filters with very sharp transition bands. As in elliptic filters [4], such abrupt cutoff is related to the presence of the transmission zero, caused by the OCSRRs. At the OCSRR's resonance, the line is opened and the injected power is reflected back to the source. Since the OCSRRs are electrically small particles, filter dimensions can be made small. Thus, such filters are of interest in applications where the main aim is to reduce dimensions and achieve extremely narrow transition bands. As shown later, filter design is simple. However, since the filter response does not follow any standard approximation (elliptic, Chebyshev, etc.), a spurious band typically appears close to the cutoff frequency. It is also shown that such spurious band can be easily eliminated by cascading further OCSRR sections tuned at the spurious band frequency.

Filter design: A prototype device example is reported in this Section to illustrate the possibilities of the approach. Specifically, a lowpass filter with the requirements depicted in Fig. 2 has been designed. The more severe requirement concerns the 40 dB rejection at 1.45 GHz with less than 3 dB insertion losses at 1.33 GHz. A substrate with dielectric constant $\epsilon_r = 11$, loss tangent $\tan\delta = 0.0023$ and thickness $h = 1.27$ mm has been considered (the metal film has thickness of $35 \mu\text{m}$). To achieve the above specifications with the smaller number of sections, it is convenient to design the OCSRRs with small inductance and

high capacitance, since this enhances the slope of the transition band. Obviously their resonance frequency must be set close to the frequency where 40 dB rejection is required. With the help of the model reported in [3], we have designed an OCSRR with a resonance frequency of $f_o = 1.84$ GHz. Owing to the presence of the ground plane metal (not accounted for in the model of [3]), we have optimised dimensions in order to adjust the OCSRR's resonance to the required value. The estimated value of the inductance and capacitance of the particle (inferred from curve fitting [1]) are $L_c = 1.55$ nH and $C_c = 4.78$ pF.

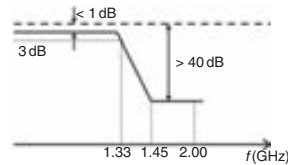


Fig. 2 Filter specifications

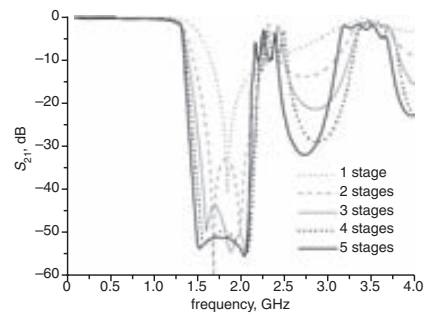


Fig. 3 Simulated transmission coefficients of designed lowpass filters with different number of OCSRR stages

Dimensions of OCSRRs are (in reference to Fig. 1), $c = 0.2$ mm, $d = 2.43$ mm, $r_{ext} = 3.5$ mm

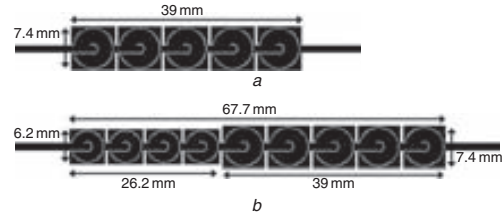


Fig. 4 Layout of fabricated lowpass filter prototype without and with spurious passband suppression

a Without spurious passband suppression
 b With spurious passband suppression
 Dimensions of four OCSRRs designed to suppress spurious band are (in reference to Fig. 1), $c = 0.2$ mm, $d = 1.83$ mm, $r_{ext} = 2.9$ mm. Other filter dimensions indicated

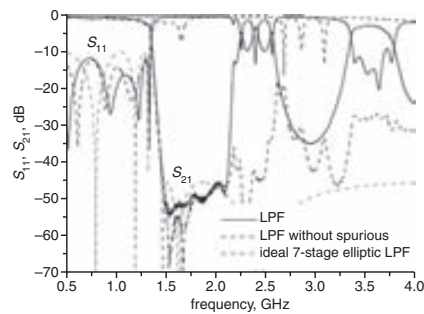


Fig. 5 Measured frequency responses of two fabricated lowpass filters shown in Fig. 4

Ideal frequency response of seven-stage elliptic filter included for comparison purposes

We have considered several lowpass filters with a different number of OCSRR stages. The simulated (by means of the Agilent Momentum commercial software) frequency responses of such filters are depicted in Fig. 3. As can be seen, five stages suffice to satisfy the specifications given in Fig. 2. The layout of the five-stage lowpass filter based on OCSRRs is depicted in Fig. 4a, and its measured frequency response is depicted in Fig. 5. This frequency response exhibits a spurious band relatively close to the 3 dB cutoff frequency. However, such spurious band can be easily suppressed by cascading additional OCSRR stages to the filter, each tuned at the onset of the spurious band. We have also fabricated this filter with spurious suppression (Fig. 4b). The measured frequency response (also depicted in Fig. 5) reveals that the spurious band is rejected. Thus, the out-of-band behaviour of the filter is improved by adding further OCSRR stages.

