Abstract  Metamaterial transmission lines are artificial lines consisting on a host conventional line (such as a microstrip or a coplanar waveguide) loaded with reactive elements. The main relevant characteristic of these propagating structures is the fact that, due to the greater number of parameters (as compared to conventional lines), it is possible to tailor (engineer) their dispersion characteristics to achieve certain functionalities not achievable though conventional lines. Enhanced bandwidth components and multiband components are the main benefits of dispersion engineering with metamaterial transmission lines. In this paper, dispersion engineering will be carefully reviewed, and several examples illustrative of the application of this technique to microwave circuit design will be reported.

Dispersion engineering with resonant-type metamaterial transmission lines

Gerard Sisó*, Marta Gil, Francisco Aznar, Jordi Bonache, and Ferran Martín

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

Received: 13 June 2008, Revised: 26 July 2008, Accepted: 29 July 2008
Published online: 12 September 2008

Key words: metamaterials; dispersion engineering; split ring resonators; multiband components; enhanced bandwidth components

PACS: 41.20.Jb, 77.22.-d, 81.05.-t, 84.40.-x, 84.40.Az, 84.40.Dc

1. Introduction

Metamaterial transmission lines are artificial lines consisting on a host (conventional) line (such as a microstrip or a coplanar waveguide) loaded with reactive elements. These propagating structures can be considered as one-dimensional (1D) metamaterials as long as they are structured on a sub-wavelength scale and their relevant characteristics (impedance and phase) can be controlled beyond what can be achieved in conventional transmission lines. However, metamaterial transmission lines are artificial lines, not artificial materials. The terminology, which is an object of some controversy, simply reflects that these artificial lines, formerly proposed in 2002 [1–3], are inspired on metamaterials, exhibit similar properties, and, in some cases, are fabricated using identical constituent particles [4, 5].

Let us clarify these points before considering in detail the main topic of this paper, namely, dispersion engineering. To this end, it is necessary to clearly define the metamaterial concept. Metamaterials are artificial materials, consisting on sub-wavelength periodic (or quasi-periodic) inclusions (“atoms”) of metals and/or dielectrics, whose electromagnetic, or optical, properties can be controlled through structuring, rather than through composition (see some of the recent published books on metamaterials [6–11]). Since the period is smaller than wavelength, effective (continuous) media properties are achieved, and it is possible to obtain properties beyond those available in nature. For instance, anisotropic metamaterials with simultaneously negative effective permeability ($\mu_{\text{eff}}$) and permittivity ($\varepsilon_{\text{eff}}$) in a certain frequency band were already implemented in 2000 by Smith and co-workers [12] (from this seminal
work, many efforts have been focused on the synthesis of isotropic metamaterials [13, 14] and metamaterials at optical wavelengths [15, 16]). These double negative (that is, with $\mu_{\text{eff}} < 0$ and $\varepsilon_{\text{eff}} < 0$) artificial materials exhibit left-handed wave propagation, which was anticipated by Veselago in 1968 [17], and is not present among natural materials. Recently, an invisible cloak was designed at microwave frequencies [18], this being a clear and genuine product of metamaterial research (for this, it is necessary to point-to-point control the material characteristics to obtain a graded anisotropic response).

Now, the key question is: what do metamaterial transmission lines have in common with metamaterials, to justify the term? Probably we can agree that metamaterial transmission lines are artificial lines with controllable characteristics. Furthermore, these artificial lines can be designed to exhibit left-handed wave propagation in certain frequency bands. Such lines are normally implemented by means of lumped or semi-lumped reactive elements. Since these elements are electrically small, the conditions to achieve homogeneity can also be achieved, that is a small period compared to signal wavelength. However, we would like to discuss a little bit this latter aspect: homogeneity. To achieve homogeneity in a material we need a small unit cell (as compared to wavelength) and also a repetition of a minimum number of them. Only under these conditions we can talk in terms of effective constitutive parameters $\mu_{\text{eff}}$ and $\varepsilon_{\text{eff}}$. However, homogeneity is not a fundamental requirement in transmission lines. Indeed homogeneity can only be achieved in a certain region of the allowed band, as will be later discussed (it is in this region where $\mu_{\text{eff}}$ and $\varepsilon_{\text{eff}}$ do have a physical meaning). However, from the point of view of microwave circuit design, the advantages of metamaterial transmission lines rely on miniaturization and on the possibility to control the dispersion diagram and characteristic impedance, rather than on homogeneity. Thus, in this paper metamaterial transmission lines are defined as artificial lines, consisting on a host line loaded with reactive elements, with controllable characteristics. Homogeneity is not considered to be a requirement for such lines. Notice that according to this, there is no need for a minimum number of unit cells to implement these artificial lines. Indeed, in most of the cases, a single cell is considered since this reduces line dimensions, as will be shown later.

With regard to the implementation of metamaterial transmission lines, there are two main approaches: LC-loaded lines [1–3], and the resonant type approach [4, 5]. In LC-loaded lines, a host line is loaded with series capacitances and shunt inductances. These lines can be implemented by using lumped loading elements, or, alternatively, by means of semi-lumped planar components such as series gaps, interdigital capacitors, grounded stubs or vias (see Fig. 1). Resonant-type metamaterial transmission lines can be implemented by loading a host line with split ring resonators (SRRs) and shunt inductive elements [4], or, alternatively, by loading a host line with complementary split ring resonators (CSRRs) and series capacitances [5, 19] (Fig. 2). Both LC-loaded lines and resonant type metamaterial transmission lines have in common with metamaterials at optical wavelengths [15, 16]). These double negative (that is, with $\mu_{\text{eff}} < 0$ and $\varepsilon_{\text{eff}} < 0$) artificial materials exhibit left-handed wave propagation, which was anticipated by Veselago in 1968 [17], and is not present among natural materials. Recently, an invisible cloak was designed at microwave frequencies [18], this being a clear and genuine product of metamaterial research (for this, it is necessary to point-to-point control the material characteristics to obtain a graded anisotropic response).

**Figure 1** Typical topologies (top view) 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; (c) microstrip structure loaded with grounded stubs and interdigital capacitors. In the microstrip structures (b and c), the ground plane is not shown.

