

Metamaterials Handbook

Applications of Metamaterials

Edited by
Filippo Capolino



CRC Press
Taylor & Francis Group
Boca Raton London New York

CRC Press is an imprint of the
Taylor & Francis Group, an **informa** business

MATLAB® is a trademark of The MathWorks, Inc. and is used with permission. The MathWorks does not warrant the accuracy of the text or exercises in this book. This book's use or discussion of MATLAB® software or related products does not constitute endorsement or sponsorship by The MathWorks of a particular pedagogical approach or particular use of the MATLAB® software.

CRC Press
Taylor & Francis Group
6000 Broken Sound Parkway NW, Suite 300
Boca Raton, FL 33487-2742

© 2009 by Taylor and Francis Group, LLC
CRC Press is an imprint of Taylor & Francis Group, an Informa business

No claim to original U.S. Government works

Printed in the United States of America on acid-free paper
10 9 8 7 6 5 4 3 2 1

International Standard Book Number: 978-1-4200-5423-1 (Hardback)

This book contains information obtained from authentic and highly regarded sources. Reasonable efforts have been made to publish reliable data and information, but the author and publisher cannot assume responsibility for the validity of all materials or the consequences of their use. The authors and publishers have attempted to trace the copyright holders of all material reproduced in this publication and apologize to copyright holders if permission to publish in this form has not been obtained. If any copyright material has not been acknowledged please write and let us know so we may rectify in any future reprint.

Except as permitted under U.S. Copyright Law, no part of this book may be reprinted, reproduced, transmitted, or utilized in any form by any electronic, mechanical, or other means, now known or hereafter invented, including photocopying, microfilming, and recording, or in any information storage or retrieval system, without written permission from the publishers.

For permission to photocopy or use material electronically from this work, please access www.copyright.com (<http://www.copyright.com/>) or contact the Copyright Clearance Center, Inc. (CCC), 222 Rosewood Drive, Danvers, MA 01923, 978-750-8400. CCC is a not-for-profit organization that provides licenses and registration for a variety of users. For organizations that have been granted a photocopy license by the CCC, a separate system of payment has been arranged.

Trademark Notice: Product or corporate names may be trademarks or registered trademarks, and are used only for identification and explanation without intent to infringe.

Library of Congress Cataloging-in-Publication Data

Applications of metamaterials / editor, Filippo Capolino.

p. cm.

"A CRC title."

Includes bibliographical references and index.

ISBN 978-1-4200-5423-1 (alk. paper)

1. Metamaterials. 2. Electronic apparatus and appliances--Materials. I. Capolino, Filippo. II. Title.

TK7871.15.M48A67 2009

621.381--dc22

2008054554

Visit the Taylor & Francis Web site at
<http://www.taylorandfrancis.com>

and the CRC Press Web site at
<http://www.crcpress.com>

Contents

Foreword	xi
Preface	xiii
Editor	xv
Advisory Board	xvii
Contributors	xix

Part I Super-Resolution

1	Negative-Refractive-Index Transmission-Line (NRI-TL) Metamaterial Lenses and Superlenses <i>Ashwin K. Iyer and George V. Eleftheriades</i>	1-1
2	Flat Lenses Formed by Photonic and Electromagnetic Crystals <i>Pavel A. Belov, Constantin R. Simovski, and Pekka Ikonen</i>	2-1
3	Subwavelength Imaging by Arrays of Metallic Rods <i>Pavel A. Belov, Mário G. Silveirinha, Constantin R. Simovski, and Yang Hao</i>	3-1
4	Super-Resolution and Near-Field Enhancement with Layers of Resonant Arrays of Nanoparticles <i>Sergiy Steshenko, Filippo Capolino, Sergei A. Tretyakov, and Constantin R. Simovski</i>	4-1
5	Super-Resolution Imaging with Hyperlens <i>Hyesog Lee, Yi Xiong, Zhaowei Liu, Cheng Sun, and Xiang Zhang</i>	5-1

Part II Cloaking

6	Cloaking and Transformation Media <i>Ulf Leonhardt and Thomas G. Philbin</i>	6-1
7	Scattering Cancellation and Plasmonic Cloaking <i>Andrea Alù and Nader Engheta</i>	7-1
8	Electromagnetic Cloaks and Concentrators <i>Arthur D. Yaghjian, Stefano Maci, and Enrica Martini</i>	8-1

Part III Circuit Applications in the Microwave, Millimeter Wave, and THz Frequency Ranges

- 9 Application of Split Ring Resonators to Microwave Circuit Design
Ferran Martín and Ricardo Marqués 9-1
- 10 Application of Electromagnetic Band Gaps to Microwave Circuit Design
Jordi Bonache, Francisco Falcone, Ignacio Gil, Joan García-García, and Ferran Martín 10-1
- 11 Small and Multiband MNG Resonators: Spiral, Prefractal, and Other Geometries
Pere J. Ferrer, José M. González-Arbesú, Jordi Romeu, Josep Parrón, Gary Junkin, and Ramón Villarino 11-1
- 12 Artificial Ferromagnetic Nanostructured Substrates for Planar Tunable Circuits
J. Spiegel, I. Huynen, L. Piraux, and A. Saib 12-1
- 13 Microwave Phase Shifters and Filters Based on a Combination of Left-Handed and Right-Handed Transmission Lines
I. B. Vendik, D. V. Kholodnyak, and P. V. Kapitanova 13-1
- 14 Magnetoinductive Waves II: Applications
O. Sydoruk, A. Radkovskaya, O. Zhuromskyy, E. Shamonina, and L. Solymar . . 14-1

Part IV Radiation Applications in the Microwave, Millimeter Wave, and THz Frequency Ranges

- 15 CRLH Metamaterial Antennas, Part I: Theory and Antenna-Related Concepts
Christophe Caloz and Tatsuo Itoh 15-1
- 16 CRLH Metamaterial Antennas, Part II: Leaky-Wave and Resonant Antennas
Christophe Caloz and Tatsuo Itoh 16-1
- 17 Recent Developments of Metamaterial-Based and Metamaterial-Inspired Efficient, Electrically Small Antennas
Richard W. Ziolkowski and Aycan Erentok 17-1
- 18 Application of Metamaterials to Microwave Patch and Leaky-Wave Antennas
Lucio Vegni, Filiberto Bilotti, Andrea Alù, and Nader Engheta 18-1
- 19 Enhancement of Directivity by Using Metamaterial Substrates
Paolo Burghignoli, Giampiero Lovat, Filippo Capolino, David R. Jackson, and Donald R. Wilton 19-1

20	Metamaterial Antenna Applications <i>J. (Yiannis) C. Vardaxoglou, Rob D. Seager, James A. Flint, Alford Chauraya, George K. Palikaras, and Charity B. Mulenga</i>	20-1
21	High-Impedance Surfaces: Applications <i>Alexandros P. Feresidis, George Goussetis, Alexander B. Yakovlev, and Constantin R. Simovski</i>	21-1
22	Controllable Metamaterials for Telecoms: Principles, Designs, and Applications—Applications in the GSM, GPRS, and UMTS Bands <i>Frédérique Gadot and André de Lustrac</i>	22-1
23	Optoelectronic Control of Metamaterials <i>Willie J. Padilla and Richard D. Averitt</i>	23-1
24	Extraordinary Transmission as an Impedance-Matching Problem <i>Francisco Medina, Francisco Mesa, Ricardo Marqués, and Diana C. Skigin</i>	24-1

Part V Applications in the Optical Frequency Range

25	Plasmonic Materials for Near-Field Optical Nanolithography <i>Richard J. Blaikie</i>	25-1
26	Enhanced Transmission at Optical Frequencies <i>L. Martin-Moreno and F. J. Garcia-Vidal</i>	26-1
27	Photonic Crystals for Integrated Communication Systems <i>Henri Benisty and Jean-Michel Lourtioz</i>	27-1
28	Photonic Applications of Two-Dimensional Quasicrystals <i>Dmitry N. Chigrin and Andrei V. Lavrinenko</i>	28-1

Part VI Fabrication Techniques for THz and Optical Metamaterials

29	Fabrication and Optical Characterization of Photonic Metamaterials <i>Stefan Linden and Martin Wegener</i>	29-1
30	Metamaterials at Optical Frequencies: Fabrication and Measurements <i>Nigel P. Johnson, Richard M. De La Rue, and Sarah A. De La Rue</i>	30-1
31	Self-Organized Structures for Metamaterials <i>Dorota A. Pawlak</i>	31-1
32	Self-Assembly and Nanochemistry Techniques for the Fabrication of Metamaterials <i>Virginie Ponsinet, Ashod Aradian, Philippe Barois, and Serge Ravaine</i>	32-1

Index		I-1
-------	--	-----

Artificial Ferromagnetic Nanostructured Substrates for Planar Tunable Circuits

12.1	Introduction	12-1
12.2	Nanowired Composite	12-2
	Topology of Nanowired Substrate • Dielectric Properties • Magnetic Properties • Losses • Tunability	
12.3	Feasibility of Tunable Single- and Double-Negative Planar Metamaterials Based on Nanowired Substrate	12-10
	Single-Negative Material • Double-Negative Material • Tunability	
12.4	Transmission Line Modeling of Stopband Effects	12-14
	Electromagnetic Response of EBG Dielectric Substrate • Tunable Electromagnetic Response of Nanowired Substrate	
12.5	Demonstration of Planar MEBG Structures	12-19
	Topology • Design of the MEBG • Tunable Electromagnetic Response of MEBG	
12.6	Defect Modes	12-24
	Defect Modes in a Dielectric EBG Material • Defect Modes in MEBG	
12.7	Potential Applications in Magneto-Optics	12-30
12.8	Conclusions	12-32
	References	12-32

J. Spiegel
Université Catholique de Louvain

I. Huynen
Université Catholique de Louvain

L. Piraux
Université Catholique de Louvain

A. Saib
Cissoïd SA

12.1 Introduction

Periodic materials and devices have been extensively studied for more than 40 years across the world. Artificial composite materials that might theoretically exhibit low or negative permittivity and permeability [1,2] and photonic or periodic bandgap (PBG) structures, more generally designed as electromagnetic bandgap (EBG) structures, where the periodicity induces some frequency bands and where the propagation of waves is stopped [3–5], are studied for this purpose. The first approach enables the design of special substrates and lenses [6] for antenna design and the second approach, including the concept of defects, is related to the design of planar filters and sensors.

The artificial composite proposed here is based on ferromagnetic nanowired materials, which are of special interest in communication systems, because higher operating frequencies can be obtained compared to classical magnetic materials used for this purpose up to now (e.g., ferrites).

It is composed of an array of randomly distributed ferromagnetic nanowires of nanometric diameter embedded in a polymer or alumina membrane. The small wire diameter, obtained by using advanced technologies developed in-house, allows an easy penetration of electromagnetic waves inside the structure and enables the insulating nature of the membrane. Moreover, due to the high aspect ratio of nanowires (>200), desired magnetic properties for microwave devices are obtained without a biasing static magnetic field. This leads to a considerable potential increase of compactness and ease of integration in microwave monolithic integrated circuits (MMICs).

PBG materials and devices were first developed in the optical range (concept of photonic crystals), but were rapidly scaled down to the microwave and millimeter waves frequencies [3,7–10]. During many years, the existence of the bandgap was associated to the periodic change in refractive index along the material under scope. Later, Kee et al. [11,12], and Tiwari et al. [13] associated for the first time the existence of the stopband in PBG materials and devices to a periodic change of the wave impedance induced by a change of the dielectric or magnetic properties. The essential role played by the periodic change in wave impedance is demonstrated experimentally by Kee et al. [12], and Saib et al. [14]. Strictly speaking, these experimental proofs are based on a variation of the guided mode characteristic impedance, since planar microstrip transmission lines (TLs) lying on the PBG substrate are used for these experiments.

In this chapter, we present ferromagnetic composite materials that exhibit tunable or negative permeability [1,15], EBG or PBG structures, and the combination of the two approaches resulting in a Magnetic Electromagnetic BandGap (MEBG) structure proven at microwave frequencies, which could be extended into the optical range. For EBG applications, a transmission line approach (TLA) has been developed, complementary to the theory of reflections developed in the optical range for the modeling of Bragg reflectors. This approach enables first the prediction and analysis of the bandgap properties of the devices under scope, both in transmission and reflection. Devices considered in this chapter indeed consist of planar TLs, lying on substrates similar to integrated Bragg-reflector topologies in optics, since they exhibit periodicity in terms of dielectric constant or magnetic permeability. Due to the presence of guiding structures on the substrate, propagation into the so-called PBG substrate is really governed by TL mechanisms. Also, we consider in this chapter the presence of defect modes, which induce transmission of signal inside the bandgap. It is known [12,16–18] that in EBG structures, when the periodicity is broken locally inside the array, very sharp transmission peaks are created in the bandgap, called defect modes. Defect-mode operation at optical wavelengths can be viewed as a resonance occurring in a cavity formed by a Fabry–Perrot resonator between two perfect Bragg reflectors.

Finally, we discuss the potential applications in magneto-optics (MOs). MO effects [19] are indeed primarily used for polarization control, based on the use of the Kerr or Faraday effect [20]. MO materials are of prime interest for integrated photonics: review papers [21,22] mention the Faraday rotation effect for designing isolators or polarizers and the realization of magnetophotonic crystals [23].

12.2 Nanowired Composite

12.2.1 Topology of Nanowired Substrate

The devices mentioned above, besides the EBG structures, are based on the so-called nanowired substrate, shown in Figure 12.1. It consists of a porous insulating polycarbonate (PC) or alumina membrane filled with metallic nanowires, ferromagnetic or not magnetic, by electrodeposition [24]. The nanowire materials are nickel (Ni), cobalt (Co), iron (Fe), their alloys, and the nonmagnetic copper (Cu). The PC membranes have a porosity from 2% to 10%, a height of 21 μm , and cylindrical pores of diameters from 30 to 500 nm. The alumina membranes have a porosity of up to 40%, a height of up to 10 μm , and cylindrical pores down to 20 nm. The pores are randomly distributed in the membrane because of their fabrication process. The characterization of the nanowired substrate

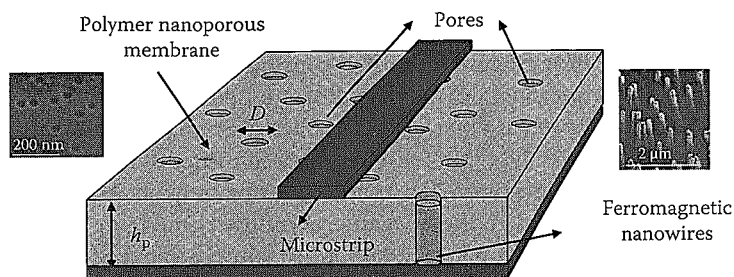


FIGURE 12.1 MSL on ferromagnetic nanowired substrate. (From Spiegel, J., de la Torre, J., Darques, M., Piriaux, L., and Huynen, I., *IEEE Microwave Wireless Component Lett.*, 7, 492, 2007. With permission.)

is important for the development of microwave devices and is done here by a microstrip approach. It permits to determine the dielectric properties, magnetic properties, and dielectric losses. For this purpose, a permittivity [25] and a permeability model were developed and validated by S-parameter measurements of microstrip lines (MSLs).