For comparison purposes, we have calculated the number of stages that are necessary to obtain the filter requirements by means of an elliptic response with a ripple of 0.5 dB [4]. An order-7 filter is required. The ideal response of such a filter (obtained from frequency and impedance transformation from the lowpass filter prototype) is also depicted in Fig. 5 for comparison.

Conclusion: We have demonstrated that OCSRRs are electrically small particles that can be applied to the design of compact microstrip lowpass filters with sharp cutoffs. We have designed and fabricated two prototype devices: one with five OCSRR stages; the other with four additional stages in order to achieve spurious-band elimination. The measured frequency responses of these prototypes are indicative of the possibilities of the proposed approach.

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F. Aznar, A. Vélez, J. Bonache, J. Menés and F. Martín (GEMMA/CIMITEC, Departament d'Enginyeria Electrònica, Universitat Autònoma de Barcelona, 08193 Bellaterra, Barcelona, Spain)

E-mail: francisco.aznar@uab.es

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Artículo M

Referencia:

F. Aznar, A. Vélez, M. Durán-Sindreu, J. Bonache and F. Martín, “Elliptic-function CPW Low-Pass Filters Implemented by Means of Open Complementary Split Ring Resonators (OCSRrs)”, *IEEE Microwave Wireless Components Letters*. Enviado.

Resumen del artículo:

En este artículo, se presenta por primera vez un filtro paso bajo elíptico implementado en tecnología coplanar (CPW) y *open complementary split ring resonators* (OCSRrs) grabados en la línea central. Los OCSRrs se comportan como un tanque resonante paralelo conectados en serie, mientras que parches metálicos grabados en la parte posterior del sustrato dan la capacidad en derivación necesaria para lograr la respuesta de la función elíptica. Se ha diseñado y fabricado un LPF elíptico de 5º orden para ilustrar las posibilidades de esta nueva propuesta. La respuesta frecuencial medida tiene un buen ajuste con la función elíptica ideal hasta el doble de la frecuencia de corte ($f_c = 1GHz$). Como los OCSRrs son resonadores eléctricamente pequeños, las dimensiones del filtro también son pequeñas (la longitud del dispositivo es $2.5cm$, es decir 0.13λ , donde λ es la longitud de onda guiada a f_c). Este trabajo ilustra las posibilidades de los OCSRrs para el diseño de filtros planares compactos y otros componentes de microondas.

Elliptic-function CPW Low-Pass Filters Implemented by Means of Open Complementary Split Ring Resonators (OCSRRs)

Francisco Aznar, *Student Member IEEE*, Adolfo Vélez, Miguel Durán-Sindreu, Jordi Bonache, *Member IEEE*, and Ferran Martín, *Senior Member IEEE*

Abstract— In this letter, an elliptic low-pass filter (LPF) implemented in coplanar waveguide (CPW) technology by etching open complementary split ring resonators (OCSRRs) in the central strip is presented for the first time. The OCSRRs behave as series connected parallel resonant tanks, whereas metallic patches etched in the back substrate side provide the required shunt capacitance to achieve the elliptic function response. An order-5 elliptic LPF has been designed and fabricated to illustrate the possibilities of this new approach. The measured frequency response is in good agreement with the ideal elliptic function up to twice the cutoff frequency ($f_c=1\text{GHz}$). Since OCSRRs are electrically small resonators, filter dimensions are also small (device length is 2.5cm , namely 0.13λ , where λ is the guided wavelength at f_c). This work is illustrative of the possibilities of OCSRRs for the design of compact planar filters and other microwave components.

Index Terms— Open complementary split ring resonator (OCSRR), Low Pass Filter (LPF), coplanar waveguide (CPW).