**Figure 2** Typical layout (unit cell) of a metamaterial transmission line loaded with SRRs (a) and CSRRs (b). The relevant dimensions of SRRs or CSRRs are indicated in (c). In (a), the host line is a CPW, also loaded with shunt connected strips; in (b) the host line is a microstrip structure also loaded with series capacitive gaps. In (a), the SRRs (in black) are etched in the back substrate side; in (b), the ground plane is depicted in gray.
terial transmission lines exhibit similar characteristics and dispersion (as will be shown in the next section), and both line types are useful for the implementation of beyond the state-of-the-art microwave and millimeter wave circuits based on dispersion engineering. This is the main claim of this review paper.

2. Metamaterial transmission lines: analysis and circuit models

Metamaterial transmission lines consist on a host line loaded with reactive elements. The unit cell of such lines can be described through a π- or a T-circuit model with lumped elements (Fig. 3). This is valid as long as the unit cell length is small as compared to signal wavelength at the frequencies of interest. According to this description, the dispersion in these lines is given by [20]:

\[
\cosh \gamma l = 1 + \frac{Z_s(\omega)}{Z_p(\omega)} \tag{1}
\]

whereas the characteristic (or image) impedance is given by:

\[
Z_B = \sqrt{Z_s(\omega) [Z_s(\omega) + 2Z_p(\omega)]} \tag{2}
\]

for a T-circuit, and by

\[
Z_B(\omega) = \sqrt{\frac{Z_s(\omega)Z_p(\omega)/2}{1 + \frac{Z_s(\omega)}{2Z_p(\omega)}}} \tag{3}
\]

for a π-circuit. \(\gamma\) is the complex propagation constant, \(l\) is the unit cell length, and \(Z_s(\omega)\) and \(Z_p(\omega)\) are the series and shunt impedances of the T- or π-circuit models. If losses are negligible (this assumption will be considered along this paper), the complex propagation constant, \(\gamma = \alpha + j\beta\) (where \(\alpha\) is the attenuation constant and \(\beta\) is the phase constant), is either purely real (i.e., \(\beta = 0\)) or imaginary (i.e., \(\alpha = 0\)). In the allowed bands, \(\alpha = 0\) and (1) can be expressed as:

\[
\cos \beta l = 1 + \frac{Z_s(\omega)}{Z_p(\omega)} \tag{4}
\]

whereas in the forbidden bands \(\beta = 0\) and (1) reduces to:

\[
\cosh \alpha l = \left| 1 + \frac{Z_s(\omega)}{Z_p(\omega)} \right| \tag{5}
\]

From (4) and (5) it follows the well known result that the sign of the series and shunt reactances are different in those regions where signal propagation is allowed. Let us now consider in detail the implications of the sings of these reactances (in the allowed regions). Two possibilities arise:

(i) \(\chi_s > 0\) and \(\chi_p < 0\), and (ii) \(\chi_s < 0\) and \(\chi_p > 0\). In the first case, the group and the phase velocity have the same sign and wave propagation is forward. In the second case, the group and phase velocities are antiparallel and wave propagation is backward (or left handed). This can be demonstrated by obtaining the derivative of \(\phi = \beta l\) with frequency from Eq. (4):

\[
\frac{d\phi}{d\omega} \bigg|_{\omega_o} = \frac{1}{\sin \phi} \left( \frac{\chi_p}{\omega_o} \frac{d\chi_s}{d\omega} \bigg|_{\omega_o} + \chi_s \frac{d\chi_p}{d\omega} \bigg|_{\omega_o} \right) \tag{6}
\]

Taking into account that the derivative of any reactance with frequency must be always positive (Foster theorem [21]), inspection of (6) reveals that if \(\chi_s > 0\) and \(\chi_p < 0\), \(\phi\) and its derivative with frequency have the same sign, and hence do the phase and group velocities. Conversely, if \(\chi_s < 0\) and \(\chi_p > 0\), \(\phi\) and its derivative with frequency have opposite sign, and the phase and group velocities are anti-parallel.

2.1. Transmission line models for the forward and backward transmission lines

From the previous analysis, it follows that the simplest transmission line model for a forward wave transmission line is the well known ladder network model of a conventional line, whereas the dual circuit, consisting on a cascade of series capacitances and shunt inductances is the simplest model of a left handed line. Fig. 4 depicts the simplest models (unit cell) of these lines as well as the simplest T-circuit models corresponding to non-propagating structures, with indication of the sign of the series and shunt reactance.

Let us now analyze the dispersion and characteristic impedance of the forward and backward transmission line models of Fig. 4. Application of Eqs. (1) and (2) gives (the sub-indices \(L\) and \(R\) are used to distinguish between the left handed – backward wave – and right handed – forward
wave – structures):  

\[
\cos \beta_R l = 1 - \frac{LC}{2} \omega^2, \tag{7}
\]

\[
Z_{BR} = \sqrt{\frac{L}{C} \left(1 - \frac{\omega^2}{\omega_{cR}^2}\right)}, \tag{8}
\]

\[
\cos \beta_L l = 1 - \frac{1}{2LC\omega^2}, \tag{9}
\]

\[
Z_{BL} = \sqrt{\frac{L}{C} \left(1 - \frac{\omega^2}{\omega_{cL}^2}\right)}, \tag{10}
\]

where \(\omega_{cR} = 2/(LC)^{1/2}\) and \(\omega_{cL} = 1/(2LC)^{1/2}\) are angular cut-off frequencies for the forward and backward wave transmission line structures, respectively. The dispersion diagram, as well as the dependence of \(Z_B\) on frequency are depicted in Fig. 5. Transmission is limited to those frequency intervals that make the phase constant and the characteristic impedance to be real numbers. It is worth mentioning that frequency dispersion is present in both structures. Even though the circuit of the first quadrant of Fig. 4 models an ideal lossless forward transmission line, where dispersion is absent, actually this circuit is only valid for frequencies satisfying \(\omega \ll \omega_{cR}\), i.e., in the long wavelength limit (corresponding to those frequencies where wavelength for guided waves satisfies, \(\lambda_g \gg l\), i.e. homogeneity conditions). To correctly model an ideal lossless transmission line at higher frequencies, we simply need to reduce the period of the structure, and accordingly the per-section inductance and capacitance of the line, \(L\) and \(C\), with the result of a higher cut-off frequency. Thus, the lumped element T-circuit model can properly describe ideal transmission lines without dispersion. To this end we simply need to select the period such that the long wavelength limit approximation holds. Under this approximation, Eqs. (7) and (8) become:

\[
\beta_R = \omega \sqrt{L'C'}, \tag{11}
\]

and

\[
Z_{BR} = \sqrt{\frac{L'}{C'}} \equiv Z_{lw} \tag{12}
\]

where \(L'\) and \(C'\) are the per-unit length inductance and capacitance of the transmission line (in Eq. (12), \(L'\) and \(C'\) can be replaced by \(L\) and \(C\) with no effect). From (11), the phase and group velocities of the forward transmission line

Figure 4 (online color at: www.lpr-journal.org) Simplest T-circuit models (unit cell) of the forward and backward wave transmission lines with indication of the sign of the series and shunt reactances. Non-propagating structures, where such reactances have the same sign are also indicated. Actually, we have represented in the horizontal axis the shunt susceptance \(B_p = -1/\chi_p\).