12.2.2 Dielectric Properties

Taking the metallic nature of our ferromagnetic nanowires together with recent developments on metamaterials based on arrays of metallic wires into account [26], we were often asked by the scientific community to clarify the exact influence of ferromagnetic nanowires on the reflective behavior and the permittivity of the resulting artificial composite material. Figure 12.2a shows the skin depth δ versus frequency f for two typical metallic materials considered in this section: nonmagnetic Cu and ferromagnetic Co. Up to 40 GHz, $\delta_{\text{Cu}} > 300$ nm and $\delta_{\text{Co}} > 750$ nm. Hence, as shown by Figure 12.2b, nanowires with diameter $D < \delta$ permit an easy penetration of the electromagnetic waves from surface ($x=r$) to center ($x=0$): for $D=50$ nm, no decrease of the magnetic field is observed over depth of the wires, while more than 80% of its magnitude is preserved for $D=500$ nm. This explains why we

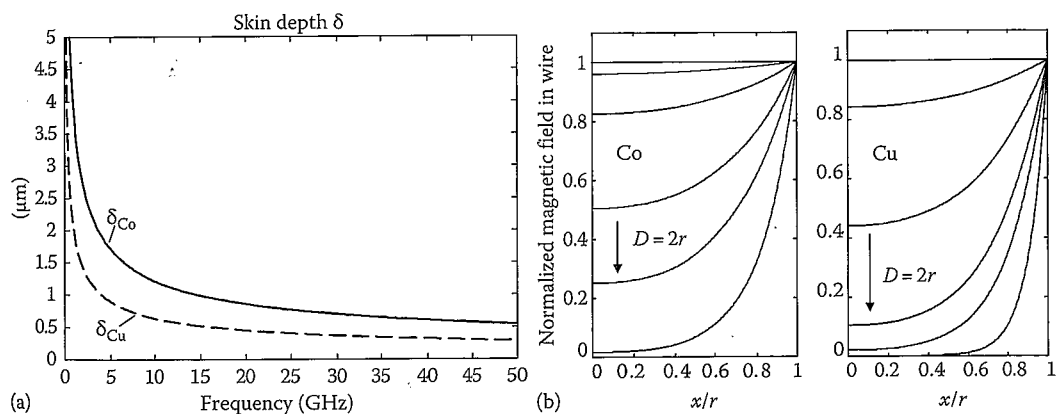


FIGURE 12.2 (a) Skin depth for Co and Cu material, calculated via expression $\delta = \sqrt{1/\pi f \mu_0 \sigma_w}$ and (b) calculated radial dependence of normalized magnetic field magnitude inside Co and Cu wire from center ($x=0$) to surface ($x=r$), for various wire diameters $D=2r$: 50 nm, 500 nm, 1 μm , 1.75 μm , 2.5 μm , and 5 μm . Calculation is made at 40 GHz, arrows indicate increase of wire diameter from top to bottom curves. (From Spiegel, J., de la Torre, J., Darques, M., Piriaux, L., and Huynen, I., *IEEE Microwave Wireless Component Lett.*, 7, 492, 2007. With permission.)

can observe a macroscopic ferromagnetic resonance (FMR) effect [24] in spite of the metallic nature of our ferromagnetic wires. From Figure 12.2b, it is concluded that unlike thick metallic conductors, nanowires with $D < 1 \mu\text{m}$ cannot act like shieldings and do not reflect electromagnetic waves. The substrate could also be viewed as an array of metallic wires embedded in a dielectric medium. For a wire diameter $D = 50 \text{ nm}$ and a typical distance between wires of $a = 0.3 \mu\text{m}$, we can estimate using Pendry formalism ([26], Equation 12.14) that the plasma frequency should be $\omega_p = 240 \text{ THz}$. This means that the substrate behaves like a negative permittivity medium below ω_p and becomes positive above. However, this formalism is only valid when wires can be considered as infinitely long compared to the wavelength and the array is perfectly periodic. These two conditions are not fulfilled, since the distribution of wires inside the membrane is random and its thickness is only a few micrometers. Hence, we make the assumption that the nanowired substrate of Figure 12.1 behaves like a classical medium in the microwave range and can be modeled by a simple equivalent circuit where nanowires act like resistors. The inductances introduced by the current flowing into nanowires are negligible, because of the low porosity of the substrate, which is around 2%. No magneto-dielectric effect occurs [27,28].

Our theoretical permittivity model is based on a MSL approach and takes geometrical (porosity and height of membrane, height of nanowires) and material parameters (choice of nanowired material) of the substrate into account. Figure 12.3a shows the 2D profile of an MSL with ferromagnetic nanowires in the supporting membrane. Figure 12.3b shows the corresponding equivalent circuit for the line capacitance per unit length. The partial capacitances of the system are calculated from parallel connections as follows:

$$C_{pa} = C_a + C_{p1} \quad (12.1)$$

$$C_{pw} = C_w + C_{p2}. \quad (12.2)$$

Thus, the total capacitance C_{tot} results from a series connection of C_{pa} and C_{pw} :

$$C_{\text{tot}} = \frac{C_{pa} C_{pw}}{C_{pa} + C_{pw}}. \quad (12.3)$$

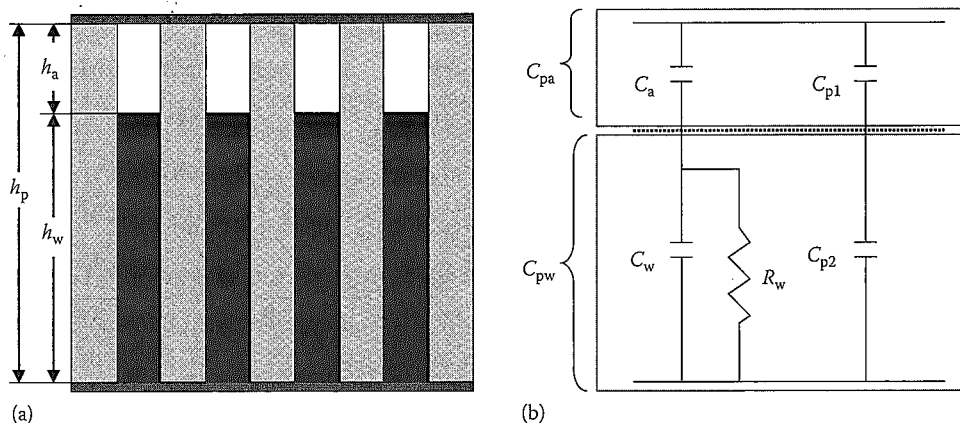


FIGURE 12.3 (a) 2D profile of a MSL including nanowires in the supporting substrate and (b) corresponding two-layer equivalent circuit of capacitance per unit length. (From Spiegel, J., de la Torre, J., Darques, M., Piraux, L., and Huynen, I., *IEEE Microwave Wireless Component Lett.*, 7, 492, 2007. With permission.)

The overall permittivity can then be calculated from the obtained capacitance C_{tot} :

$$\epsilon_{\text{eff}} = \frac{C_{\text{tot}} h_p}{\epsilon_0 W_s} = \frac{h_p}{\frac{h_w}{\epsilon_{\text{pw}}} + \frac{h_a}{\epsilon_{\text{pa}}}}, \quad (12.4)$$

where

W_s is the width of the microstrip

h_p is the membrane height

h_w is the height of the nanowires

h_a is the air gap, between the top of the nanowires and the bottom of the MSL as defined on

Figure 12.3a

and the permittivities are obtained from layer capacitances, namely for the upper layer,

$$\epsilon_{\text{pa}} = \frac{C_{\text{pa}} h_a}{\epsilon_0 W_s} = p + (1 - p) \epsilon_{\text{rb}} (1 - j \tan \delta_b) \quad (12.5)$$

and for the bottom layer,

$$\epsilon_{\text{pw}} = \frac{C_{\text{pw}} h_w}{\epsilon_0 W_s} = p(1 - j \frac{\sigma_w}{\omega \epsilon_0}) + (1 - p) \epsilon_{\text{rb}} \cdot (1 - j \cdot \tan \delta_b), \quad (12.6)$$

where

p is the porosity of the membrane defined as the fraction of total surface occupied by pore sections in a given surface of membrane

σ_w is the conductivity of the wires

$\tan \delta_b$ is the dielectric loss tangent factor of the bulk material

ϵ_{rb} is the dielectric constant of the bulk material

Simulations have been performed with this effective permittivity formula (Equation 12.4). The wire conductivity is $\sigma_w = 1.72 \cdot 10^7$ S/m (Co), the dielectric losses of the used membrane $\tan \delta_b = 0.1$, and its bulk permittivity $\epsilon_b = 2.89$. The chosen simulation frequency here is 40 GHz, but it was verified that a similar behavior is observed at lower frequencies. Figure 12.4a shows the result of ϵ_{eff} depending on membrane's porosity and on the height of nanowires as a percentage of the substrate height. The dependence on membrane's porosity is linear: the permittivity increases, when the porosity decreases. Furthermore the permittivity increases with increasing nanowire height. The dependence on the nanowire height for a given porosity ($p = 2\%$, Figure 12.4b) starts with the value of the empty porous membrane. Up to a nanowire height of 40% of the substrate height, the dependence is quite linear. For higher nanowires, the effective permittivity increases dramatically due to the more important impact of the capacitance of air C_a . This is due to the fact that the air gap between the surface of the nanowires and the bottom side of the microstrip decreases. So C_a is higher in a filled membrane than in an empty one. It is inversely proportional to its corresponding height h_a . The simulation result has been validated by S -parameter measurements on different sets of MSLs. This leads to the extraction of the product $\epsilon_{\text{eff}} \cdot \mu_{\text{eff}}$, effective permittivity $\times$ effective permeability. It has been obtained by using a TLA. S parameters of a TL indeed involve its complex propagation constant, which is proportional to $\sqrt{\epsilon_{\text{eff}} \cdot \mu_{\text{eff}}}$. Its extraction is made possible using the chain matrix formalism detailed in Section 12.4.

The measured samples have a height of the nanowires of 20%, 40%, and 55% of the substrate height [25]. A reference example without nanowires is also included. Figure 12.5a shows the real part of the product $\epsilon_{\text{eff}} \cdot \mu_{\text{eff}}$. A resonance is observed around 26 GHz and is due to the FMR occurring in

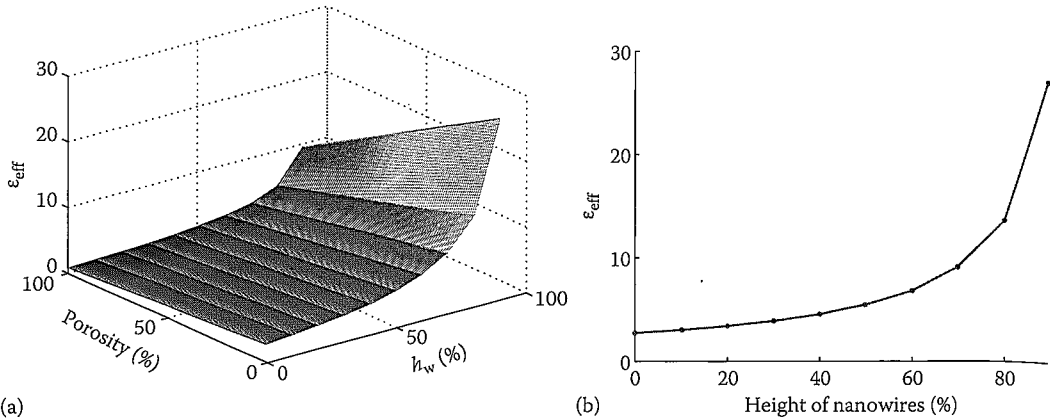


FIGURE 12.4 (a) 3D permittivity view depending on porosity of membrane and height of nanowires, and (b) 2D view of permittivity's real part for $p = 2\%$. (From Spiegel, J., de la Torre, J., Darques, M., Piriaux, L., and Huynen, I., *IEEE Microwave Wireless Component Lett.*, 7, 492, 2007. With permission.)

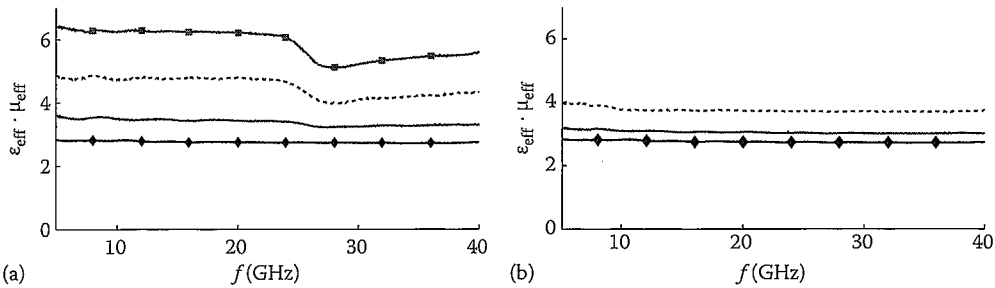


FIGURE 12.5 Extraction of real part of $\epsilon_{eff} \cdot \mu_{eff}$ from measurements of MSLs on composite substrates with three heights of (a) Co and (b) Cu nanowires in a porous membrane of porosity $p = 2\%$: empty sample (diamond), 20% sample (solid), 40% sample (dashed), and 55% sample (square). (From Spiegel, J., de la Torre, J., Darques, M., Piriaux, L., and Huynen, I., *IEEE Microwave Wireless Component Lett.*, 7, 492, 2007. With permission.)

Co material [24]: theory predicts a strong variation of the permeability's real part and an extremum on its imaginary part responsible for losses [29]. Supposing that $\mu_{eff} = 1$ in all frequency regions besides the resonance frequency, these results can be compared to the simulation. At 40 GHz, the predictions for all samples derived from Figure 12.4b are in good accordance with the measurements (Figure 12.5a), namely $\epsilon_{eff} = 2.76$ for 0% (empty membrane), 3.4 for 20%, 4.5 for 40%, and 6.1 for 55%, respectively. Moreover, similar results are observed when replacing Co by Cu wires in the same membrane (Figure 12.5b): again the measured permittivity increases with wire height, and is in good agreement with the simulation as well as with corresponding values for Co wires. Here the extraction leads directly to the effective permittivity since $\mu_{eff} = 1$ in absence of magnetic material.

12.2.3 Magnetic Properties

The nanowired substrate filled with ferromagnetic nanowires exhibits a wideband absorption phenomenon in the frequency range (6–30 GHz). It is due to the FMR, which is present even if there is no

external direct current (DC) biasing magnetic field. In this case, the frequency position of the FMR depends only on the ferromagnetic material filling the pores [30]. The effective permeability can be modeled by a variational approach. First of all, the propagation constant γ_{eff} has to be calculated [31]:

$$\gamma_{\text{eff}} \approx \gamma_p + \frac{j\omega \int \int \int_{\Delta V_m} \vec{E}^* \epsilon_0 (\epsilon_{\text{eff}} - \epsilon_{\text{rp}}) \vec{E} dV}{2\Delta L \int \int_{\Delta S} \vec{E} \cdot \vec{H} dS} + j\omega \frac{\int \int \int_{\Delta V_m} \vec{H}^* \epsilon_0 (\bar{\mu} - 1) \vec{H} dV}{2\Delta L \int \int_{\Delta S} \vec{E} \cdot \vec{H}^* dS}, \quad (12.7)$$

where

γ_p is the propagation constant of the porous polymer

ϵ_{rp} is the permittivity of the porous polymer

Both determined by the TL theory.

The permittivity of the composite ϵ_{eff} is extracted by the model given in Section 12.2.2. The permeability tensor of the composite $\bar{\mu}$ can then be deduced from the susceptibility tensor:

$$\bar{\mu} = \begin{bmatrix} \mu & -j\kappa & 0 \\ j\kappa & \mu & 0 \\ 0 & 0 & 1 \end{bmatrix} \quad (12.8)$$

with

$$\mu = 1 + p \cdot \mu_w \quad (12.9)$$

$$\kappa = p \cdot \kappa_w. \quad (12.10)$$

where p is the porosity of the membrane. The elements μ_w and κ_w of the wire tensor can be calculated by

$$\mu_w = 1 + \frac{\omega_m (\omega_e + j\omega\tilde{\alpha})}{(\omega_e + j\omega\tilde{\alpha})^2 - \omega^2} \quad (12.11)$$

$$\kappa_w = \frac{j\omega_m \omega}{(\omega_e + j\omega\tilde{\alpha})^2 - \omega^2} \quad (12.12)$$

with

$$\omega_m = \frac{\tilde{\gamma} M_s}{\mu_0} \quad (12.13)$$

$$\omega_e = \tilde{\gamma} H_{\text{eff}}, \quad (12.14)$$

where

$\tilde{\alpha}$ is the damping factor

$\tilde{\gamma}$ is the gyromagnetic factor

M_s is the saturation magnetization

H_{eff} is the effective magnetic field, given by

$$\tilde{\gamma} H_{\text{eff}} = \begin{cases} \tilde{\gamma} H_{\text{DC}} + \frac{1}{2} \omega_m (1 - 3p) & \text{if } H_{\text{DC}} \neq 0 \\ \frac{1}{2} \omega_m & \text{if } H_{\text{DC}} = 0. \end{cases} \quad (12.15)$$

The FMR occurs when the denominator in Equation 12.11 is equal to zero [32]. At remanent state, in the absence of external applied DC magnetic field H_{DC} , the FMR occurs for the so-called FMR angular frequency $\omega_r = \frac{1}{2} \omega_m$ [30]. The factor 1/2 holds for the demagnetizing factors of a single cylindrical nanowire. Its nonzero value explains why FMR can occur in absence of a DC biasing field. In Equation 12.7 the volume ΔV_m considers a small length of line ΔL , having as transverse section S , the product $W_s \cdot h_p$, with W_s width of the MSL and h_p membrane height. The effective permeability for the ferromagnetic nanowired substrate can finally be extracted by

$$\mu_{\text{eff}} = \frac{\gamma_{\text{eff}} \cdot c_0}{j\omega \sqrt{\epsilon_{\text{eff}}}}, \tag{12.16}$$

where
 c_0 is the speed of light in vacuum
 ϵ_{eff} is given by Equation 12.4

Simulations have been performed with this effective permeability formula. Figure 12.6a shows the real part and Figure 12.6b imaginary part of the effective permeability. The geometric parameters of the device are porosity of membrane $p = 2\%$, height of the polymer $h_p = 21 \mu\text{m}$, and width of the MSL $W_s = 150 \mu\text{m}$. The simulated MSLs contain Co nanowires with a height of 20%, 40%, and 55% of the substrate height. The strong variation of the permeability's real part and the extremum of its imaginary part at around 26 GHz describes the FMR occurring in Co material. The higher the nanowires, the larger is the variation of the real part. The imaginary part shows that the magnetic losses increase with the height of the nanowires. These simulation results have been validated by measurements on a set of MSLs with the same material parameters. The extraction of μ_{eff} has been done using the TLA. Figure 12.7a shows the real and Figure 12.7b the imaginary parts of μ_{eff} . The measurements are in good accordance with the simulation results. A variation in the real part of the permeability is visible as well as the extremum of the imaginary part. The small differences between measurement and simulation result occur due to fabrication incertainties, concerning the width and height of the MSL as well as the fact that the nanowires are randomly distributed. It should be emphasized that the simulations and the measurements presented here are performed at remanent state, meaning that no external DC magnetic field is applied. Thus the experimental observation of FMR validates Equation 12.15.