I. INTRODUCTION

Elliptic low-pass filters (LPFs) are of interest for the elimination of unwanted harmonics and spurious bands in microwave systems, especially in those applications where sharp cut-off and small dimensions are a severe requirement. Due to the presence of finite-frequency transmission zeros in the elliptic-function response, the synthesis of filters with simultaneously narrow transition bands and low order (and hence size) is possible. The two commonly used network structures for elliptic-function low-pass prototype filters are depicted in Fig. 1. These dual realizations prevent signal from being transmitted at the resonance frequencies of the different resonators.

Most elliptic-function LPFs have been implemented in microstrip technology [1]–[8] and described by the networks of Fig. 1. In this paper, novel elliptic LPFs described by the network of Fig. 1(a) and implemented in coplanar waveguide (CPW) technology are proposed. In a previous work, CPW

low-pass elliptic filters were proposed by etching interdigital capacitances in the central strip with an electrical short in one of the fingers in order to implement the series connected parallel resonators of the network of Fig. 1(a) [9]. In this paper, the series connected parallel resonators of the circuit of Fig. 1(a) are implemented by means of open complementary split ring resonators (OCSRRs), an electrically small open parallel resonator recently introduced by some of the authors [10]. The shunt capacitances are implemented by etching metallic patches in the back side of the substrate connected to the ground planes through vias. As will be shown, the reported OCSRR-based prototype device LPF exhibits state of the art performance and very compact size.

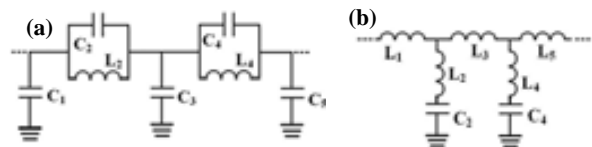


Fig. 1. Low-pass elliptic-function prototype filters with (a) series connected parallel resonators and (b) shunt connected series resonators. The circuits correspond to a 5th order prototype.

II. CPW TRANSMISSION LINES WITH SERIES CONNECTED OCSRRS

A typical topology of a series connected square-shaped OCSRR in a CPW transmission line is depicted in Fig. 2. As was discussed in [10], the structure can be described by means of an open parallel resonant tank, where the inductance, L , is the inductance of the metallic strip between the ring slots, and the capacitance, C , can be approximated by the capacitance of a square patch of side length l surrounded by a metallic plane at a distance c of its edge. As compared to the complementary split ring resonator (CSRR) [11], the inductance of the OCSRR is 4 times larger, whereas the capacitance is roughly the same (for identical dimensions and substrate). Thus, the OCSRR is electrically smaller than the CSRR by a factor of two, and hence the OCSRR is very interesting for size reduction in applications where open parallel resonators are needed. The formulas providing the element values of the OCSRR (inductance and capacitance) from its geometry are given in [12], and hence they are not reproduced here. According to this, the series connected OCSRR in a CPW transmission line can be simply modeled as shown in Fig. 2(b).

This work has been supported by MEC (Spain) through the project TEC2007-68013-C02 META-INNOVA, and by MCI (Spain) through the project EMET CSD2008-00066 of the CONSOLIDER Ingenio 2010 Program. Special thanks are also given to *Generalitat de Catalunya* for funding CIMITEC. Ferran Martín is also in debt with the ICREA Foundation for giving him an ICREA Academia Award.

F. Aznar, A. Vélez, M. Durán-Sindreu, J. Bonache and F. Martín are with GEMMA/CIMITEC (Departament d'Enginyeria Electrònica), Universitat Autònoma de Barcelona, 08193 BELLATERRA (Barcelona), Spain. E-mail: Ferran.Martin@uab.es.

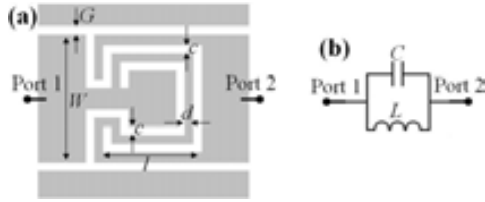


Fig. 2. Typical topology of a square-shaped OCSRR etched in the central strip of a CPW transmission line (a) and equivalent circuit model (b). The two terminals –ports– of the open resonator are indicated. Relevant dimensions are also indicated.