Figure 5 Typical dispersion diagram of a forward (a) and backward (b) transmission line model. The dependence of the normalized \((Z_B/Z_{lw})\) Bloch impedance with frequency is shown in (c) and (d) for the forward and backward lines, respectively. In (a) and (b), the portion of the curves in bold correspond to power propagation from left to right, as is usually considered.
whereas for the backward transmission line, the constitutive parameters are:
\[ \varepsilon_{\text{eff}} = -\frac{1}{\omega^2 L}, \quad \mu_{\text{eff}} = -\frac{1}{\omega^2 C} \]
and they are both negative, a sufficient condition to obtain left handed wave propagation. The sign of the effective permeability and permittivity is also indicated in Fig. 4.

2.2. The composite right/left handed (CRLH) transmission line

The composite right/left handed transmission line concept was first introduced in [22] to explain the forward and backward wave transmission in practical metamaterial transmission lines, that is, host lines loaded with series capacitances and shunt inductances (see Fig. 1). The model of such CL-loaded lines is depicted in Fig. 6. At low frequencies, the loading elements are dominant and wave propagation is backward. However, at high frequencies, line parameters are dominant and forward wave transmission arises. The dispersion relation and the characteristic impedance of these CRLH lines are (from 1 and 2):
\[ \cos \beta l = 1 - \frac{\omega^2}{2\omega_R} \left( \frac{1}{\varepsilon_{\text{eff}}} \right) \left( 1 - \frac{\omega^2}{\omega_p^2} \right), \quad (25) \]
\[ Z_B = \sqrt{\frac{L_R}{C_R}} \left( 1 - \frac{\omega^2}{\omega_s^2} \right) - \frac{L_R^2 \omega^2}{4 \left( 1 - \frac{\omega^2}{\omega_p^2} \right)^2}, \quad (26) \]
where the following variables
\[ \omega_R = \frac{1}{\sqrt{L_R C_R}}, \quad (27) \]
\[ \omega_L = \frac{1}{\sqrt{L_L C_L}}, \quad (28) \]
and the series and shunt resonance frequencies
\[ \omega_s = \frac{1}{\sqrt{L_R C_L}}, \quad (29) \]
\[ \omega_p = \frac{1}{\sqrt{L_L C_R}}, \quad (30) \]
have been introduced to simplify the mathematical formulas. Expressions (25) and (26) are depicted in Fig. 7. The left handed and right handed band can be easily identified. The gap in between such transmission bands is delimited by the following frequencies:
\[ \omega_{G1} = \min(\omega_s, \omega_p), \quad (31) \]
\[ \omega_{G2} = \max(\omega_s, \omega_p). \quad (32) \]
In the long wavelength limit, Eq. (25) rewrites as:
\[
\beta = \frac{s(\omega)}{l} \sqrt{\frac{\omega^2}{\omega_R^2} \left(1 - \frac{\omega_s^2}{\omega^2}\right) \left(1 - \frac{\omega_p^2}{\omega^2}\right)}
\]
(33)
where \(s(\omega)\) is the following sign function:
\[
s(\omega) = \begin{cases} 
-1 & \text{if } \omega < \min(\omega_s, \omega_p) \\
+1 & \text{if } \omega > \max(\omega_s, \omega_p)
\end{cases}
\]
(34)
From the phase constant (Eq. 33), the phase and group velocities can be easily inferred, these velocities being of opposite sign in the left handed band and both being positive in the right handed band.

With regard to the constitutive parameters for the CRLH transmission line, they can be inferred as previously indicated, i.e.:
\[
\varepsilon_{\text{eff}} = \frac{C_R}{l} - \frac{1}{\omega^2 L_L l},
\]
(35)
\[
\mu_{\text{eff}} = \frac{L_R}{l} - \frac{1}{\omega^2 C_L l},
\]
(36)
and they can be positive or negative, depending on the frequency range.

One particular case of interest is the balanced case, where the series and shunt resonance coincide and the gap collapses. In this case there is a continuous transition between the left handed and the right handed bands, and the group velocity is finite at the so called transition frequency \(\omega_o = \omega_s = \omega_p\). It is worth mentioning that the characteristic impedance varies very smoothly in the vicinity of the transition frequency, where it takes the maximum value given by:
\[
Z_B = \sqrt{\frac{L_R}{C_R}} = \sqrt{\frac{L_L}{C_L}}
\]
(37)

The typical dispersion diagram and characteristic impedance of a balanced CRLH line are depicted in Fig. 8.

### 2.3. Resonant type CRLH metamaterial transmission lines

Those lines with unit cell depicted in Fig. 2 do also exhibit CRLH behaviour. Let us briefly present the circuit model of both the SRR- and the CSRR-based lines and analyze their relevant characteristics.

#### 2.3.1. SRR-loaded lines

The circuit model of the unit cell of SRR-loaded CRLH lines is depicted in Fig. 9a [23]. Actually, the circuit model depicted in Fig. 9a is an improved version of the first proposed model of such lines [4]. In this model, \(L\) and \(C\) are the inductance and capacitance of the line, \(L_p\) is the inductance of the shunt connected strips (or vias in a microstrip...
The π-circuit model depicted in Fig. 9b is actually very similar to the T-circuit model depicted in Fig. 6. It exhibits CRLH behaviour, and the characteristic impedance exhibits similar features. However there is one important difference: the presence of a transmission zero close (to the left) to the left handed transmission band. This transmission zero is at the origin in the model of Fig. 6. The transmission zero at a finite frequency has been used by the authors for the design of band pass filters with sharp cut-off transitions [25]. Obviously, the structure can be designed to be balanced.