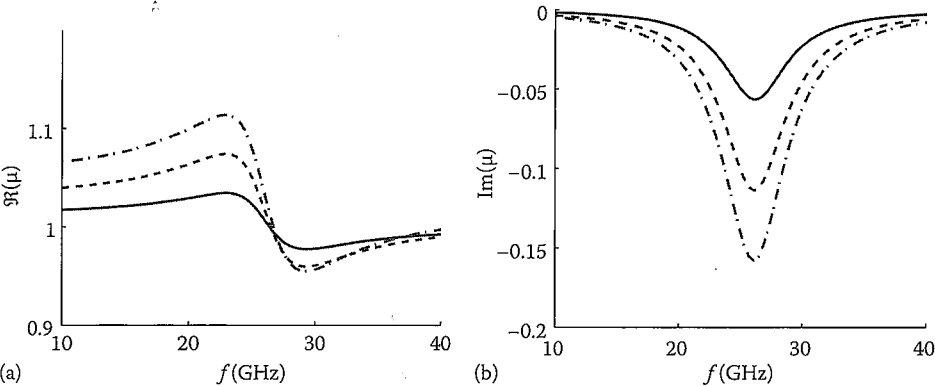


FIGURE 12.6 Simulation of effective permeability of MSLs on composite substrate in a 2% porosity membrane with three heights of Co nanowires, 20% sample (solid), 40% sample (dashed), and 55% sample (dashed dotted): (a) real and (b) imaginary parts. (From Spiegel, J., de la Torre, J., Darques, M., Piraux, L., and Huynen, I., *IEEE Microwave Wireless Component Lett.*, 7, 492, 2007. With permission.)

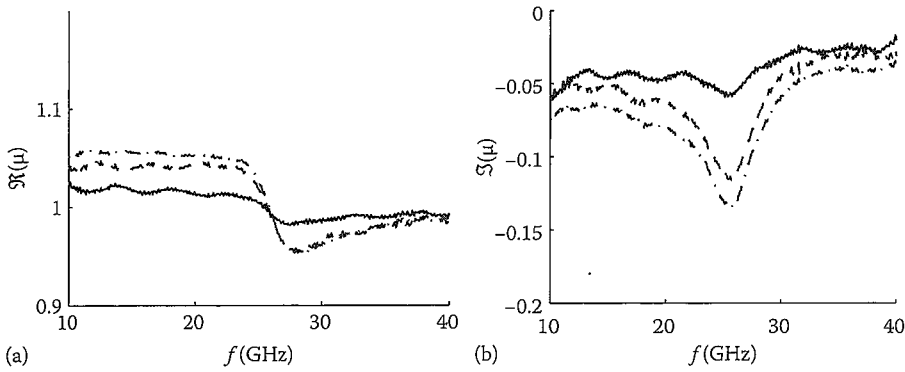


FIGURE 12.7 Extraction of effective permeability from measurements of MSLs on composite substrate in a 2% porosity membrane with three heights of Co nanowires, 20% sample (solid), 40% sample (dashed), and 55% sample (dashed dotted): (a) real and (b) imaginary parts. (From Spiegel, J., de la Torre, J., Darques, M., Piriaux, L., and Huynen, I., *IEEE Microwave Wireless Component Lett.*, 7, 492, 2007. With permission.)

12.2.4 Losses

Various types of losses can be found in nanowired material. The focus here is given to the dielectric losses of the bulk polymer membrane, magnetic and conductive losses of ferromagnetic materials. The effective dielectric losses have been simulated based on the developed theoretical permittivity model:

$$\tan \delta_{\text{eff}} = -\frac{\Im(\epsilon_{\text{eff}})}{\Re(\epsilon_{\text{eff}})}. \quad (12.17)$$

Effective losses were also extracted from measurement by

$$\tan \delta_{\text{eff}} = -\frac{\Im(\epsilon_{\text{eff}} \cdot \mu_{\text{eff}})}{\Re(\epsilon_{\text{eff}} \cdot \mu_{\text{eff}})}. \quad (12.18)$$

Figure 12.8 shows a 3D view of the simulation result. It shows that the loss tangent factor (Equation 12.17) is not depending on the nanowire height. It can also be seen that the dielectric losses decrease with porosity, because the volume of lossy bulk membrane is reduced. To validate the loss model, Figure 12.9a shows the measured loss tangent factor (Equation 12.18) of membranes filled with Cu. The curves show the same results for the filled and for the empty membranes, differing by measurement noise. This confirms that the presence of conductive nanowires does not influence dielectric losses. Comparing these measurement results to those of membranes filled with ferromagnetic Co nanowires, shown in Figure 12.9b, a strong variation of the curves can be seen around 26 GHz for Co wires. The increase of loss occurs due to the FMR [24]. In contrast, beyond the FMR region, the curves follow those of Cu nanowires and empty membrane.

12.2.5 Tunability

Another interesting feature of ferromagnetic nanowired materials is their tunability in frequency by applying an external magnetic DC field. Depending on the porosity of the membrane, the FMR frequency is shifted toward higher frequencies for low porosity and toward lower frequencies for high porosity, due to the strong dipolar coupling between the nanowires (see Section 12.3.3). No change

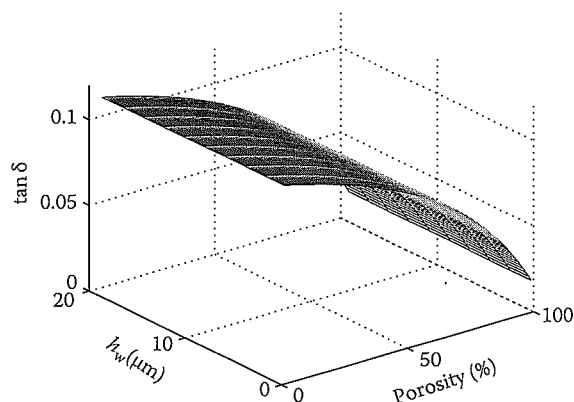


FIGURE 12.8 Dielectric loss simulation depending on porosity of membrane and height of nanowires. (From Spiegel, J., de la Torre, J., Darques, M., Piraux, L., and Huynen, I., *IEEE Microwave Wireless Component Lett.*, 7, 492, 2007. With permission.)

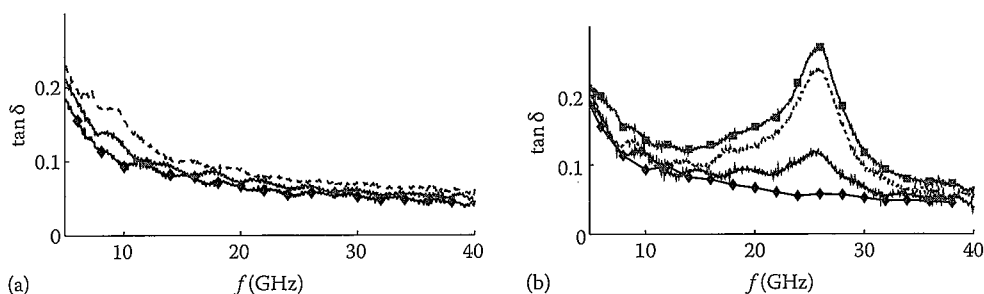


FIGURE 12.9 Loss tangent factor extracted from measurements of MSLs on composite substrate filled with (a) Cu and (b) Co nanowires of 0% (diamond), 20% (solid), 40% (dashed), and 55% (square) of substrate height. (From Spiegel, J., de la Torre, J., Darques, M., Piraux, L., and Huynen, I., *IEEE Microwave Wireless Component Lett.*, 7, 492, 2007. With permission.)

occurs in the magnitude of the permeability components. Figure 12.10 shows the real part of the effective permeability of a MSL, extracted from S -parameter measurements on a low porosity ($p = 10\%$) membrane by applying different external magnetic DC fields up to 10 kOe. The FMR frequency shift toward higher frequencies is clearly observable.

12.3 Feasibility of Tunable Single- and Double-Negative Planar Metamaterials Based on Nanowired Substrate

Metamaterials can be defined as artificial effectively homogeneous electromagnetic structures with unusual properties not readily available in nature [33]. They include single-negative (SNG) material, where either the permeability or the permittivity is negative and double-negative (DNG) material, where both permeability and permittivity are negative. These materials consist of a periodic repetition of basic cells, which have a length shorter than the guided wavelength. This is the reason why at macrolevel the medium behaves like a continuous medium with effective electromagnetic properties. They can find application in the microwave range, for example, in filters, couplers, and leaky wave

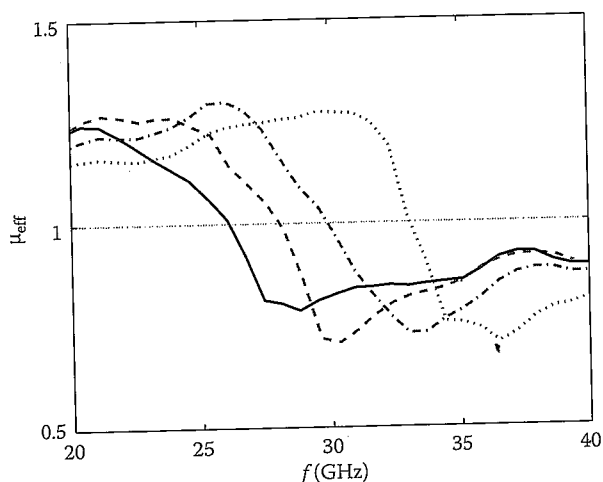


FIGURE 12.10 Real part of effective permeability with applied external magnetic field of 0 kOe (solid), 6 kOe (dashed), 8 kOe (dashed dotted), and 10 kOe (dotted) extracted from measurements of MSL on a low porosity membrane ($p = 10\%$).

antennas. This section presents SNG and DNG materials built out of ferromagnetic nanowired substrates proposed in Section 12.2.1 and the possibility of tuning the material by applying an external magnetic DC field.

12.3.1 Single-Negative Material

Considering the ferromagnetic nanowired substrate, a closer study of the influence of the membrane's porosity and the nanowire height on the effective permeability has been performed, with the result that a SNG material can be obtained, where the permeability is negative. Figure 12.11a shows the simulation of μ_{eff} by changing the porosity (10%, 25%, 35%, and 40%) and keeping the nanowire height at 60% of the substrate height. Figure 12.11b shows additionally the zoom of the simulation for the negative permeability zone. With a porosity of 35%, a bandwidth of 2 GHz can be obtained

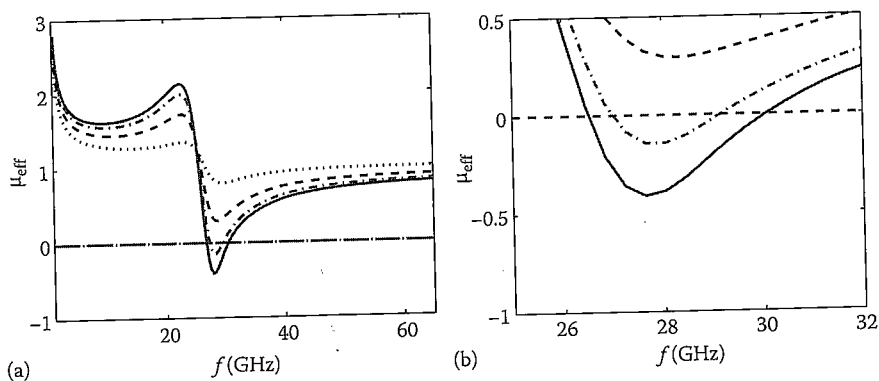


FIGURE 12.11 (a) Simulation of effective permeability by varying the porosity: 10% (dotted), 25% (dashed), 35% (dashed dotted), and 40% (solid), height of nanowires 60% of substrate height; and (b) zoom on negative permeability zone.

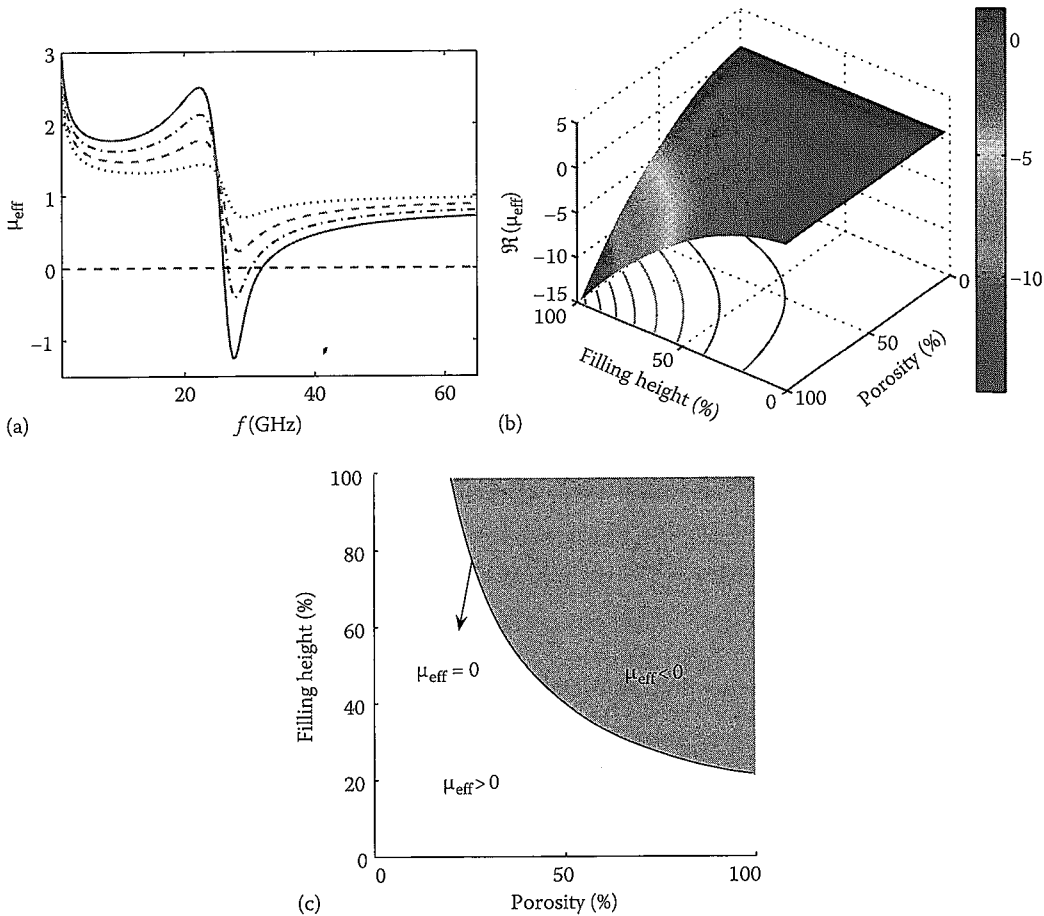


FIGURE 12.12 (a) Simulation of effective permeability by varying the height of the nanowires: 20% (dotted), 40% (dashed), 60% (dashed dotted), and 80% (solid), porosity 40%; (b) 3D view of effective permeability by varying the porosity and the height of the nanowires at $f = 27$ GHz (where effective permeability is minimal); and (c) values of porosity and filling height yielding $\mu_{\text{eff}} = 0$ at $f = 27$ GHz.

and for 40%, a bandwidth of 3.5 GHz, where μ_{eff} is negative. As a result, the porosity has to be very high to obtain a negative permeability. But the porosity is not the only parameter that influences the permeability. If the height of the nanowires is too low, there will be no negative permeability zone. This result is shown in Figure 12.12a, where the porosity is 40% and the height of the nanowires varies from 20% to 80% in steps of 20%. It can be seen that the two parameters are coupled to each other (Figure 12.12b). The porosity has to be high enough and so should the height of the nanowires for obtaining a negative permeability zone: permitted values are given by the upper area in Figure 12.12c. As a result, a high volume of ferromagnetic material is required. This will not be possible with a PC membrane. So the membrane's material has to be changed into a material that provides a higher porosity such as alumina [34].

12.3.2 Double-Negative Material

The SNG material can be extended to a DNG material by adding a shunt inductance, either with a lumped element or by short-circuiting the MSL with nanowires. The equivalent circuit for this DNG

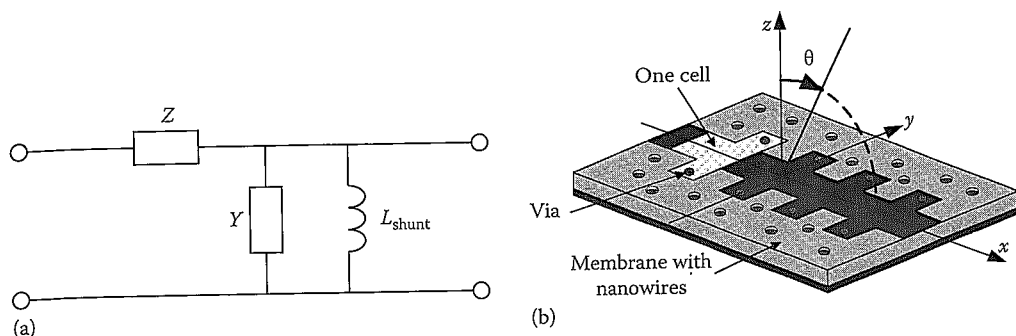


FIGURE 12.13 (a) Equivalent circuit of MSL with shunt inductance and (b) concept of DNG material in MSL technology. Stubs with vias form parallel inductor L_{shunt} .