III. FILTER DESIGN AND RESULTS

We have considered the design of an order-5 elliptic-function LPF with a passband ripple of $L_A=0.1\text{dB}$, a cutoff frequency of $f_c=1\text{GHz}$ and an equal-ripple stop band starting frequency of 1.4085GHz (with stop band ripple of $L_A=39.59\text{dB}$) [13]. The element values corresponding to this elliptic filter response are (referred to the circuit of Fig. 1a): $C_1=3.20\text{pF}$, $C_2=0.58\text{pF}$, $L_2=9.44\text{nH}$, $C_3=5.02\text{pF}$, $C_4=1.73\text{pF}$, $L_4=6.87\text{nH}$, $C_5=2.41\text{pF}$. Filter implementation in CPW technology has been realized by means of two series connected OCSRRs (etched in the central strip) cascaded between electrically short (to avoid the parasitic effects of the series inductance) transmission line sections with a metallic patch etched in the back substrate side and connected to the ground planes by means of vias. These patches are necessary to obtain the high required values of the shunt capacitors.

The procedure to determine the dimensions of the metallic patches is as follows: we obtain the S-parameters of the short CPW transmission line section with the metallic patch below it (from electromagnetic simulation), and, from these results, we infer the shunt (capacitive) reactance of the π -circuit of such transmission line section (following standard formulas [14]). Then, we extract the capacitance and compare it to the nominal value. The procedure is repeated (by modifying the patch dimensions) until good agreement results. In practice, we have modified the length, l_p , of the patch, while its width has been set to a fixed value ($w_p=15.84\text{mm}$). The value of the patch capacitance as a function of its length is depicted in Fig. 3. Thus, Fig. 3 has been used to determine patch dimensions.

The dimensions of the OCSRRs have been determined by forcing their resonance frequencies to the required values, and by forcing the susceptance slopes to be identical to those of the lumped element resonators (of the elliptic-function model) at resonance. Depending on the values of the resonator inductance and capacitance, it is necessary to substantially modify the topology of the OCSRR in order to obtain the desired values of the elements. Thus, the formulas that link the geometry of the OCSRR and the element values (reported in [12]), are only used as reference. Optimization is required to provide a good description of the resonators in the vicinity of resonance. Fig. 4 depicts resonator's susceptance for the two parallel resonators of the filter of Fig. 1(a) in the vicinity of resonance. Also in this figure are depicted the susceptances that have been inferred from electromagnetic simulation of the optimized OCSRR topologies (shown in the inset of the figure). As required, there is good agreement between the two pair of curves in the vicinity of resonance.

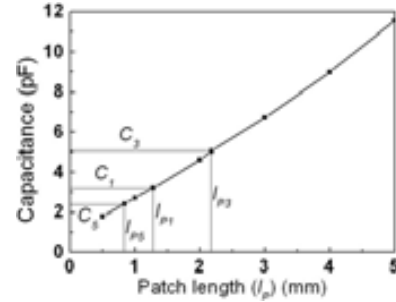


Fig. 3. Capacitance of the patch as a function of its length, obtained as described in the text. The Rogers RO3010 substrate with measured dielectric constant $\epsilon_r=11.5$ and thickness $h=0.635\text{mm}$ has been considered.

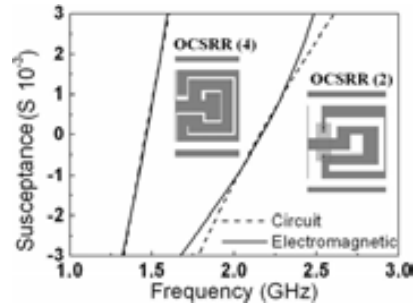


Fig. 4. Susceptance of the two series connected parallel resonators of the designed elliptic-function LPF. The susceptance inferred by means of electromagnetic simulation has been obtained from the optimized layout topologies included in the figure. The Rogers RO3010 substrate with measured dielectric constant $\epsilon_r=11.5$ and thickness $h=0.635\text{mm}$ has been considered.

The layout and photograph of the fabricated LPF are depicted in Fig. 5. The device has been fabricated on the Rogers RO3010 substrate with measured dielectric constant $\epsilon_r=11.5$ and thickness $h=0.635\text{mm}$. Besides the patch capacitances, there is an additional strip in the back substrate side that connects (through metallic vias) the external metallic regions of OCSRR (2) (the one which exhibits the upper resonance frequency). With this strategy, we suppress a parasitic slot mode that is otherwise generated in this resonator, as revealed from the simulated magnetic currents. This parasitic mode obscures the first resonance frequency of the OCSRR (2). Thus, by etching such additional metallic strip, the stop band filter response is improved, and the two transmission zeros can be clearly identified.