2.3.2. CSRR-loaded lines

The circuit model of the unit cell of CSRR-loaded CRLH transmission lines is depicted in Fig. 10a [26]. In this circuit model, $C_s$ is the series capacitance of the gap or interdigital capacitor, $C_f$ is the fringing capacitance, $C_L$ is the line capacitance, $L$ is the line inductance and the CSRR are described by the resonant tank $L_c - C_c$. This circuit can be transformed to the circuit shown in Fig. 10b, where the following transformations apply:

\[
C_g = 2C_s + C_{\text{par}}, \quad C = \frac{C_{\text{par}}(2C_s + C_{\text{par}})}{C_s},
\]

with $C_{\text{par}} = C_L + C_f$. Notice that the circuit depicted in Fig. 10b is the formerly proposed circuit of CSRR-loaded lines [27]. Such circuit is correct, but the interpretation of some of the parameters was inaccurate. Thus, $C_g$ actually accounts for the series capacitance plus the capacitance of the line and the fringing capacitance, and $C$, the coupling capacitance, depends also on the series capacitance. Indeed, this capacitance can be very large if the gap is wide, as has been verified [28]. Notice that in the limit where $C_C$ is large (as compared to $C_{\text{par}}$), $C_g$ and $C$ have the interpretation given in [27], but these conditions are not easily achievable in practice.

The model of Fig. 10b is very similar to the model of Fig. 6. CSRR-loaded lines do also exhibit CRLH behaviour and, as occurs in SRR loaded lines, these lines exhibit a transmission zero to the left of the left handed band [29]. Since CSRR-loaded CRLH transmission lines are the resonant-type metamaterial transmission lines that have been more exhaustively used in microwave circuit design based on dispersion engineering, we will consider these lines in more detail. The dispersion relation and the characteristic impedance of these CSRR-based CRLH lines are (from 1 and 2):

\[
\cos \beta l = 1 + \frac{C}{2C_g} \left( 1 - \frac{\omega_s^2}{\omega_p^2} \right) \left( 1 - \frac{\omega_s^2}{\omega_f^2} \right), \quad (44)
\]

\[
Z_B = \sqrt{\frac{L}{C_g} \left( 1 - \frac{\omega_p^2}{\omega_C^2} \right) - \frac{L^2 \omega_f^2}{4} \left( 1 - \frac{\omega_s^2}{\omega_f^2} \right)^2 + \frac{L}{C} \left( 1 - \frac{\omega_s^2}{\omega_f^2} \right)}, \quad (45)
\]

where $\omega_s$ and $\omega_p$ are the series and shunt resonance, respectively. By forcing the balance condition, that is, $\omega_s = \omega_p = \omega_C$, the gap between the left handed and the right handed region disappears, and the characteristic impedance exhibits continuity in the vicinity of the transition frequency.
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 dielectric constant $\varepsilon_r = 10.2$ and thickness $h = 1.27$ mm. Dimensions are: line width $W_m = 0.8$ mm, internal radius $r = 6.3$ mm, ring width $c = 0.4$ mm and ring separation $d = 0.2$ mm; the interdigital capacitor, formed by 28 fingers separated 0.16mm, has been used to achieve the required capacitance value. Reprinted with permission from [29]. Copyright 2007 IEEE.

$\omega_o$ (notice however that the maximum value of the characteristic impedance is slightly shifted to the right of $\omega_o$). By balancing the line, the structure may exhibit a very broadband [29]. This, combined with the presence of a transmission zero, has been used for the design of ultra wide band filters [30, 31]. Fig. 11a depicts the layout of a unit cell CSRR-loaded CRLH microstrip line that has been designed to be balanced. The measured and simulated dispersion, shown in Fig. 11b, reveals that the structure is roughly balanced (perfect balance is difficult to achieve in practice). The transmission coefficient is depicted in Fig. 11c. As expected, a broad band that includes the left handed and the right handed regions, and a sharp cut-off at the lower band edge are obtained. Other unbalanced CSRR-loaded CRLH lines have been reported in the recent literature, and it has been shown that the model of Fig. 10 describes accurately the frequency response up to the second band [32].

To end this section we would like to mention that SRR-loaded and CSRR-loaded metamaterial transmission lines exhibit a very similar behaviour. It has been found that the effect of varying the strip width in SRR-loaded CPWs is the same as varying the gap width in CSRR-loaded microstrip lines. This is not actually surprising since these type of lines are roughly dual, and their equivalent circuits are formally circuit duals. This aspect is discussed in [33].

3. Dispersion engineering: principles and applications

Dispersion engineering means to tailor the dispersion characteristics of metamaterial transmission lines to achieve certain functionalities, which are normally not achievable by means of conventional lines. Dispersion engineering can be mainly applied to the design of enhanced bandwidth components and to the design of multiband components. Let us now see the principles behind these applications, and then some illustrative results.

3.1. Principles of dispersion engineering

3.1.1. Bandwidth enhancement: principles and limitations

Let us start by discussing the principles behind bandwidth enhancement. The operative bandwidth of microwave components is given by that frequency interval where the required characteristics are satisfied within certain limits. In distributed circuits, bandwidth is limited by the phase shift experienced by transmission lines and stubs when frequency is varied from the nominal operating value. In a conventional transmission line of length $l$, the phase of the line at a certain angular frequency $\omega_o$ is given by:

$$\phi_o = \beta l = \frac{l}{v_p} \omega_o$$

where $v_p$ is the phase velocity of the line. According to the previous comment, bandwidth is intimately related to the derivative of $\phi$ with frequency, also known as group delay. From this, it follows that the shorter the line is, the broader the bandwidth becomes. In other words, bandwidth is inversely proportional to the required phase of the line, which is dictated by the design. This means that the operative bandwidth is not an easily controllable parameter in conventional distributed circuits. The reason is the limited number of degrees of freedom of conventional transmission lines. However, in metamaterial transmission lines, the loading elements provide additional parameters and certain control of the phase response is expected. Namely, one expects that the required phase does not dictate bandwidth.
It has been demonstrated for instance that in series power dividers based on 360° phase shifting lines, bandwidth can be enhanced by using artificial lines [34]. The key question is: is it possible to improve the bandwidth, i.e. decrease the group delay, of conventional lines by means of metamaterial transmission lines regardless of the required phase shift? Let us discuss now this aspect (a detailed analysis is given in [35]).