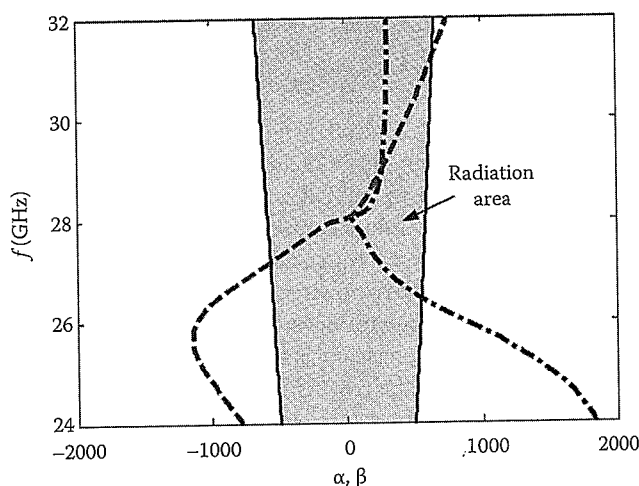


FIGURE 12.14 Radiation cone with attenuation constant α (dashed dotted) and phase constant β (dashed) for microstrip topology of Figure 12.13b on nanowired substrate with porosity $p = 40\%$ and filling height $= 60\%$.

material, based on the equivalent circuit for an MSL [35] lying on the nanowired substrate, is shown in Figure 12.13a. It shows a shunt-inductance L_{shunt} connected in parallel to the admittance per unit length Y_m of the MSL, which represents a negative shunt admittance. This shunt inductance is easy to achieve in planar microstrip technology using a via short-circuit ending (Figure 12.13b). Figure 12.14 shows the extraction of the phase constant β and the attenuation constant α for a microstrip topology on a nanowired substrate of porosity $p = 40\%$ and filling height 60% . Left-handed propagation occurs from about $f = 26$ to 28 GHz, where β is negative and increasing with frequency, indeed in the range where μ_{eff} is negative. This basis cell can be used to create leaky-wave antennas, which support a fast-wave mode. The principal interest of a leaky-wave antenna is the tuning of the beamdirection with frequency. The metamaterial adds the possibility of scanning the radiation beamdirection (angle θ in Figure 12.13b) over 180° . From leaky-wave theory, radiation occurs when propagation constant $|\beta|$ is lower than free-space propagation constant $|k_0|$, i.e., inside the radiation cone, shown in Figure 12.14. The beamdirection scanning is possible, because β varies here continuously from $-|k_0|$ to $|k_0|$ as detailed in [36].

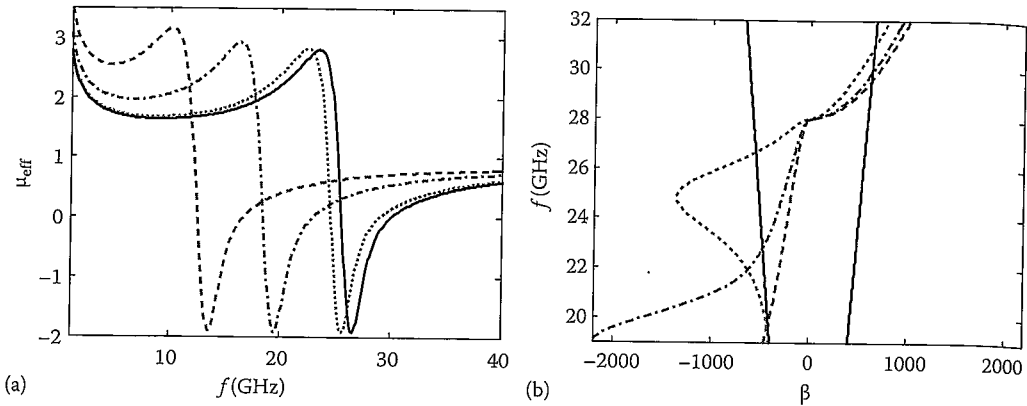


FIGURE 12.15 (a) Real part of effective permeability (with 0 Oe (solid) included) and (b) β with applied external magnetic field 6 (dashed), 8 (dashed dotted), and 10 kOe (dotted).

12.3.3 Tunability

As already mentioned, these metamaterials based on ferromagnetic nanowired substrates are tunable in frequency by applying an external magnetic DC field. Figure 12.15a shows a simulation of the effective permeability of a SNG material with an applied external field of 0 Oe, 6 kOe, 8 kOe, and 10 kOe. The applied magnetic field has no further influence on the curves than shifting them on the frequency axis to lower frequencies. This shift to lower frequencies occurs in high-porosity membranes because of the strong dipolar coupling between the nanowires [37] instead of a shift to higher frequencies as shown in Section 12.2.5 for low porosity membranes. The same behavior is visible for DNG materials. Figure 12.15b shows the result of the phase constant β by applying an external magnetic field of the values mentioned above. As a result, the frequency zone, where the structure has a left-handed behavior, can be varied with the applied external magnetic field. So the beamangle of the antenna can be varied either by frequency or by applying an external magnetic field by one fixed frequency.

12.4 Transmission Line Modeling of Stopband Effects

The TLA analyzes materials and designs devices for electronic circuits. This approach considers the structure under scope as a combination of equivalent TLs supporting the propagation of signals. It yields to important information about its frequency response, in terms of bandwidth and resonant frequencies induced by material properties or by reflection of waves inside the device.

12.4.1 Electromagnetic Response of EBG Dielectric Substrate

Assuming that the EBG effect is due to a periodic change of characteristic impedance along a guiding planar structure [11,12,14], the simple model illustrated in Figure 12.16a can be used to predict its frequency response [38,39]. It can be applied, for example, to an MSL lying on a dielectric substrate where periodic rectangular holes filled by air are etched. Figure 12.16b shows the resulting structure, having N ($= 6$) MSL sections over a dielectric substrate, and $N - 1$ ($= 5$) MSL sections over air holes. Applying the notations of Figure 12.16a to the structure of Figure 12.16b results in

$$Z_{c1} \propto \frac{1}{n_{\text{eff1}}}, \quad n_{\text{eff1}} = \sqrt{\epsilon_{r1}}, \quad \gamma_1 = j\beta_1 = jn_{\text{eff1}}k_0 \quad (12.19)$$

$$Z_{c2} \propto \frac{1}{n_{\text{eff2}}}, \quad n_{\text{eff2}} = \sqrt{\epsilon_{r2}}, \quad \gamma_2 = j\beta_2 = jn_{\text{eff2}}k_0, \quad (12.20)$$

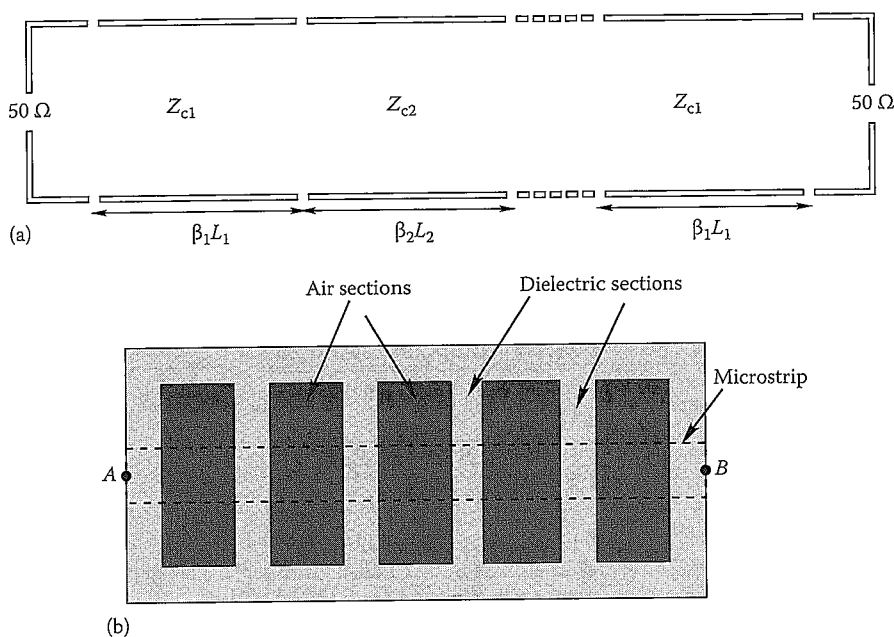


FIGURE 12.16 Topologies under scope for dielectric EBG: (a) equivalent TL topology for the modeling of planar EBG structures and (b) topology of the MSL on the dielectric EBG substrate with $N = 6$ and substrate permittivity of 10.2, according to [14]. (From Saib, A. and Huynen, I., *J. Opt. A: Pure Appl. Opt.*, 7(2), S124, 2005. With permission.)

where

- n_{effi} is the effective refractive index associated with propagation along TL section of type i
- Z_{ci} is its characteristic impedance
- $\gamma_i = j\beta_i$ is its complex propagation constant
- $k_0 = \omega/c_0$ is the free-space propagation coefficient
- c_0 is the velocity of light in vacuum

Values of refractive indexes are calculated for MSL sections of constant width, lying on dielectric substrates of respective dielectric constant $\epsilon_{r1} = \epsilon_{rb} = 10.2$ and $\epsilon_{r2} = \epsilon_{rair} = 1$.

Since the topology in Figure 12.16 is made of N sections of type 1 and $N - 1$ sections of type 2, the chain matrix formalism described in [40] is applied to each type of section. Its corresponding chain matrix noted $[T]_i$ is given by

$$[T]_i = \begin{bmatrix} A_i & B_i \\ C_i & D_i \end{bmatrix} = \begin{bmatrix} \text{ch}(\gamma_i L_i) & Z_{ci} \text{sh}(\gamma_i L_i) \\ \frac{1}{Z_{ci}} \text{sh}(\gamma_i L_i) & \text{ch}(\gamma_i L_i) \end{bmatrix}. \quad (12.21)$$

Coefficients of the chain matrix include parameters n_{effi} , Z_{ci} and the length L_i of the section. The cascade of N TL sections of type 1 and $N - 1$ TL sections of type 2 is represented as the product of their chain matrices:

$$[T]_{\text{tot}} = \begin{bmatrix} A & B \\ C & D \end{bmatrix} = \{[T]_1 \cdot [T]_2\}^{N-1} \cdot [T]_1. \quad (12.22)$$

The total chain matrix is then converted into a 2×2 scattering matrix $[S]$ referenced to the reference impedances Z_R at the input and output ports using classical matrix conversion formulas [41]:

$$[S] = \begin{bmatrix} S_{11} & S_{21} \\ S_{21} & S_{22} \end{bmatrix}, \quad (12.23)$$

where

$$S_{11} = \frac{AZ_R + B - CZ_R^*Z_R - DZ_R^*}{AZ_1 + B + CZ_RZ_R + DZ_R} \quad (12.24)$$

$$S_{12} = \frac{2(AD - BC)\sqrt{\Re(Z_R)\Re(Z_R)}}{AZ_R + B + CZ_RZ_R + DZ_R} \quad (12.25)$$

$$S_{21} = \frac{2\sqrt{\Re(Z_R)\Re(Z_R)}}{AZ_R + B + CZ_RZ_R + DZ_R} \quad (12.26)$$

$$S_{22} = \frac{-AZ_R^* + B - CZ_RZ_R^* - DZ_R}{AZ_R + B + CZ_RZ_R + DZ_R}. \quad (12.27)$$

As a result, the transmission coefficient S_{21} between input and output of the total structure is obtained. Studying the frequency response of this coefficient will illustrate the behavior of the EBG effect. Figure 12.17a shows the dependence of the bandgap shape on the number of sections N : the attenuation level at the central frequency of the stopband increases with N , so that the structure will show a perfect bandgap when $N \rightarrow \infty$. At fixed number of sections ($N = 6$, as in Figure 12.16b), increasing the ratio (or mismatch) between characteristic impedances Z_{c2} and Z_{c1} , the width of the stopband and the maximal attenuation are increased (Figure 12.17b). This illustrates that the bandgap is created by a difference between impedances Z_{c2} and Z_{c1} , since it totally disappears when their ratio $r = 1$. Structures of Figure 12.17a and b are designed and simulated assuming that phase shifts $\beta_i L_i$ of sections of type 1 and 2 are identical. Their lengths L_i verify:

$$L_1 = \frac{\beta_2}{\beta_1} L_2 = \frac{n_{\text{eff}2}}{n_{\text{eff}1}} L_2. \quad (12.28)$$

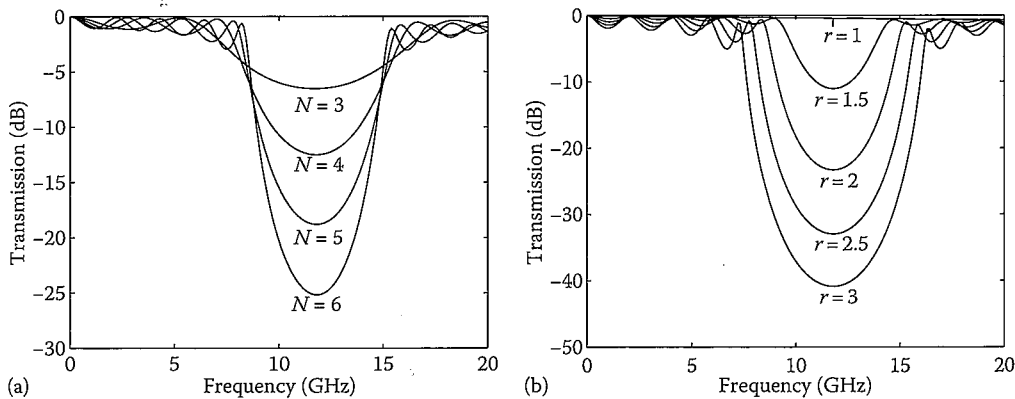


FIGURE 12.17 Simulated transmission coefficient for configuration of Figure 12.16b when varying (a) the number N of dielectric sections. (From Saib, A. and Huynen, I., *Electromagnetics*, 26, 261, 2006. With permission.) (b) Value of the characteristic impedance ratio $r = Z_{c2}/Z_{c1}$. (From Saib, A. and Huynen, I., *J. Opt. A: Pure Appl. Opt.*, 7(2), S124, 2005. With permission.)

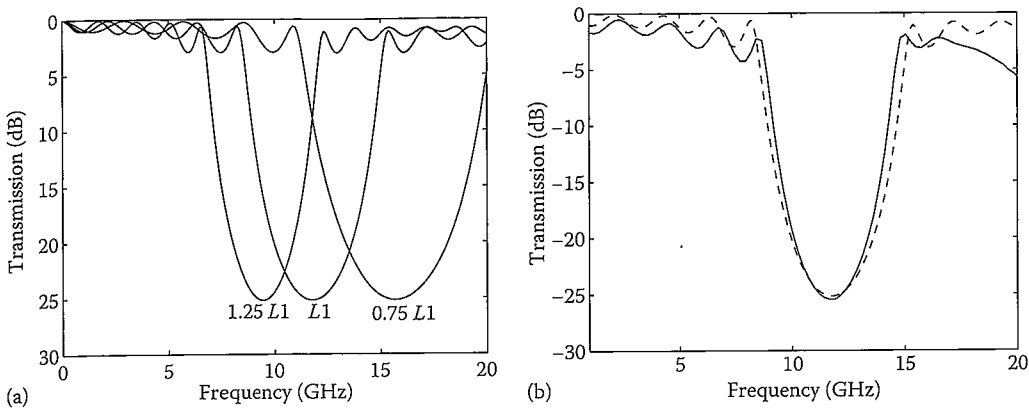


FIGURE 12.18 (a) Simulated transmission coefficient for configuration of Figure 12.16b when varying the length L_1 , while keeping $\beta_1 L_1 = \beta_2 L_2$. (From Saib, A. and Huynen, I., *Electromagnetics*, 26, 261, 2006. With permission.) (b) Measured transmission coefficient (solid line) corresponding to that simulated (dashed line) in Figure 12.17a for $N = 6$ as in Figure 12.16b. (From Saib, A. and Huynen, I., *J. Opt. A: Pure Appl. Opt.*, 7(2), S124, 2005. With permission.)

The center frequency of the bandgap, noted f_0 , is derived from the condition

$$\beta_1 L_1 = \beta_2 L_2 = \pi/2 \quad (12.29)$$

as

$$f_0 = \frac{c_0}{4L_2 n_{\text{eff}2}} = \frac{c_0}{4L_1 n_{\text{eff}1}}. \quad (12.30)$$

At a fixed number of sections ($N = 6$), varying the lengths L_i of the sections does not modify the shape of the bandgap (normalized width of stopband and maximal attenuation level), but modifies its center frequency (Figure 12.18a), which corresponds to phase shifts $\beta_i L_i$ equal to $\pi/2$. Finally, a validation of this model is illustrated at Figure 12.18b for the topology of Figure 12.16b, with $N = 6$. The measured transmission coefficient (solid line) agrees very well with the simulated one (dashed line, identical to curve $N = 6$ in Figure 12.17a).