The simulated (inferred from the Agilent Momentum commercial software) and measured (obtained by means of the Agilent E8364B vector network analyzer) frequency responses of the device are shown in Fig. 6. The circuit simulation of the model of Fig. 1(a) with the above indicated element values (corresponding to the ideal elliptic-function response) is also depicted in Fig. 6. The agreement between experimental data and the elliptic function response is good up to approximately 2GHz. At higher frequencies, the mismatch is due to the fact that the lumped element models of the different filter elements (OCSRRs and shunt capacitors) fail. Measured insertion losses in the pass band differ in less than 0.3dB to the ideal response, and return losses are better than 15dB. The frequency response exhibits a sharp cutoff, with a stop band rejection better than 25dB up to roughly

4.5GHz. It is also remarkable that filter length is small i.e., 2.5cm, which corresponds to 0.13λ , where λ is the guided wavelength at the cutoff frequency. This combination of size (relative to the filter cutoff frequency) and performance (especially for which concerns the width of the stop band) is not found in the filters provided in references [1-8]. We have also compared the dimensions and performance of the filter of Fig. 5 with other LPFs based on CSRRs [15,16]. The filter of [15] also provides good stop band rejection, but it is much larger. The filter reported in [16] is also much larger and in-band performance (insertion and return losses) is clearly inferior as compared to the filter proposed in the present paper.

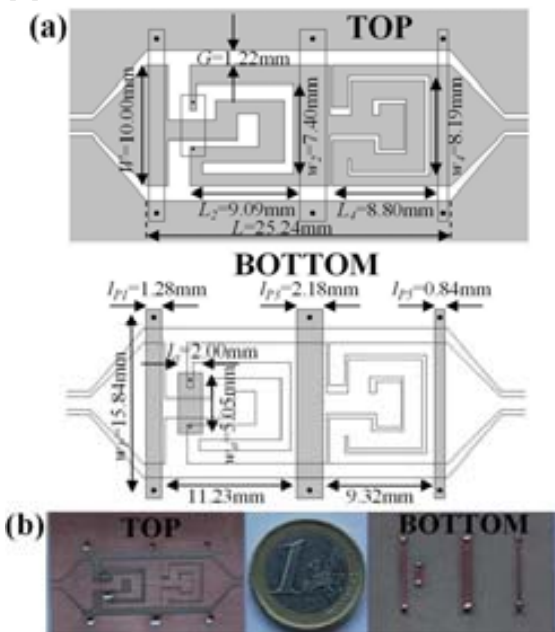


Fig. 5. Layout (a) and photograph (b) of the fabricated elliptic-function LPF.

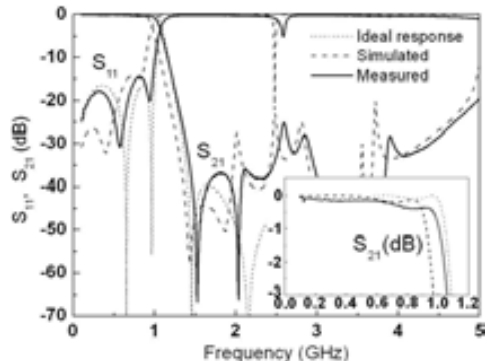


Fig. 6. Simulated and measured frequency response of the fabricated filter. The ideal elliptic-function response is also depicted. The detail of the in-band insertion losses are depicted in the inset. The measured in-band losses are better than those reported in [3] and [7] (in such references a graph similar to that of the inset is reported).

IV. CONCLUSION

In conclusion, novel elliptic-function low pass filters based on OCSRRs and implemented in CPW technology have been

proposed for the first time. The OCSRRs, acting as series connected resonant tanks, provide the transmission zeros of the filter response, whereas the shunt connected capacitances of the elliptic function circuit model have been implemented by means of metallic patches etched in the back substrate side and connected to the ground planes through vias. An order-5 elliptic low pass filter has been designed and fabricated. The measured frequency response has been found to be in very good agreement with the ideal elliptic function up to twice the cutoff frequency. The two transmission zeros of the ideal frequency response can be clearly identified in measurement. Measured in-band losses are good, and the stop band exhibits a rejection level better than 25dB up to 4.5 times the cutoff frequency. The main conclusion of this paper is that OCSRRs are useful for the design of high performance compact elliptic LPFs in CPW technology.

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Apéndice B

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