Assuming that for a certain transmission line, the required phase and characteristic impedance at the operating frequency are \( \phi_0 \) and \( Z_0 \), Eqs. (2) and (4) can be inverted, and the series and shunt impedances must take the following values at the design frequency (a T-circuit has been assumed for the unit cell):

\[
Z_s = Z_o \sqrt{\frac{\cos \phi_0 - 1}{\cos \phi_0 + 1}}, \quad (47)
\]

\[
Z_p = \frac{Z_o}{\sqrt{\cos^2 \phi_0 - 1}}. \quad (48)
\]

On the other hand, the derivative of the phase shift with frequency is given by (6). By introducing (47) and (48) in (6), and after some simple calculation, the derivative of phase at the operating frequency is:

\[
\frac{d\phi}{d\omega} \bigg|_{\omega_o} = \frac{1}{Z_o} \left( \frac{d\chi_s}{d\omega} \bigg|_{\omega_o} - (\cos \phi_0 - 1) \frac{d\chi_p}{d\omega} \bigg|_{\omega_o} \right). \quad (49)
\]

The two right hand side terms in (49) are effectively positive since according to the Foster reactance theorem [21], the derivative of any purely reactive reactance or susceptance with frequency is always positive, and \( \cos \phi_0 \leq 1 \). To enhance bandwidth, we must force the derivative of \( \phi \) with frequency to be as small as possible. Thus, according to (49), the frequency derivatives of the series and shunt reactances must be both as small as possible. From the Foster theorem (see the appendix of [35]), it follows that the optimum metamaterial transmission line model to minimize Eq. (49) is the canonical dual transmission line. Hence, it is convenient to calculate Eq. (49) for this particular case. By introducing

\[
\chi_s = -j \frac{1}{2C_L \omega}, \quad (50)
\]

\[
\chi_p = jL_L \omega. \quad (51)
\]

in (49), we obtain, after some straightforward calculation:

\[
\frac{d\phi}{d\omega} \bigg|_{\omega_o} = \frac{2}{\omega_o} \sqrt{\frac{1 - \cos \phi_0}{1 + \cos \phi_0}} = \frac{2}{\omega_o} \tan \frac{|\phi_o|}{2}. \quad (52)
\]

Thus for the optimum transmission line model (in terms of the operative bandwidth) the derivative of phase with frequency does not depend on the required characteristic impedance. It is simply determined by the operating frequency and phase.

It is important to take into account that Eq. (52) is the derivative of the phase with frequency corresponding to a single unit cell. If a number \( N \) of cells is used to obtain the nominal phase shift, \( \phi_o \), then the phase shift of either cell is \( \phi_o / N \), and Eq. (52) rewrites as:

\[
\frac{d\phi}{d\omega} \bigg|_{\omega_o} = \frac{2}{\omega_o} N \tan \frac{|\phi_o|}{2N}. \quad (53)
\]

A simple analysis of Eq. (11) reveals that \( d\phi/d\omega \) decreases as \( N \) increases, namely,

\[
\frac{2}{\omega_o} N \tan \frac{|\phi_o|}{2N} > \frac{2}{\omega_o} (N + 1) \tan \frac{|\phi_o|}{2(N + 1)} \quad (54)
\]

for \( N = 1, 2, \ldots, \infty \). Therefore, in terms of bandwidth, the optimum solution is an \( N \)-stage artificial transmission line with \( N \to \infty \). In this case, (53) can be expressed as:

\[
\frac{d\phi}{d\omega} \bigg|_{\omega_o} = |\phi_o| \omega_o. \quad (55)
\]

According to (55) the derivative of phase with frequency for a dual transmission line in the considered limiting case \( (N \to \infty) \) is identical to that of a conventional line with identical phase (but different sign) at the same frequency. Therefore, from this result we may conclude that in those applications where the sign of the phase shift is irrelevant (for instance in 90° impedance inverters and many microwave components based on them [36]), it is not possible to enhance bandwidth by using artificial lines. If the number of cells is limited to a finite number, the derivative of phase with frequency increases, and the operative bandwidth is degraded, as compared to that of a conventional transmission line. Bandwidth improvement can be obtained if the sign of the phase is relevant. In this case we have to compare the dual transmission line with phase \( \phi_o \) (\( \phi_o \) being negative) with a conventional transmission line with equivalent phase shift, that is \( 2\pi + \phi_o \). From this comparison, it is obvious (from (55)) that as long as the required phase shift is higher than \( \pi \) (or lower than \(-\pi\)), the dual transmission line exhibits smaller phase dependence with frequency and hence it exhibits a better solution in terms of bandwidth. If the number of stages is limited to a finite number, then the limiting phase above which the artificial lines exhibit a better phase response (lower derivative) is no longer \( \pi \) rad. For a single stage dual transmission line, such limiting phase can be inferred by simply forcing (52) to be \((2\pi + \phi_o)/\omega_o\). It gives \( \phi_o = 0.7\pi \) rad, or a positive phase (conventional line of 1.3\pi rad). It means that in applications where the sign of the phase shift is fundamental, by using a single-stage dual transmission line, bandwidth can be improved if the required (positive) phase shift is higher than 1.3\pi rad.

As it has been highlighted above, the optimum metamaterial transmission line for bandwidth enhancement is the dual transmission line. Thus, the phase shifts (\( \pi \) rad for the infinite stage structure and 1.3\pi rad for the one-stage
transmission line) below which the dual transmission line is not able to provide an improved bandwidth are fundamental limits. Namely, with CRLH transmission lines (either CL-loaded or resonant type) this limits are even lower. However, in spite of this, bandwidth enhancement is possible since many microwave components are based on phase differences between transmission lines. To enhance bandwidth, it is necessary that the dispersion characteristics exhibit similar slopes in the vicinity of the operating frequency, and this can be achieved though the considered artificial CRLH lines, as will be shown in the Sect. 3.2 (see Fig. 12).