It should be noted that the relationship between S parameters and chain matrix elements can be inverted. Measured S parameters of a TL can be converted to chain matrix elements [41], from which the complex propagation constant γ_i can be extracted. This technique was used in Section 12.2.2 for obtaining the product $\sqrt{\epsilon_{\text{eff}} \cdot \mu_{\text{eff}}}$ of a TL on a nanowired substrate.

12.4.2 Tunable Electromagnetic Response of Nanowired Substrate

Since the analytical TL model developed here is based on the determination of the propagation constant and the characteristic impedance of MSLs on nanowired substrate, the propagation constant γ_{eff} of the MSL on nanowired substrate is given by the variational formulation (Equation 12.7). Its characteristic impedance $Z_{c \text{ eff}}$ is derived from the characteristic impedance $Z_{c \text{ p}}$ of a MSL of same width W_s lying on the bulk polymer substrate of dielectric constant ϵ_{rb} as follows:

$$Z_{c \text{ eff}} = Z_{c \text{ p}} \sqrt{\frac{\epsilon_{\text{rb}} \cdot \mu_{\text{eff}}}{\epsilon_{\text{eff}}}}. \quad (12.31)$$

The permittivity ϵ_{eff} of the nanowired substrate is given by the analytical formulation (Equation 12.4) developed in Section 12.2.2, while μ_{eff} is obtained from Equation 12.16. Once the propagation constant

and the characteristic impedance of the MSL lying on the nanowired substrate are determined, the corresponding scattering matrix $[S]$ of a TL section of length L is given by Equations 12.21 and 12.23. The variational model is validated by vector network analyzer (VNA) measurements on a CoFe—nanowired substrate having a porosity of 3.5% and a nanowire diameter of 30 nm. The comparison between simulation (dashed lines) and measurement (solid lines) is shown in Figure 12.19. A very good agreement is observed both in magnitude and phase of the scattering parameters (transmission S_{21} and reflection S_{11}). The oscillations observed on the scattering parameters and the high reflection level are due to impedance mismatch. Indeed, the width of the microstrip is $w = 150 \mu\text{m}$ due to technological constraints, instead of $w = 60 \mu\text{m}$ necessary to have a 50Ω characteristic impedance. The position and width of the FMR is perfectly predicted by the variational-TL model: a strong absorption associated to FMR losses is observed at zero DC field, inducing a stopband filter behavior. It is due to the extremum in the imaginary part of the effective permeability observed in Figures 12.16 and 12.17. The fact that the FMR is observed without any external DC magnetic field is related to Equation 12.15. The low transmission level observed outside the FMR range is mainly due to dielectric losses of the polymer membrane and conductor losses of the metallic layers. Those two effects are correctly modeled. It should be noted that the nanowired substrate MSL is a reciprocal and symmetrical device (i.e., $S_{12} = S_{21}$ and $S_{11} = S_{22}$). This is due to the symmetrical electromagnetic fields configuration in the transverse section of the microstrip, which is not favorable for nonreciprocal

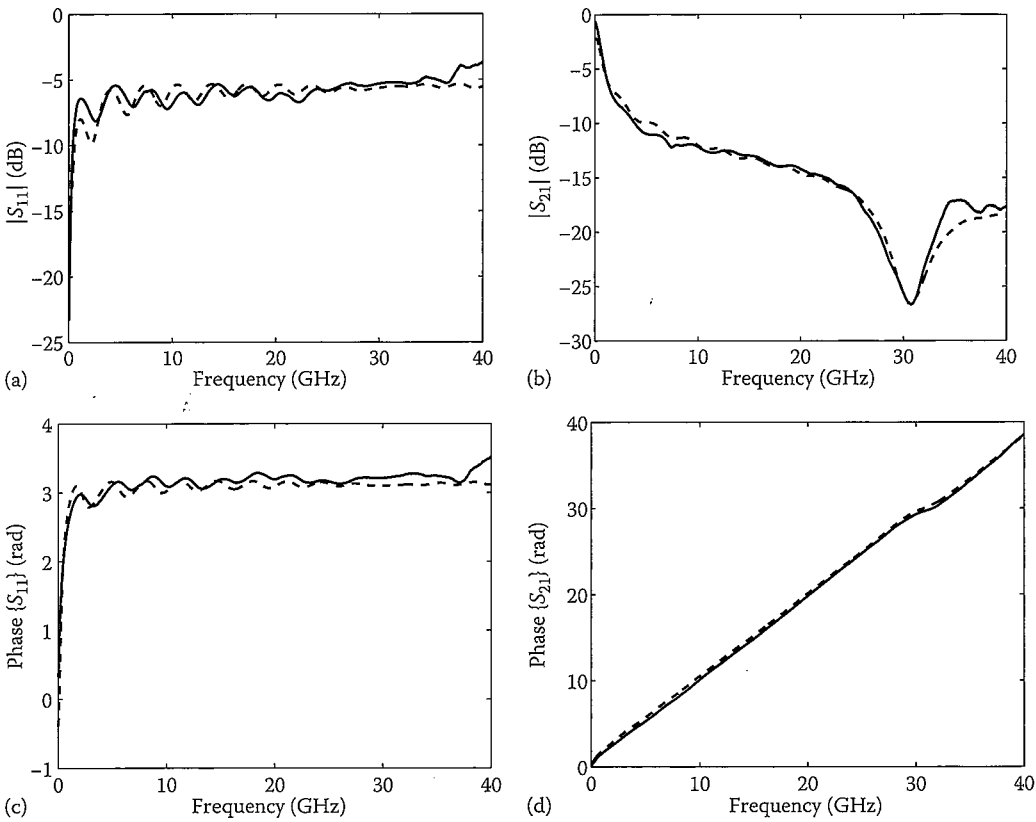


FIGURE 12.19 Measured (solid lines) and simulated (dashed lines) scattering parameters of MSL on a CoFe—nanowired substrate ($p = 3.5\%$ and $D = 30 \text{ nm}$), MSL ($W_s = 150 \mu\text{m}$, $h_{\text{poly}} = 21 \mu\text{m}$, and $L = 1.75 \text{ cm}$): (a) $|S_{11}|$, (b) $|S_{21}|$, (c) phase $\{S_{11}\}$, and (d) phase $\{S_{21}\}$.

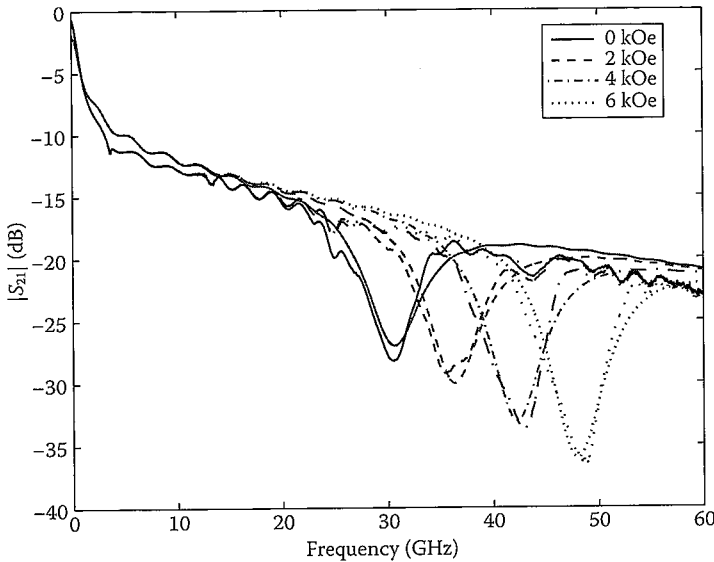


FIGURE 12.20 Measured (black curves) and simulated (gray curves) transmission factor of MSL on a CoFe—nanowired substrate ($p = 3.5\%$ and $D = 30$ nm), MSL ($w = 150$ μm , $h_{\text{poly}} = 21$ μm , and $L = 1.75$ cm) versus the frequency for a DC magnetic field $H_{\text{dc}} = 2, 3$, and 4 kOe (dashed, dashdotted, and dotted lines, respectively) applied parallel to the wire axis and for $H_{\text{dc}} = 0$ kOe (solid line).

propagation along the microstrip length. As a consequence of Equation 12.15, the FMR absorption and hence the stopband effect observed in the transmission factor are tunable with the application of a static DC field. When a DC magnetic field is applied to the structure in the direction parallel to the wire axis (Figure 12.20), the FMR is shifted toward high frequencies. Again the agreement between the simulations (gray curves in Figure 12.20) and the measurements (black curves in Figure 12.20) is very good. The variational model is able to describe the propagation of electromagnetic waves inside the nanowired MSLs at any magnetic state.

12.5 Demonstration of Planar MEBG Structures

12.5.1 Topology

The TLA validated for planar dielectric EBGs can be successfully applied to the design and experimental demonstration of MEBG devices operating at microwaves. The idea is to transpose the dielectric/air structure of Figure 12.16b to a MEBG structure. To this end, dielectric sections in the topology of Figure 12.16b are replaced by a particular composite substrate: it consists of an array of parallel ferromagnetic nanowires electrodeposited into a porous polymer membrane, following the method reported in [24,42]. The resulting MEBG topology is illustrated in Figure 12.21a: the air sections are now replaced by the bulk polymer membrane, while each nanowired area (polymer + nanowires) forms a composite magnetic material replacing dielectric sections of Figure 12.16b. Figure 12.21b shows the picture of an in-house fabricated MEBG.

12.5.2 Design of the MEBG

The design of a MEBG is exemplary shown with $\text{Ni}_{80}\text{Fe}_{20}$ nanowires. The resulting scalar effective permeability (Equation 12.16) exhibits a measured FMR without DC field around 16 GHz [43],

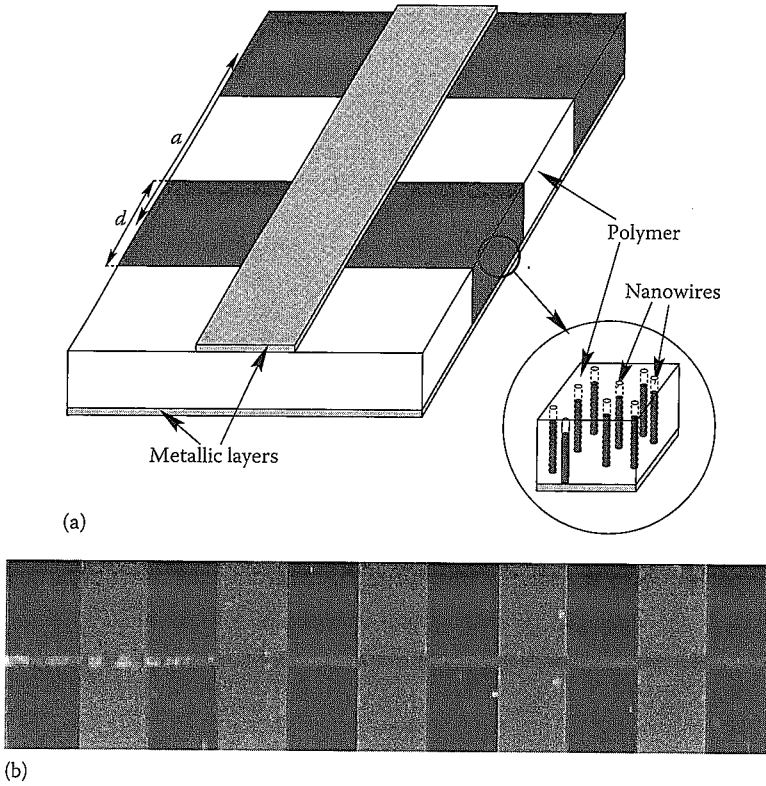


FIGURE 12.21 (a) Topology of MEBG and (b) in-house fabricated MEBG. (From Saib, A. and Huynen, I., *J. Opt. A: Pure Appl. Opt.*, 7(2), S124, 2005. With permission.)

according to Equation 12.11. Of course the relative permeability μ_{rb} of the bulk polymer sections is equal to 1. It can be easily verified from Equations 12.11 through 12.16, and according to [32], that above the FMR frequency the real part of μ_{eff} will be less than 1. Equations 12.19 and 12.20 valid for the dielectric case of Section 12.4 becomes now

$$Z_{c1} = Z_{c\,eff}, \quad n_{eff1} = \sqrt{\epsilon_{eff} \cdot \mu_{eff}}, \quad \gamma_1 = \gamma_{eff} = jn_{eff1}k_o \quad (12.32)$$

$$Z_{c2} = Z_{c\,p}, \quad n_{eff2} = \sqrt{\epsilon_b}, \quad \gamma_2 = j\beta_2 = jn_{eff2}k_o, \quad (12.33)$$

where $Z_{c\,p}$ and $Z_{c\,eff}$ are characteristic impedances for the bulk polymer and nanowired MSL sections, respectively, while the permittivity ϵ_{eff} of the nanowired substrate is given by the analytical formulation in Equation 12.4, γ_{eff} is the propagation constant of the nanowired section given by the variational formulation in Equation 12.7, and μ_{eff} is obtained from Equation 12.16. By virtue of Equation 12.31, and since $\Re(\mu_{eff}) < 1$ above the FMR, which occurs at 16 GHz in Figure 12.22a, $Z_{c\,eff} < Z_{c\,p}$ in this frequency region, so that a EBG effect is possible ($r > 1$ in Figure 12.17b). The characteristic impedances $Z_{c2} = Z_{c\,p}$ and $Z_{c1} = Z_{c\,eff}$ for the polymer and nanowired sections are shown in Figure 12.22b. Adjusting then the length of the polymer and nanowired sections to make the phase shift of each section equal to $\pi/2$ at 29 GHz enables to form a bandgap around this frequency, as shown by the simulation of transmission parameter S_{21} (thin line) in Figure 12.23. The other

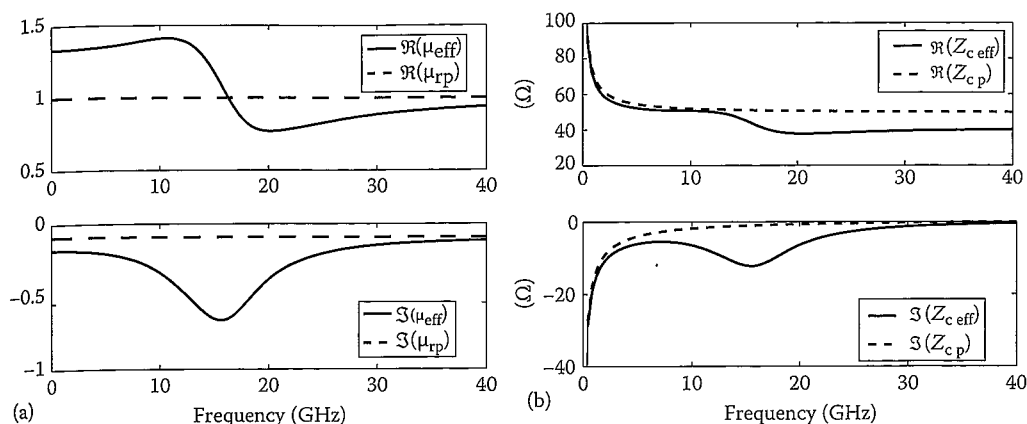


FIGURE 12.22 (a) Effective permeability of magnetic nanowired (solid) and relative permeability of polymer (dashed) sections, and (b) characteristic impedance of magnetic nanowired (solid) and polymer (dashed) sections. (From Saib, A. and Huynen, I., *Electromagnetics*, 26, 261, 2006. With permission.)