3.1.2. Dual-band operation

Dual-band components are components exhibiting certain functionality at two arbitrary frequencies. Conventional distributed microwave components typically exhibit a periodic response and hence they exhibit the required functionality at the design frequency and its odd harmonics. However, such components cannot be considered multi-band components since the operating frequencies cannot be set to those values corresponding to system requirements (namely, arbitrariness is missing). However, the greater number of parameters of metamaterial transmission lines (as compared to conventional lines) is useful for the design of dual-band components through dispersion engineering. To achieve dual-band operation, it is necessary to satisfy the phase and impedance requirements (for the transmission lines and stubs of the circuit) at the two design frequencies, \( f_1 \) and \( f_2 \). Let us consider that the required phases are \( \phi_1 \) and \( \phi_2 \) at the design frequencies. As can be seen in Fig. 13, in general, such phases cannot simultaneously be satisfied. However, by using artificial lines, we can tailor the dispersion diagram to set the phases to the required values at the two operating frequencies (such as Fig. 13 illustrates). One important aspect is that, contrary to conventional lines, metamaterial transmission lines exhibit frequency dependent characteristic impedance. Thus, such artificial lines must be designed to satisfy also the impedance requirements at the design frequencies.

For the implementation of dual-band metamaterial transmission lines, one approach is to cascade multiple effectively homogeneous unit cells. For instance, a balanced line in the vicinity of the transition frequency exhibits roughly constant (i.e., frequency independent) characteristic impedance. Thus, by setting this impedance to the required value and adjusting the slope of the dispersion diagram so that the phases of the unit cells are \( \phi_1/N \) and \( \phi_2/N \) at the design frequencies, dual-band operation can be achieved. Practical dual band components have been indeed implemented by cascading conventional lines (with positive phase shift) and backward lines implemented by means of lumped elements. As long as the phase shift per unit cell in the backward lines is small, the phase can be approximated by:

\[
\phi_C = \phi_R - \frac{N}{\omega \sqrt{L_C C_L}}
\]

where the sub-index \( C \) denotes the phase of the composite structure and \( N \) is the number of unit cells of the backward wave line. Following this strategy, several dual band components have been implemented (see for instance [37–40]).

Another approach consists on implementing such dual-band artificial lines by means of a single unit cell, designed to satisfy the impedance and phase requirements at the operating frequencies. In general this approach cannot be applied, but it is useful, for instance, for the design of dual-band 90° transmission lines (of application in many circuits as impedance or admittance inverters). The key idea behind this approach is to design the line so that the lower frequency lies within the left handed band, and the upper frequency within the right handed band. According
to this, we must force the electrical length and characteristic impedance to \( \phi = \phi_1 = -90^\circ \) and \( Z_B = Z_1 \), at the lower frequency, and \( \phi = \phi_2 = +90^\circ \) and \( Z_B = Z_2 \) at the upper frequency (notice that in principle we consider that the characteristic impedance can be different at each frequency). Let us consider the implementation of these dual-band impedance inverters by means of CSRR-loaded CRLH transmission lines. The lumped element model of the unit cell is described by means of 5 independent parameters, whereas the phase and impedance pairs represent 4 conditions. Thus, these parameters cannot be univocally determined unless an additional condition is imposed. If we force the structure to be balanced, that is, to exhibit a continuous transition between the left handed and right handed bands, the series and shunt resonance must be identical [8, 9]. Under these conditions the electrical parameters of the circuit model can be isolated. The following results are obtained:

\[
L = \frac{2(Z_1\omega_1 + Z_2\omega_2)}{\omega_2^2 - \omega_1^2}, \quad (57)
\]

\[
C_g = \frac{\omega_1^2 - \omega_2^2}{2\omega_1\omega_2(Z_1\omega_1 + Z_2\omega_2)}, \quad (58)
\]

\[
C = \frac{Z_1\omega_2 + Z_2\omega_1}{(Z_2^2 - Z_1^2)\omega_1\omega_2}, \quad (59)
\]

\[
C_c = \frac{Z_1\omega_2 + Z_2\omega_1}{(\omega_2^2 - \omega_1^2)Z_1Z_2}, \quad (60)
\]

\[
L_c = \frac{(Z_1\omega_1 + Z_2\omega_2)(\omega_2^2 - \omega_1^2)Z_1Z_2}{\omega_1\omega_2(Z_1\omega_2 + Z_2\omega_1)^2}, \quad (61)
\]

with \( \omega_1 = 2\pi f_1 \) and \( \omega_2 = 2\pi f_2 \). Inspection of Eqs. (57)–(61) reveals that \( L_c \), \( C_g \), \( C_c \), and \( L_c \) are always positive. However, \( C \) may be negative if \( Z_1 > Z_2 \) or, even, if \( C = \infty \) for a dual-band impedance inverter with identical impedances \( Z_1 = Z_2 \). If \( Z_1 > Z_2 \) (\( C < 0 \)), the inverter cannot be synthesized by means of a balanced cell described by the model of Fig. 10. If \( Z_1 = Z_2 \), as most applications require, the inverter can be synthesized by balancing the line, but substituting the capacitance \( C \) with an electrical short. In practice, this is not possible by using the cell topology shown in Fig. 2b, although this is possible through the CL-loaded approach. Hence, in order to implement dual-band quarter wavelength impedance inverters with identical impedances (at the frequencies of interest) by means of CRLH CSRR-based cells, such cells must be designed without the restriction of being balanced. Nevertheless, the circuit parameters of the series branch, \( L \) and \( C_g \) are still univocally determined under the unbalance condition (namely, Eqs. (57) and (58) hold). The reason is simple: since the \( \pm 90^\circ \) phases lead to \( Z_e = -Z_p \) at both \( \omega_1 \) and \( \omega_2 \) (see Eq. (4)), it follows from (3) that \( Z_1 = jZ_p(\omega_1) \) and \( Z_2 = -jZ_p(\omega_2) \), and Eqs. (57) and (58) result. The conditions involving the shunt branch are simply \( Z_1 = -jZ_p(\omega_1) \) and \( Z_2 = +jZ_p(\omega_2) \). Thus, \( C \), \( L_c \) and \( C_c \) are not univocally determined unless the structure is balanced. In general, it is convenient to design the dual-band impedance inverters by means of un-balanced unit cells. This provides more flexibility to the electrical parameters and it eases the implementation of the structure through the layout of Fig. 2b. Furthermore, as has been mentioned, if \( Z_1 = Z_2 \) (as usual), the structure must necessarily be un-balanced.

3.2. Some illustrative applications of dispersion engineering

In this section, several illustrative applications of dispersion engineering, based on resonant-type metamaterial transmission lines, are reported.

3.2.1. Examples of enhanced bandwidth components

Let us start with the design of some microwave components with enhanced bandwidth. At this point we must mention that the main aim is to enhance bandwidth in terms of the phase response. In other words, the aim has been to implement devices based on phase differences, and tailor the dispersion diagram to achieve quasi-parallel phase slopes of the different transmission lines at the design frequency, as has been discussed before. Obviously, since the characteristic impedance varies with frequency, it is not clear a priori that the bandwidth in terms of matching or transmission is superior, as compared to conventional implementations. Let us start by considering the design of a rat-race hybrid coupler [41, 42]. A typical layout of the conventional implementation is depicted in Fig. 14. It is essentially a 4-port device consisting on a 1.5λ ring structure (where \( \lambda \) is the guided wavelength at the design frequency, \( f_0 \)) with the ports equally spaced in the upper half of the ring. Since the 270° (0.75λ) line section is formally equivalent to a −90° LH line, the conventional 0.75λ line can be substituted by a LH artificial line designed to provide the required characteristic impedance (70.7 Ω) and phase (−90°) at the operating frequency. This was done previously by Okabe et al. [43] by using the dual transmission line approach, where the host line was loaded with lumped inductors and

![Typical layout of a rat-race hybrid coupler.](image-url)
capacitors in shunt and series connection, respectively. In this paper, we will consider the resonant-type version of the rat-race, where not only the 270° line is replaced with an artificial −90° LH line, but also the three 90° transmission lines are implemented as artificial right handed lines based on CSRRs. This allows us to achieve a further controllability on the frequency dependence of phase in the lines, with the result of an excellent performance in terms of phase-balance. Moreover, the design is fully compatible with planar technology since no lumped elements are used.

Concerning the implementation of left handed lines with CSRRs (actually CRLH lines), this has been considered before. However, we must discuss how to implement the right handed lines by means of CSRRs. This can be done through the topology shown in Fig. 15. As compared to the unit cell of the left handed line, in the right handed unit cell the gap is substituted with a grounded stub, acting as a shunt inductor. Since the series reactance is no longer negative, left handed wave propagation is not possible in these CSRR based lines. The typical dispersion diagram of these structures is shown in Fig. 16. Two right handed bands separated by a frequency gap do appear. In order to implement electrically small unit cells, operation in the first band is a due. For this reason, we do not consider the forward wave band of a CRLH CSRR-based unit cell for the implementation of a right handed transmission line. As mentioned before, the advantage of using artificial lines for these lines is the possibility to control the dispersion diagram so that roughly identical phase slopes between the right handed (+90°) lines and left handed (−90°) lines can be obtained.

The topology of the designed rat-race is depicted in Fig. 17, and the photograph of the fabricated prototype is depicted in Fig. 18, where it is compared to the conventional implementation. The artificial lines have been all designed to exhibit a characteristic impedance of $Z_0 = 70.71 \, \Omega$ at the design frequency (1.6GHz). The determination of the layout of each line has been done following the procedure described before [36, 44]. The device has been fabricated on the Rogers RO3010 substrate with dielectric constant $\varepsilon_r = 10.2$ and thickness $h = 635 \mu m$. The active area (excluding access lines) of the new hybrid coupler based on LH and RH artificial lines is 3.62 cm², whereas the conventional one occupies an area of 10.33 cm², i.e., the meta-coupler is roughly 3 times smaller. The simulated and measured impedance matching, coupling and isolation for both couplers are depicted in Fig. 19. In Fig. 20, it is depicted the phase-balance for the $\Sigma$ and $\Delta$ ports, namely $\phi(S_{42}) - \phi(S_{32})$ and $\phi(S_{41}) - \phi(S_{31})$, respectively. The CSRR-based coupler exhibits good isolation, coupling and matching. These magnitudes are comparable to those of the conventional device. Specifically, the measured power splitting between ports 3 and 4 (considering port 1 as the input port) exhibits similar characteristics in terms of flat-
The slight discrepancies between simulation and measurement in the conventional coupler are attributed to fabrication related tolerances. Reprinted with permission from [41]. Copyright 2007 IEEE.

Another example corresponds to the implementation of a power divider with quadrature output signals [45]. This has been implemented by means of a Y-junction with output lines exhibiting a phase difference of 90° at the design frequency. The metamaterial based device is depicted in Fig. 21, where it is compared to a conventional implementation. +90° and +180° output lines are used in the conventional implementation, whereas zero-degree and −90° lines are considered in the metamaterial-based device. The characteristic impedance of these output lines is 50 Ω, whereas the 90° admittance inverter has a characteristic impedance of 35.35 Ω to preserve input matching. The metamaterial based device exhibits better performance in terms of phase

Figure 19 (online color at: www.lpr-journal.org) Impedance matching ($S_{11}$), coupling ($S_{31}, S_{41}$) and isolation ($S_{21}$) for the CSRR-based hybrid coupler (a) and conventional coupler (b). The slight discrepancies between simulation and measurement in the conventional coupler are attributed to fabrication related tolerances. Reprinted with permission from [41]. Copyright 2007 IEEE.

Figure 20 (online color at: www.lpr-journal.org) Phase balance for the $\Delta$ (a) and $\Sigma$ (b) ports of the fabricated rat-race hybrids. Reprinted with permission from [41]. Copyright 2007 IEEE.

Figure 21 Comparative layouts (drawn to scale) for the CSRR based (a) and conventional (b) quadrature phase shifters. The active area of these devices is indicated by means of dashed rectangles. For the −90° left handed line, dimensions are: $c = 0.44$ mm, $d = 0.44$ mm, $r_{\text{ext}} = 5.10$ mm, $g = 0.17$ mm, and $w = 3.12$ mm. For the zero-degree balanced line dimensions are: $c = 0.51$ mm, $d = 0.51$ mm, $r_{\text{ext}} = 5.98$ mm, $g = 0.16$ mm, and $w = 8.12$ mm. The devices have been implemented on the Rogers RO3010 substrate with dielectric constant $\varepsilon_r = 10.2$, thickness $h = 635 \mu$m. Reprinted with permission from [45].
shift since the operative bandwidth is superior, as can be seen in Fig. 22 (the operative bandwidth is determined by that frequency region where deviation of the phase shift from the nominal value is smaller than a certain tolerance). Fig. 23 depicts the power splitting between ports 2 and 3 for both the conventional and the metamaterial quadrature phase shifter. Power splitting in the metamaterial phase shifter exhibits a soft dependence on frequency in the vicinity of the design frequency, although the conventional device exhibits a broader bandwidth since the characteristic impedance does not depend on frequency. Nevertheless, the relevant parameter in a quadrature phase shifter is the phase balance and it is clearly superior in the proposed metamaterial-based device. Concerning dimensions, it must be pointed out that the active region of the metamaterial phase shifter is smaller than that of the conventional implementation (both are indicated in Fig. 21).