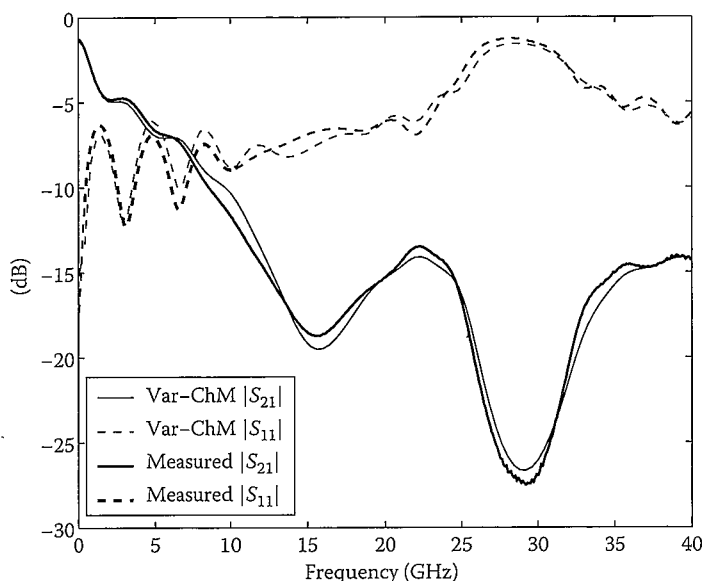


FIGURE 12.23 Measured (thick lines) and simulated (thin lines) transmission (solid lines) and reflection (dashed lines) coefficients of the MEBG of Figure 12.21 at the remanent state, calculated using the variational model. (From Saib, A. and Huynen, I., *Electromagnetics*, 26, 261, 2006. With permission.)

attenuation peak observed at 16 GHz is due to the FMR absorption effect, which induces an extremum in the imaginary part of μ_{eff} at this frequency. The origins of the two peaks are indeed clearly different: reflection coefficient S_{11} is high at 29 GHz, in accordance with a stopband effect, while around 16 GHz, it remains moderate, proving that absorption at this frequency is due to FMR losses. Figure 12.23 also shows that the modeling of the EBG effect involving magnetic materials (thin lines) is successfully validated by measurements (thick lines). To our best knowledge, it is the first time that a magnetic PBG effect is realized experimentally at microwaves [15]. As mentioned in Section 12.2.2 for the FMR peak,

the MEBG peak is not due to any periodicity of the nanowires, since they are randomly distributed, but due to the periodic cascade of polymer sections and nanowired composite sections. However, this MEBG peak is possible due to the ferromagnetic nature of nanowires, which induces still a significant change in characteristic impedance at frequencies far above the FMR (Figure 12.22b). Various samples with different periodic parameters (a , d), where a is the cell length and d is the length of the magnetic section, as indicated in Figure 12.22a, have been fabricated using $\text{Ni}_{80}\text{Fe}_{20}$ nanowires (Figure 12.21). Figure 12.24 shows the simulated and measured transmission and reflection of two MEBGs without applied static magnetic field: MEBG 1 (solid lines) with $a = 3.636$ mm, $d = 1.818$ mm, and MEBG 2 (dashed lines) with $a = 3.076$ mm, $d = 1.538$ mm. The total length of the MEBG structures is 2 cm. In the transmission curves, we can see two peaks. The first one, at 16 GHz, corresponds to the FMR frequency of $\text{Ni}_{80}\text{Fe}_{20}$ nanowires at the remanent state. The second one is due to the EBG effect at 21 GHz for MEBG 1 and 26 GHz for MEBG 2. These last two frequencies are in good agreement with those predicted by the simulation (Figure 12.24a). In all samples considered here, we take $a = 2d$ in

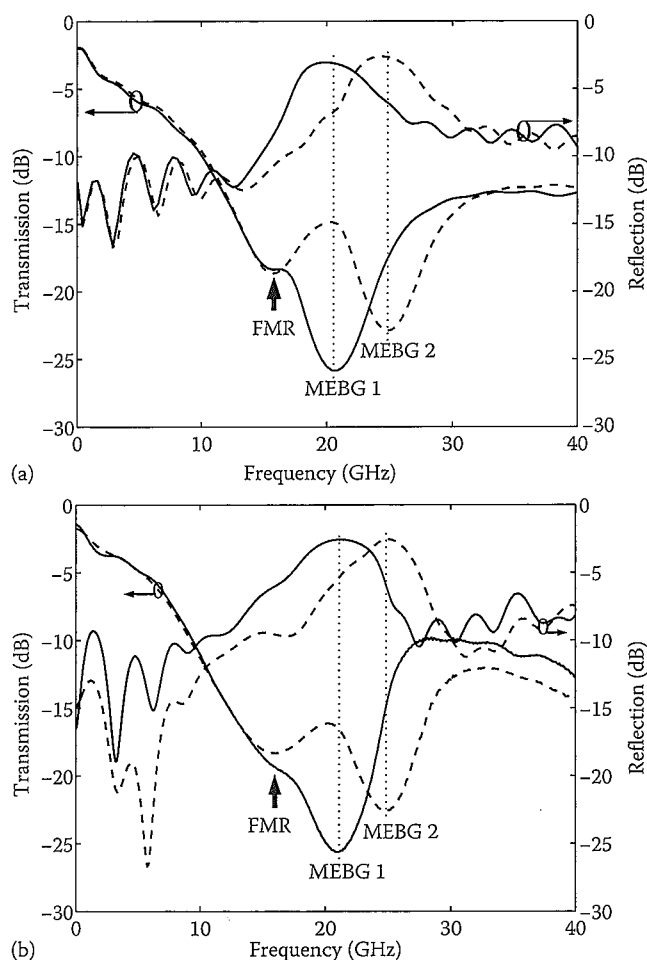


FIGURE 12.24 (a) Simulated and (b) measured microwave transmission and reflection of two MEBGs at zero DC magnetic field: MEBG 1 (solid lines) with $a = 3.636$ mm, $d = 1.818$ mm and MEBG 2 (dashed lines) with $a = 3.078$ mm, $d = 1.538$ mm.

order to simplify the design. The FMR peak is located at the same position for both MEBG 1 and MEBG 2 because it only depends on the ferromagnetic material filling the pores [30].

The possibility to design EBG materials with ferromagnetic nanowired substrate at the remanent state ($H_{dc} = 0$) has been demonstrated. Even if the FMR phenomenon, which is due to the ferromagnetic nature of nanowires, is necessary for the creation of the MEBG; the position of the bandgap peak can be adjusted separately by the periodicity. From Figure 12.23, it is observed that the main challenge for designing efficient microwave devices is the reduction of losses: outside the stopband or FMR absorption band, the transmission level has to be kept close to 0 dB. This is mainly due to the high dielectric losses of the porous membrane (measured loss tangent $\tan \delta = 0.1$). Solutions currently investigated consist of replacing the PC polymer template by other porous templates having lower dielectric losses as alumina.

We observe here an EBG effect with a periodic material where no significant contrast in the refractive index is present: $\sqrt{\epsilon_{eff} \cdot \mu_{eff}} = \sqrt{4.6 \cdot 0.6} \approx 1.66$ and $\sqrt{\epsilon_{rb} \cdot \mu_{rb}} = \sqrt{2.89 \cdot 1} \approx 1.7$ at the center frequency of the bandgap. This confirms the role of the wave impedance ($\sqrt{\mu/\epsilon}$) in the creation of EBGs with dielectric and magnetic properties. Let us compare the wave impedances of the magnetic and bulk polymer sections. Around the center frequency of the bandgap $\sqrt{\mu_{eff}/\epsilon_{eff}} \approx 0.36$ and $\sqrt{\mu_{rb}/\epsilon_{rb}} \approx 0.59$, which corresponds to a considerable contrast close to 40%. In general, for classical dielectric EBGs, a high contrast in the refractive index leads necessarily to a similar contrast in the wave impedance, since they are inversely proportional. For our MEBG, even if there is no contrast in the refractive index, we still have a significant contrast in the wave impedance due to the magnetic properties and especially for $\mu < 1$. Therefore, a magnetic EBG can be seen as the dual of a dielectric EBG.

12.5.3 Tunable Electromagnetic Response of MEBG

Figure 12.25 shows the effect of a DC magnetic field applied parallel to the nanowires on the MEBG of Figure 12.21. Since the bandgap is located at 29 GHz far away from the FMR frequency, the influence

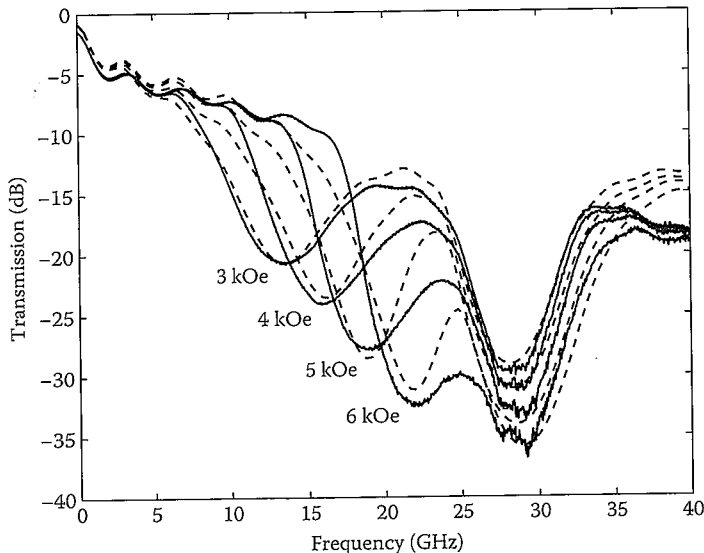


FIGURE 12.25 Measured (solid lines) and simulated (dashed lines) transmission coefficients of the MEBG of Figure 12.21 under a static magnetic field H_{dc} ranging from 3 to 6 kOe applied parallel to the nanowires.

of the external magnetic field on the MEBG peak and on the FMR peak can be seen separately. If the FMR and MEBG frequencies are close to each other, the resulting peaks will be coupled and the effects of the magnetic field on each one cannot be dissociated. The position of the FMR absorption varies linearly with the static magnetic field. On the other hand, the MEBG peak remains almost at the same frequency. This weak influence of the magnetic field on the MEBG can be explained by the small variation at 29 GHz of the magnetic permeability μ_{eff} within the DC magnetic field range considered in Figure 12.25. The FMR equation combined with the variational model of the nanowired substrate gives a variation between 0.6 and 0.7 for the real part of μ_{eff} , which is too small to affect the position of the bandgap.

12.6 Defect Modes

It is known that very sharp transmission peaks in the bandgap, called defect modes, are created in EBG structures, when the periodicity is broken locally inside the array. In this section, we first present and validate experimentally a theoretical explanation of defect mode operations in finite dielectric EBG structure of Section 12.4, via the simple TL formalism used in the previous sections. Using this formalism, we then investigate the effect of defects on the frequency response of the MEBG structure of Section 12.5.

12.6.1 Defect Modes in a Dielectric EBG Material

Using the TLA, modeling the presence of a defect just requires modification of the TL parameters of the section where the defect is present: the defect can be created by modifying either the material of the section (dielectric constant or permeability) or its length [44]. As an example, Figure 12.26a illustrates the creation of a defect in the topology of Figure 12.16b [14]: one of the air sections is filled by the dielectric substrate of permittivity $\epsilon_r = 10.2$, so that in the modeling of the corresponding section, β_2 and Z_{c2} of Equation 12.20 have to be replaced by β_1 and Z_{c1} , respectively. Using the chain matrix formalism described in Section 12.4, the simulated transmission coefficients for the three different configurations of Figure 12.26a can be calculated. The position of the defect is expected to have an influence on the frequency response of the structure since the array is finite (11 sections) [14]. Figure 12.26b shows the simulated transmission coefficients for four different configurations: when no defect is present in the structure (solid line, identical to curve $N = 6$ in Section 12.4, Figure 12.17a, corresponding to Figure 12.16b) and when the defect is moved over the EBG length (the dashed line corresponds to the configuration (1) of Figure 12.26a, the line with dots to (2), and the dashed-dotted line to (3)). The sensitivity of the sharp transmission peak to the defect position can be simply explained in the following manner [14]. From Figure 12.17a, it is known that the maximal attenuation in the bandgap for a finite array is an increasing function of the number of periodic cells, so that for an infinite array, the transmission in the bandgap is expected to be zero, and the reflection is total. Using the concept of highly ideal reflective mirror mentioned by Yablonovitch [4,18] and Tiwari et al. [13], or the analogy in optics with the Fabry-Perrot cavity between two perfect Bragg reflectors, the infinite array can indeed be viewed as a perfect reflector. The structures of Figure 12.26a consist of a left and a right EBG material separated by a defect. Each (left or right) EBG, having less periodic cells than the reference structure of Figure 12.16b, is a less efficient reflector and exhibits a higher transmission level in the bandgap. When the defect is centered (Figure 12.26a(1)), the structure behaves like two identical finite periodic arrays, each having three dielectric sections and two air sections, and they are separated by a TL section on the dielectric substrate: the defect. The two periodic arrays have bandgaps exactly situated in the same frequency band and with the same intensity, since they have the same periodicity and length. They behave like two symmetrical reflectors creating the same strong value of reflection coefficient at the input and output of the defect section [14].

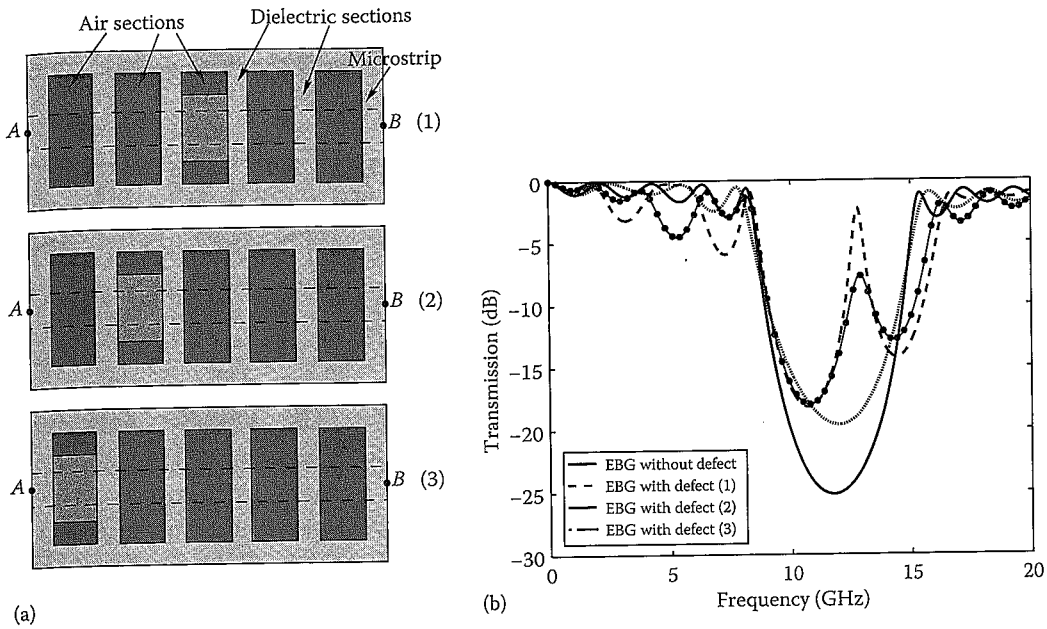


FIGURE 12.26 (a) Topology of the dielectric EBG with three defect positions, and (b) simulated transmission coefficients for three configurations of (a) and for the EBG without defect of Figure 12.16b. (From Saib, A. and Huynen, I., *J. Opt. A: Pure Appl. Opt.*, 7(2), S124, 2005. With permission.)

Hence, at a specific electrical length (occurring at 13 GHz), the defect creates an adequate phase shift between the two reflectors to allow resonance inside the defect section. This later acts like a resonant cavity, yielding a sharp transmission peak (dashed curve in Figure 12.26b). On the contrary, when the defect is moved to the left (Figure 12.26a(2) and (3)), the symmetry is broken: the left reflector, having a smaller number of sections than the right reflector, induces a smaller value of reflection level at the input of the defect section than that at the output. Hence, even if the phase shift created by the defect section satisfies the resonance condition, the cavity is imperfect because the quality of the left reflector is lowered by reducing the number of sections.

The simulation results, shown in Figure 12.26b, are successfully validated by the measurement of Figure 12.27a. Therefore, the model can be used for computing reflection coefficients seen in the input plane of the defect section from the left (Γ_{left}) and from the right (Γ_{right}), for each of the three positions of defect shown in Figure 12.26a. The expressions of those reflection coefficients are

$$\Gamma_{\text{left}} = \frac{Z_{\text{in left}} - Z_{c2}}{Z_{\text{in left}} + Z_{c2}}, \quad \Gamma_{\text{right}} = \frac{Z_{\text{in right}} - Z_{c2}}{Z_{\text{in right}} + Z_{c2}}, \quad (12.34)$$

where $Z_{\text{in left}}$ and $Z_{\text{in right}}$ are the impedances in the input plane of the defect, seen from the left and the right, respectively. They can be easily computed from the components of the global chain matrix of (right reflector + defect section), and of the global chain matrix of left reflector.

Figure 12.27b shows the magnitude and phase of the product $\Gamma_{\text{left}} \cdot \Gamma_{\text{right}}$. The resonance condition inside the defect can be expressed by the fact that the impedance seen from the right $Z_{\text{in right}}$ in the input plane of the defect section is equal to the opposite of the impedance seen from the left $Z_{\text{in left}}$ in this input plane [14]. In terms of the reflection coefficient, this means that the reflection coefficient seen from the left Γ_{left} must be equal to the inverse of the reflection coefficient seen from the right Γ_{right} , or that the product of those two reflection coefficients (which are complex numbers) should

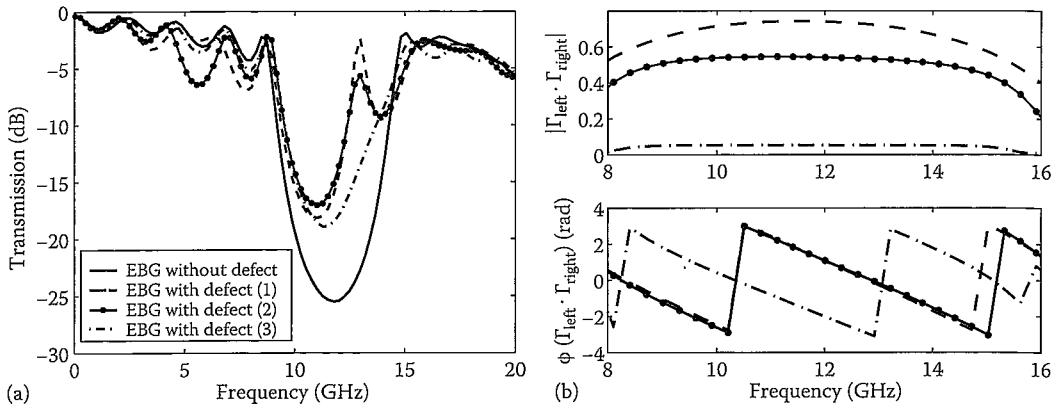


FIGURE 12.27 (a) Measured transmission coefficients corresponding to those simulated in Figure 12.26b, according to [14] and (b) calculated magnitude and phase of the product $\Gamma_{\text{left}} \cdot \Gamma_{\text{right}}$ for the three defect positions of Figure 12.26a. (From Saib, A. and Huynen, I., *J. Opt. A: Pure Appl. Opt.*, 7(2), S124, 2005. With permission.)

be equal to unity. It is observed from Figure 12.27b that around 13 GHz, for a centered defect, the product is closer to 1 than for the two other positions of the defect. At this frequency, which corresponds to the phase of the product equal to zero, the defect mode appears. When the defect moves to the left (Figure 12.26a(2) and (3)), the zero phase condition is still fulfilled, but with a lowered magnitude of the product. Therefore, the defect peak transmission is lowered in the configuration (2) and completely disappears in configuration (3) where the product magnitude is close to zero.