In the previous examples, the metamaterial-based components have been implemented through the resonant type approach. However, some other works have shown the possibilities of bandwidth enhancement by using CL-loaded lines (see for instance the design of broadband series power dividers [34], phase shifters [46, 47] and Wilkinson baluns [48]).

3.2.2. Dual-band components

Dual-band impedance inverters based on CSRRs have been applied to the design of branch line hybrid couplers and power dividers operative at the GSM bands (0.9GHz and 1.8GHz), as illustrative examples of the application of CSRR-based metamaterials to dual band components [49]. The frequency ratio is 2, this being a good choice to demonstrate the possibilities of the approach. The branch line hybrid coupler is depicted in Fig. 24, where it is also shown the power splitting, matching and isolation, as well as the phase response of the device. Dual band operation at the design frequency is clearly achieved. Apart from dual band operation, the device is also small on account of the small electrical size of the impedance inverters forming the branch line coupler.

The dual band power divider has been implemented by using complementary spiral resonators (CSRs) [50]. As compared to CSRRs, the electrical size of CSRs is roughly one half [9, 51]. This allows further size reduction. The dual band device is depicted in Fig. 25, where it is also depicted the power splitting and matching. Dual band operation is confirmed. Other dual-band power dividers based on metamaterial resonators are reported in [50] and [52].

Dual band devices based on the CL-loaded approach have also been reported in various works [37–40, 53, 54].

4. Conclusions

In conclusion, dispersion engineering with metamaterial transmission lines has been reviewed. The main relevant characteristics of metamaterial transmission lines, including both the CL-loaded lines and the resonant type lines have been pointed out. It has been shown that such artificial lines exhibit a composite right/left handed behavior, which
has been confirmed through measurement, and corroborated from the analysis of the equivalent circuit model of the different lines. The principles for bandwidth enhancement and dual-band operation have been highlighted, and several prototype devices illustrative of both applications of dispersion engineering have been reported. Such reported examples correspond to microwave components based on the resonant type approach of metamaterial transmission lines, since this is the approach pioneered by the authors (although other reported devices based on the CL-loaded approach have been pointed out, as has been properly quoted).

Small size and compatibility with planar technology are additional benefits of the use of metamaterial transmission lines. To end this paper, we would like to mention that metamaterial transmission lines have been also applied to the design of many other microwave components where miniaturization and/or performance improvement has been achieved (filters, diplexers, coupled line couplers, phase shifting lines, etc.). Also, due to the unusual propagation properties of metamaterial transmission lines, such lines have been applied to the design of leaky-wave antennas (for further information on these topics, the authors recommend the books [6–9] and references therein).

Acknowledgements: This work has been supported by MEC by project contract TEC2007-68013-C02-02 METAINNOVA and by a FPU Grant (Ref. AP2005-4523) 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.
Gerard Sisó was born in Barcelona in 1981. He received the Industrial Engineering Diploma, specialising in Electronics from the Universitat Politécnica de Catalunya in 2004 and the Electronics Engineering degree from the Universitat Autònoma de Barcelona in 2006. He is now working toward his PhD degree. His research interests include microwave circuits based on metamaterial transmission lines, specially enhanced bandwidth components and dual-band devices.

Marta Gil Barba was born in Valdepeñas (Ciudad Real), Spain, in 1981. She received the degree in Physics from the Universidad de Granada in 2005. She is currently working toward her PhD degree in subjects related to metamaterials and microwave circuits. She is presently holder of a national research fellowship from the FPU Programme of the Education and Science Spanish Ministry (MEC).

Francisco Aznar Ballesta was born in 1978 in Granada (Spain). He received the degree in Electronics Engineering from the Universidad de Granada in 2005. He studied during two years in Darmstadt, Germany, doing the project in FG Mikroelektronische Systeme in the Technische Universität Darmstadt. He is currently working toward his PhD and is also Assistant Professor at the Universitat Autònoma de Barcelona.

Jordi Bonache was born in Cardona (Barcelona), Spain, in 1976. He received the Physics and Electronics Engineering degrees from the Universitat Autònoma de Barcelona, Bellaterra (Barcelona), Spain, in 1999 and 2001, respectively and the Ph.D. degree in Electronics Engineering in 2007 from the same University. In 2000, he joined the High Energy Physics Institute of Barcelona (IFAE), where he was involved in the design and implementation of the control and monitoring system of the MAGIC telescope. In 2001, he joined the department d’Enginyeria Electrònica, Universitat Autònoma de Barcelona, where he is currently an Assistant Professor. His research interests include active and passive microwave devices and metamaterials.

Ferran Martín was born in Barakaldo (Vizcaya), Spain in 1965. He received the B.S. Degree in Physics from the Universitat Autònoma de Barcelona (UAB) in 1988 and the PhD degree in 1992. From 1994 up to 2006 he has been Associate Professor in Electronics at the Departament d’Enginyeria Electrònica (Universitat Autònoma de Barcelona), and from 2007 he is Full Professor of Electronics. In recent years, he has been involved in different research activities including modelling and simulation of electron devices for high frequency applications, millimeter wave and THz generation systems, and the application of electromagnetic bandgaps to microwave and millimeter wave circuits. He is now very active in the field of metamaterials and their application to the miniaturization and optimization of microwave circuits and antennas. He is the head of the Microwave and Millimeter Wave Engineering Group (GEMMA Group) at UAB, and a partner of the Network of Excellence of the European Union METAMORPHOSE. He has authored and co-authored over 250 technical conference, letter and journal papers and he is co-author of the monograph on Metamaterials entitled Metamaterials with Negative Parameters: Theory, Design and Microwave Applications (John Wiley & Sons Inc.). He has recently launched CIMITEC, a research Center on Metamaterials (supported by CIDEM - Catalan Government) where he acts as Director. Dr. Ferran Martín and CIMITEC have received the 2006 Duran Farel Prize for Technological Research.

References