12.6.2 Defect Modes in MEBG

The aim of this section is to show the possibility of having tunable defect modes in an MEBG structure. Therefore, a theoretical explanation of defect modes operation is proposed in a structure of order 21. Furthermore, the effect of defect position and length on the frequency response of the structure is discussed. Figure 12.29 shows the transmission coefficient simulated using the variational model (Equation 12.7) for the configurations depicted in Figure 12.28b through d (solid lines), compared to that of Figure 12.28a (dashed line), where no defect is present (taken as reference). The dark sections of Figure 12.28 have the magnetic properties of $\text{Ni}_{80}\text{Fe}_{20}$ nanowires and the white sections have same properties as the PC membranes. The defect of the MEBG in Figure 12.28b is obtained by increasing the length of the magnetic section situated at the center of the structure of Figure 12.28a from $l = d$ to $l = 2d$ (with $d = \lambda_g/4$ at 56 GHz, λ_g being the guided wavelength). A very sharp transmission peak in the middle of the bandgap (around 56 GHz in Figure 12.29), which is typical signature of defect modes, is observed. The first peaks, situated at 16 GHz, correspond to the FMR phenomenon of $\text{Ni}_{80}\text{Fe}_{20}$ nanowires in the absence of an applied magnetic field. The slight decrease in the transmission level around this peak is attributed to the small difference between the ferromagnetic volumes present in the topology of Figure 12.28a and b. The defect mode transmission peak does not recover completely (to reach 0 dB) because of the high magnetic losses of $\text{Ni}_{80}\text{Fe}_{20}$ nanowires. If we suppress mathematically those spurious losses, the transmission goes to 0 dB. The position of the defect is expected to have an influence on the frequency response of the structure since the array is finite [14]. The sensitivity of the sharp transmission peak to the defect position can be simply explained as for the dielectric case in Section 12.6.1. When the defect is centered (Figure 12.28b), the structure behaves like two identical finite periodic arrays, behaving like two symmetrical reflectors, creating the same strong value of reflection coefficient at the input and output of the defect section [14]. On the

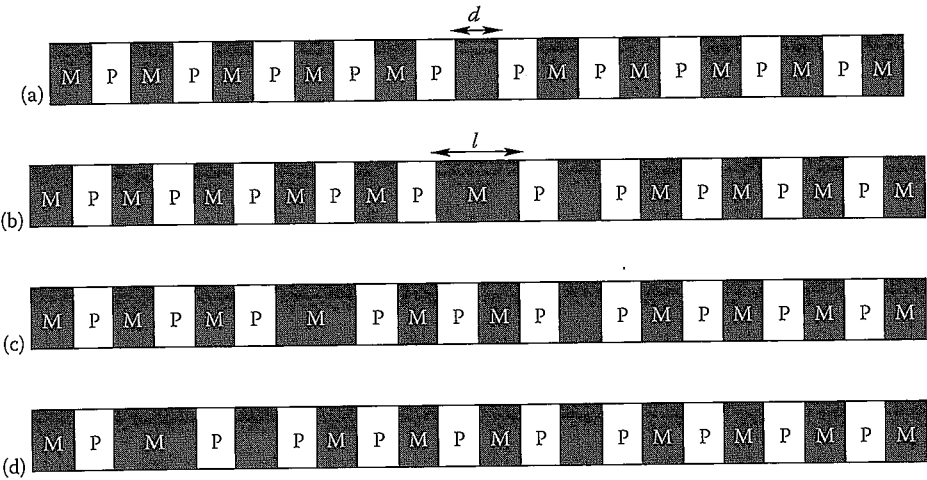


FIGURE 12.28 Topology of MEBG substrate with defect. Dark sections noted “M” represent magnetic sections: polymer with ferromagnetic nanowires. White sections noted “P” represent polymer membrane. (a) shows a top view of MEBG without defect composed of 10 polymer sections and 11 magnetic sections having the same length d . (b), (c), and (d) consist of three MEBGs with defect. They consist of a similar configuration as (a) except for (b) the center, (c) the fourth, and (d) the second, magnetic section that has different length (noted l). (From Saib, A. and Huynen, I., *Electromagnetics*, 26, 261, 2006. With permission.)

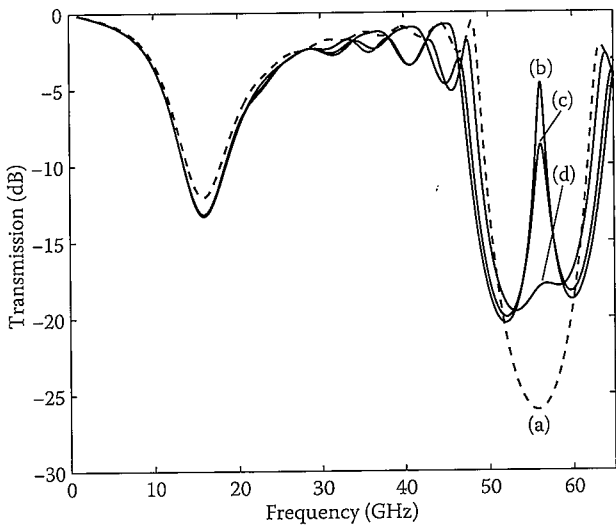


FIGURE 12.29 Simulated transmission coefficient of MSL on the MEBG substrate of the four configurations of Figure 12.28 showing the effect of the defect position in the array. (From Saib, A. and Huynen, I., *Electromagnetics*, 26, 261, 2006. With permission.)

contrary, when the defect is moved to the left (Figure 12.28c and d), the symmetry is broken: the left reflector having smaller number of sections than the right reflector induces a smaller value of reflection level at the input of the defect section than at the output (the propagation is supposed to occur from left to right). This approach can be verified via the simple modeling explained in Section 12.6.1.

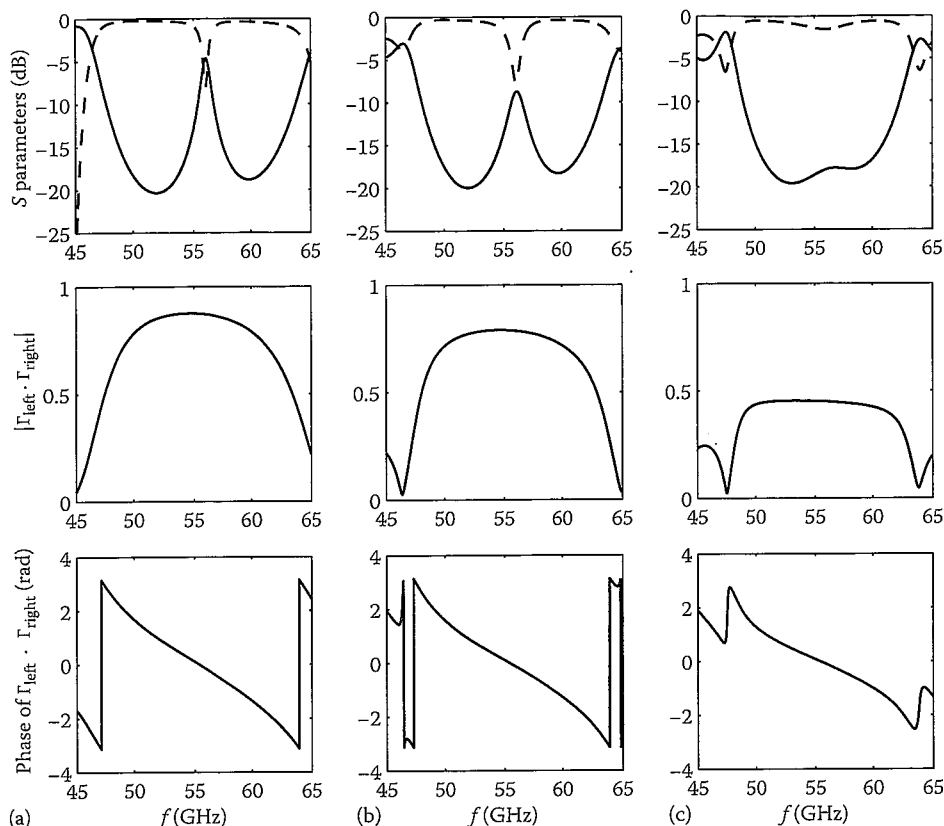


FIGURE 12.30 Calculated reflection (dashed lines) and transmission (solid lines) coefficients (S parameters), obtained using the variational model, and the magnitude and phase of the product of the left and right reflection coefficients ($\Gamma_{\text{left}} \cdot \Gamma_{\text{right}}$) seen from the input and output of the defect section, for each position of the defect: (a) corresponds to Figure 12.28b, (b) to Figure 12.28c, and (c) to Figure 12.28d. (From Saib, A. and Huynen, I., *Electromagnetics*, 26, 261, 2006. With permission.)

Figure 12.30 shows the reflection and transmission coefficients (S parameters), obtained using the variational model, and the magnitude and phase of the product of the left and right reflection coefficients seen at the input of the defect section, for each position of the defect. The resonance condition inside the defect can be expressed by the fact that the left reflection coefficient Γ_{left} must be equal to the inverse of the right reflection coefficient, Γ_{right} . It is observed from Figure 12.30a that for a centered defect at 56.5 GHz, the product $\Gamma_{\text{left}} \cdot \Gamma_{\text{right}}$ is closer to 1 than for the two other positions of the defect. At this frequency, which corresponds exactly to the phase of the product equal to zero, the defect mode appears. The only reason why the transmission does not recover to zero at the defect frequency is that the product magnitude is lower than 1 (0.92 from Figure 12.30a), which is due to the finite dimensions of the two arrays and to the high magnetic losses. When the defect moves to the left (Figure 12.30b and c), the zero phase condition is still fulfilled, but with a lowered magnitude of the product. Therefore, the defect peak transmission is lowered in Figure 12.30b and completely disappears in Figure 12.30c where the product magnitude is less than 0.5. Figure 12.31 shows the effect of the defect length (l) on the position of the defect mode transmission peak for the configuration of Figure 12.28b, where the position of the defect is not changed. By adjusting the length of the defect section, the position of the defect mode transmission peak can sweep across the whole

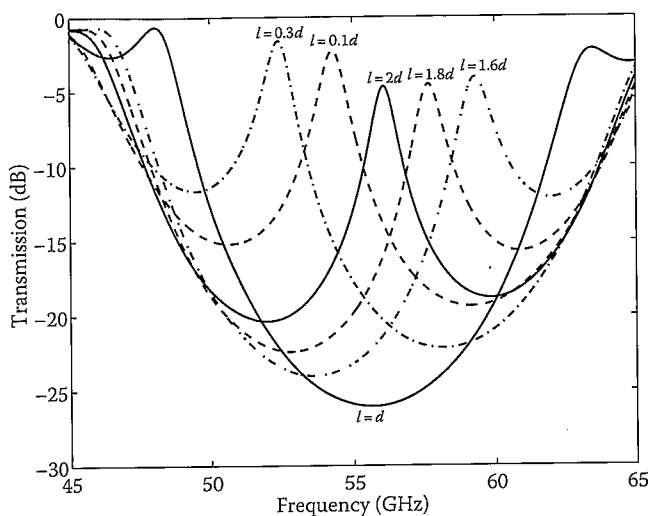


FIGURE 12.31 Effect of the defect length (l) on the position of the defect mode transmission peak for the configuration of Figure 12.28. (From Saib, A. and Huynen, I., *Electromagnetics*, 26, 261, 2006. With permission.)

bandgap. The sharp transmission peak is exactly situated in the middle of the bandgap when the length of the defect section $l = 2d$. When the defect length $l = d$, the structure is equivalent to that without defect and the sharp transmission peak completely disappears. It might be surprising to have a lower transmission level at the defect mode frequency for $l = 2d$, since it is the most favorable for the defect mode as shown in Figure 12.30a. However, in this case, the electrical length of the defect section is the highest, which induces more magnetic losses, so that the transmission level is lowered. Figure 12.32 shows the transmission coefficient corresponding to the configuration of Figure 12.28b

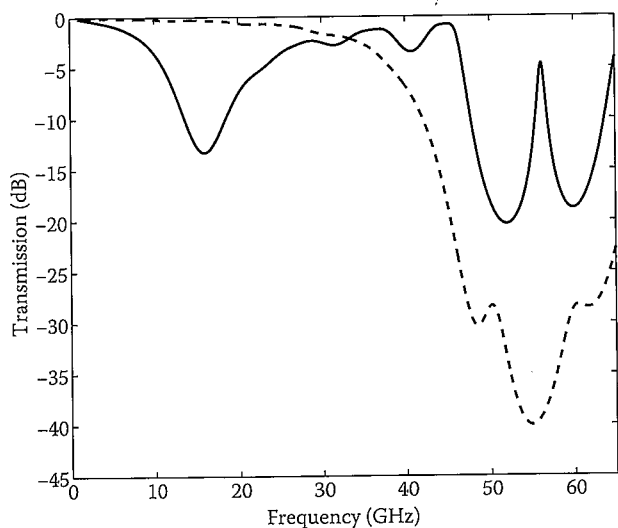


FIGURE 12.32 Transmission coefficient of the configuration of Figure 12.28b when no magnetic field is applied (solid lines), and under a static magnetic field applied parallel to the nanowires (dashed lines). (From Saib, A. and Huynen, I., *Electromagnetics*, 26, 261, 2006. With permission.)

when no magnetic field is applied (solid lines), and under a static magnetic field applied parallel to the nanowires (dashed lines). When an external static magnetic field of 12 kOe is applied to the structure, the FMR peak is shifted toward the defect mode frequency, and thus the defect mode is destroyed. In addition to the fact that the resonance condition for defect mode becomes unsatisfied, since the ferromagnetic properties of the magnetic sections are changed via the magnetic field, the FMR and EBG effects are added to each other, giving a low transmission level of -40 dB. Therefore, the effect of the magnetic field could give tunable properties to the MEBG structure with defect. This can also be used for designing high magnetic field sensors.

12.7 Potential Applications in Magneto-Optics

The TL formalism presented in the previous sections can be easily transposed to the optical range [38]. It is known that materials can present a MO effect, corresponding to dielectric properties in the optical range that vary with an applied magnetic field. The MO effect is modeled by the dual of the ferromagnetic effect presented in Section 12.5. Considering that the material has a spontaneous magnetization along the z -axis or that an external DC magnetic field is applied along this direction. Its dielectric susceptibility tensor has the following expression, where second-order terms are neglected [22,45]:

$$\bar{\bar{\chi}}_r = \begin{bmatrix} \chi_{xx} & \chi_{xy} & 0 \\ \chi_{yx} & \chi_{xx} & 0 \\ 0 & 0 & \chi_{xx} \end{bmatrix}, \quad (12.35)$$

where

$$\chi_{xx} = n_{\text{eff}}^2 - 1 \quad \text{with } n \text{ being the refractive index of ferromagnetic material} \quad (12.36)$$

$$\chi_{xy} = -\chi_{yx} = jaH_{dc} \quad (12.37)$$

$$\alpha = \mu_0 \frac{2c_0 V \sqrt{1 + \chi_{xx}}}{\omega}. \quad (12.38)$$

In Equation 12.38, V is the so-called Verdet coefficient. A nonzero value of this coefficient is responsible for the presence of off-diagonal components in the susceptibility tensor, proportional to the DC magnetic field, H_{dc} , applied along the z -axis (or spontaneous magnetization M_z existing inside the material along this axis). For gases and nonferromagnetic solids, the Verdet coefficient is usually weak, but can be significantly enhanced in the case of ferromagnetic materials [19]. Hence, considering that the MO effect (Equation 12.38) occurs in ferromagnetic nanowires presented in Section 12.2, we apply the same approach as for the FMR modeling equations 12.9 through 12.11 at microwaves to obtain the permittivity tensor in the optical range and resulting scalar permittivity is associated to a MO effect in the ferromagnetic nanowired substrate of Figure 12.21a:

$$\bar{\bar{\epsilon}} = \begin{bmatrix} \epsilon_{xx} & \epsilon_{xy} & 0 \\ \epsilon_{yx} & \epsilon_{xx} & 0 \\ 0 & 0 & \epsilon_{xx} \end{bmatrix} \quad (12.39)$$

$$\epsilon_{xx} = (1 - p) \cdot \epsilon_{rp} + p \cdot (1 + \chi_{xx}) \quad (12.40)$$

$$\epsilon_{xy} = p \cdot \chi_{xy}, \quad \epsilon_{yx} = p \cdot \chi_{yx}, \quad \epsilon_{xz} = \epsilon_{zx} = \epsilon_{yz} = \epsilon_{zy} = 0. \quad (12.41)$$

$$\epsilon_{rMO} = \frac{\epsilon_{xx}^2 + \epsilon_{xy}^2}{\epsilon_{xx}} \quad (12.42)$$

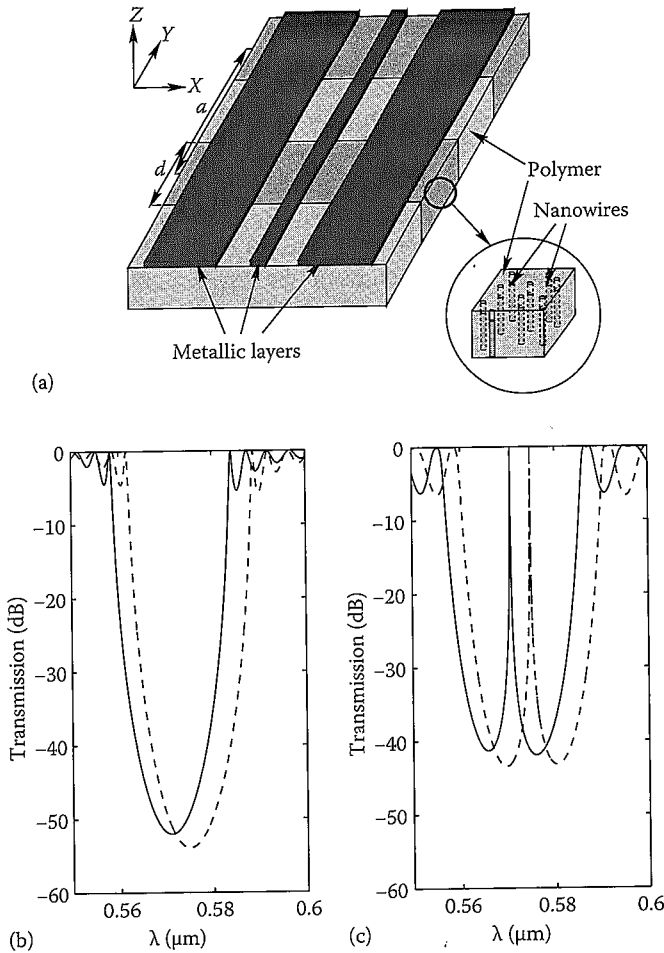


FIGURE 12.33 (a) CPW topology of MEBG for MO configuration, (b) variation of transmission coefficient (S_{21} , solid line) when a 15% variation of the DC magnetization along the z -axis is applied (dashed line), and (c) variation of transmission coefficient in presence of a defect located as in Figure 12.28 with $l = 2d$ (S_{21} , solid line) when a 15% variation of the DC magnetization along the z -axis is applied (dashed line). (From Saib, A. and Huynen, I., *Electromagnetics*, 26, 261, 2006. With permission.)

Figure 12.33 illustrates the concept of MO effect using a MEBG substrate with Co nanowires. The scalar permittivity (Equation 12.42) resulting from the tensor (Equation 12.39) is associated to a wave propagation where the electric field component has to be perpendicular both to the z -axis and to the direction of propagation [19]. Thus, the microstrip configuration used for inducing the MEBG effect associated to FMR at microwaves cannot be used for MO effects, since in that case, the electric field is parallel to the reference z -axis of the tensor (Equation 12.39), which is also the axis of the nanowires. Therefore, we propose another dual configuration for the guiding structure, consisting of a coplanar waveguide (CPW) lying on the periodic MEBG substrate of Figure 12.33a. In this TL, the electric field is present mainly across the slots and directed along the x -axis, hence being perpendicular both to the z -axis and to the y -propagation axis (Figure 12.33a). Equations 12.19 and 12.20 valid for the dielectric EBG are transposed to the case where the dielectric contrast involves the scalar permittivity ϵ_{rMO} (Equation 12.42):

$$Z_{c1} = Z_{cp \text{ MO}} \propto \frac{1}{n_{eff1}}, \quad n_{eff1} = \sqrt{\epsilon_{rp}}, \quad \gamma_1 = j\beta_1 = jn_{eff1}k_0 \quad (12.43)$$

$$Z_{c2} = Z_{c \text{ mag MO}} \propto \frac{1}{n_{eff2}}, \quad n_{eff2} = \sqrt{\epsilon_{r \text{ MO}}}, \quad \gamma_2 = j\beta_2 = jn_{eff2}k_0, \quad (12.44)$$

where $Z_{c \text{ poly MO}}$ and $Z_{c \text{ mag MO}}$ are characteristic impedances at optical wavelengths for the polymer and nanowired microstrip sections, respectively, with $\epsilon_{r \text{ MO}}$ given by Equation 12.42. Applying the TL chain formalism already used in the previous sections, Figure 12.33b shows that the maximum of bandgap attenuation generated in this structure can be tuned at optical wavelengths from $\lambda_0 = 570$ to 575 nm when the applied magnetic field or spontaneous magnetization along the z -axis is varied by 15%. Adding a defect centered in the periodicity as in Figure 12.28a induces as expected a sharp transmission peak in the bandgap for a signal wavelength λ equal to 570 nm (Figure 12.33c). Varying by 15% the magnetization or applied DC field shifts the peak to $\lambda = 575$ nm, so that at the initial wavelength of 570 nm, the attenuation becomes maximal. Controlling the structure using a small variation of the DC magnetic state of the ferromagnetic part of the EBG substrate enables thus to switch on or off the sharp transmission peak at 570 nm.

12.8 Conclusions

In this chapter, we have illustrated the potential of a novel nanocomposite material consisting of porous polymer or alumina membrane containing ferromagnetic nanowires. As for dielectric EBG substrates, we presented a TL formalism for planar structures that exploits this material as a microwave substrate. Planar substrates having a scope for this analysis are based on the topology of Bragg reflectors used in the optical frequency range: however, they include TLs as guiding structures, so that propagation inside does not consist of free space mechanism. Both dielectric EBG and MEBG substrates are investigated. It is shown that a MEBG structure based on ferromagnetic nanowires and coupled to defect modes can offer promising tuning possibilities for filter design, operating either at microwaves based on the FMR effect, or in the optical range, based on the tunable MO effect induced on the optical dielectric properties of ferromagnetic nanowires. Another application is the realization of SNG and DNG metamaterials exploiting the negative permeability induced by the nanowired substrate. Devices using nanowired composite substrates were investigated, based either on their ferromagnetic behavior for designing nonreciprocal devices [29,46,47], or for filtering applications combining their tunable FMR absorption effect with MEBG effects [40]. As mentioned in [15,19,21], the main challenge for designing efficient devices based on magnetic materials is the reduction of losses: for microwave applications using the nanowired substrate, solutions currently investigated consist of replacing the PC polymer template by other porous templates having lower dielectric losses as alumina. At optical wavelengths, the MO effect of interest has to be used in a wavelength window where the absorbance of the material remains moderate. Nanoscaled ferromagnetic materials could be good candidates.

References

1. J.B. Pendry, Magnetism from conductors and enhanced nonlinear phenomena, *IEEE Trans. Microwave Theory Tech.*, 47, 2075–2084, 1999.
2. S. Maslovski, S. Tretyakov, and P. Belov, Wire media with negative effective permittivity: A quasi-static model, *Microwave Opt. Technol. Lett.*, 35(1), 4–51, October 2002.
3. M.A.G. Laso, M.J. Erro, D. Benito, M.J. Garde, T. Lopetegi, F. Falcone, and M. Sorolla, Analysis and design of 1-d photonic bandgap microstrip structures using a fiber grating model, *Microwave Opt. Technol. Lett.*, 22(4), 223–226, August 1999.

4. E. Yablonovitch, Inhibited spontaneous emission in solid-state physics and electronics, *Phys. Rev. Lett.*, 58(20), 2059, May 1987.
5. K.M. Ho, C.T. Chang, and C.M. Soukoulis, Existence of a photonic gap in periodic dielectric structures, *Phys. Rev. Lett.*, 65, 3152–3155, 1990.
6. J.B. Pendry, Negative refraction makes a perfect lens, *Phys. Rev. Lett.*, 85(18), 3966, 2000.
7. K.M. Ho, C.T. Chang, and C.M. Soukoulis, Existence of a photonic gap in periodic dielectric structures, *Phys. Rev. B*, 65, 3152–3155, 1990.
8. N.C. Karmakar and M.N. Mollah, Investigations into nonuniform photonic-bandgap microstripline low-pass filters, *IEEE Trans. Microwave Theory Tech.*, 51, 564–572, 2003.
9. V. Radisic, Y. Qian, R. Coccioli, and T. Itoh, Novel 2-d photonic bandgap structure for microstrip lines, *IEEE Microwave Guided Wave Lett.*, 8(2), 69–71, February 1998.
10. I. Rumsey, M. Piket-May, and P. K. Kelly, Photonic bandgap structures used as filters in microstrip circuits, *IEEE Microwave Guided Wave Lett.*, 8, 336–338, 1998.
11. C.-S. Kee, J.Y. Park, S.J. Kim, H.C. Song, Y.S. Kwon, N.H. Myung, S.Y. Shin, and H. Lim, Essential parameter in the formation of photonic band gaps, *Phys. Rev. E*, 59(4), 4695, April 1999.
12. C.-S. Kee, J.-E. Park, and H. Lim, Roles of wave impedance and refractive index in photonic crystals with magnetic and dielectric properties, *IEEE Trans. Microwave Theory Tech.*, 47(11), 2148, November 1999.
13. M.K. Tiwari, K.K. Gupta, H.C. Gupta, and D.C. Dube, Photonic bandgap studies on periodic metallic structures in the microwave region, *Microwave Opt. Technol. Lett.*, 35(6), 483, December 2002.
14. A. Saib, R. Platteborze, and I. Huynen, Experimental demonstration of the origin of photonic bandgap creation and associated defect modes in microwave planar circuits, *Microwave Opt. Technol. Lett.*, 41(1), 5–9, April 2004.
15. A. Saib, D. Vanhoenacker, I. Huynen, A. Encinas, L. Piroux, E. Ferain, and R. Legras, A magnetic photonic bandgap material at microwave frequencies based on ferromagnetic nanowires, *Appl. Phys. Lett.*, 83(12), 2378–2380, September 2003.
16. R.D. Meade, K.D. Brommer, A.M. Rappe, and J.D. Joannopoulos, Photonic bound states in periodic dielectric materials, *Phys. Rev. B*, 44, 13772–13774, 1991.
17. M.M. Sigalas, C.M. Soukoulis, R. Biswas, and K.H. Ho, Effect of the magnetic permeability on photonic bandgaps, *Phys. Rev. B*, 56(3), 959–962, July 1997.
18. E. Yablonovitch, T.J. Gmitter, R.D. Meade, A.M. Rappe, K.D. Brommer, and J.D. Joannopoulos, Donor and acceptor models in photonic band structure, *Phys. Rev. Lett.*, 59, 4695–4698, 1999.
19. M.J. Freiser, A survey of magneto-optic effects, *IEEE Trans. Magn.*, 4, 152–161, 1968.
20. D. Goldring, Z. Zalevsky, G. Shabtay, D. Abraham, and D. Mendlovic, Magneto-optic-based devices for polarisation control, *J. Opt. A*, 6, 98–105, 2004.
21. R. Cobian, L.J. Cruz Riveira, N. Kim, S.Y. Sung, and X. Qi, Magneto-optic materials for integrated photonics, *IEEE LEOS Newsl.*, 17(2), 3–5, April 2003.
22. I.L. Lyubchanskii, N.N. Dadoenkova, M.I. Lyubchanskii, E.A. Shapovalov, and T. Rasing, Magnetic photonic crystals, *J. Phys. D*, 36, R277–R287, 2003.
23. M. Inoue, K. Arai, T. Fuji, and M. Abe, One-dimensional magnetophotonic crystals, *J. Appl. Phys.*, 85, 5768–5779, 2003.
24. I. Huynen, G. Goglio, D. Vanhoenacker, and A.V. Vorst, A novel nanostructured microstrip device for tunable filtering applications at microwaves, *IEEE Microwave Guided Wave Lett.*, 9(10), 401–403, October 1999.
25. J. Spiegel, J. de la Torre, M. Darques, L. Piroux, and I. Huynen, Permittivity model for ferromagnetic nanowired substrates, *IEEE Microwave Wireless Component Lett.*, 7, 492–494, 2007.
26. J.B. Pendry, A.J. Holden, W.J. Stewart, and I. Youngs, Extremely low frequency plasmons in metallic mesostructures, *Phys. Rev. Lett.*, 76(25), 4773–4776, 1996.

27. J. Spiegel and I. Huynen, Microwave properties of ferromagnetic nanowires and applications to tunable devices, *Solid State Phenomena*, 152–153, 389–392, 2009.
28. V. Boucher and D. Ménard, Electromagnetic properties of ferromagnetic nanowire arrays. *J. Appl. Phys.*, 103, 07E720–07E720-3, 2008.
29. A. Saib, M. Darques, L. Piraux, D. Vanhoenacker-Janvier, and I. Huynen, Design of a unbiased microwave circulator using a magnetic nanowired substrate, *IEEE Trans. Microwave Theory Tech.*, 53(6), 2043–2049, June 2005.
30. A. Encinas, M. Demand, L. Vila, L. Piraux, and I. Huynen, Tunable remanent state resonance frequency in arrays of magnetic nanowires, *Appl. Phys. Lett.*, 81(11), 2032–2034, September 2002.
31. A. Berk, Variational principles for electromagnetic resonators and waveguides, *IEEE Trans. Antennas Propag.*, AP-4, 104–111, April 1956.
32. B. Lax and K.J. Button, *Microwave Ferrits and Ferrimagnetics*, McGraw-Hill, New York, 1962.
33. C. Caloz and T. Itoh, *Electromagnetic Metamaterials: Transmission Line Theory and Microwave Applications*, Wiley Interscience, New York, 2006.
34. S. Shingubara, Fabrication of nanomaterials using porous alumina templates, *J. Nanopart. Res.*, 5, 17–30, 2003.
35. G. Matthei, L. Young, and E. Jones, *Microwave Filters, Impedance Matching Networks and Coupling Structures*, McGraw-Hill, New York, 1964.
36. J. Spiegel, S. Eggermont, and I. Huynen, Investigation on ferromagnetic nanowired substrates for leaky-wave antennas, *Proceedings of Metamaterials*, 2008.
37. A. Encinas-Oropesa, M. Demand, I. Huynen, U. Ebels, and L. Piraux, Dipolar interactions in arrays of nickel nanowires studied by ferromagnetic resonance, *Phys. Rev. B*, 63, 104415, 2001.
38. A. Saib and I. Huynen, Transmission lines on periodic bandgap metamaterials: From microwaves to optics applications, *J. Opt. A: Pure Appl. Opt.*, 7(2), S124–S132, February 2005.
39. A. Saib and I. Huynen, Periodic metamaterials combining ferromagnetic nanowires and dielectric structures for planar circuits applications, *Electromagnetics*, 26, 261–277, 2006.
40. A. Saib, I. Huynen, and D. Vanhoenacker-Janvier, Design of a stopband filter based on a magnetic photonic bandgap material, in *Proceedings of the 33th European Microwave Conference*, Munich, Germany, October 2003, pp. 809–812.
41. A. Dobrowolski, *Introduction to Computer Methods for Microwave Circuits Analysis and Design*, Artech House, Boston London, 1991.
42. L. Piraux, S. Dubois, and S. Demoustier-Champagne, Template synthesis of nanoscale materials using the membrane porosity. *Nucl. Instrum. Methods Phys. Res. B*, 131, 357–363, 1997.
43. A. Encinas-Oropesa, M. Demand, U. Ebels, I. Huynen, and L. Piraux, Effect of dipolar interaction on the ferromagnetic resonance properties in arrays of magnetic nanowires, *J. Appl. Phys.*, 89(11), 6704, June 2001.
44. A. Saib and I. Huynen, Defect modes in magnetic photonic band-gap materials based on ferromagnetic nanowires, in *Proceedings of the Bianisotropics 2004 Conference*, Ghent, Belgium, September 22–24, 2004, p. 4.
45. C.Y. You nad S.C. Shin, Derivation of simplified analytic formulae for magneto-optical Kerr effects, *Appl. Phys. Lett.*, 69, 1315–1317, 1996.
46. A. Saib, M. Darques, L. Piraux, D. Vanhoenacker-Janvier, and I. Huynen, Design of a unbiased microwave circulator using a magnetic nanowired substrate, in *Proceedings of the 34th European Microwave Conference (EuMC04)*, Amsterdam, the Netherlands, October 2004, pp. 1353–1356.
47. A. Saib, D. Vanhoenacker, I. Huynen, M. Darques, and L. Piraux, Unbiased microwave circulator based on ferromagnetic nanowires arrays of tunable magnetization state, *J. Phys. D: Appl. Phys.*, 28(16), 2759–2763, August 2005.