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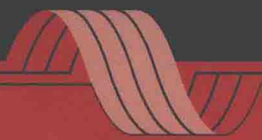
Low-Profile Natural and Metamaterial Antennas



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Low-Profile Natural and Metamaterial Antennas

Analysis Methods and Applications

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Preface

This book is written for antenna engineers, researchers, graduate students, and advanced undergraduate students, who need to realize low-profile antennas that are required for modern communication systems, such as mobile and earth-satellite communications systems.

Low-profile antennas have been an ongoing topic of interest in the antenna field for over four decades. Their inherent characteristics of compactness, low aerodynamic resistance, and durability make them attractive for use in advanced/smart base station antennas, vehicular antennas, satellite communication antennas, radar antennas, and the like. As these systems advance, the need for smaller, more robust low-profile designs to meet ever-stricter system requirements has driven the field of low-profile antenna design forward. With this trend has come the need for new antenna design references; this book is intended to help meet this need.

Normally, electromagnetic properties in nature are right-handed. Antennas with right-handed properties are designated as natural antennas. On the other hand, antennas having electromagnetic properties that are not found in naturally occurring materials are designated as metamaterial-based antennas (simply referred to as metamaterial antennas). Note that metamaterials are realized artificially, typically by arraying small conducting elements periodically in one, two, or three dimensions. These artificial materials are generally referred to as left-handed materials, because they obey the left-hand rule, as opposed to the right-hand rule observed for natural materials. This unique property leads to antenna implementations that otherwise could not be realized, including a number of low-profile antenna implementations.

Designing a low-profile antenna brings with it a unique challenge—overcoming the inherent tendency for antenna performance to decrease as the height is decreased. This book covers recent progress in low-profile natural and metamaterial antennas and is composed of three parts: Introduction (Part I), Low-Profile Natural Antennas (Part II), and Low-Profile Metamaterial-Based Antennas (Part III).

Part I has three chapters (Chapters 1–3). Chapter 1 defines the natural and metamaterial-based antennas to be covered in this book, based on the *propagation phase constant* of the current along the radiation element. Chapters 2 and 3 provide the fundamentals for readers to be able to write computer programs based on the method of moments (MoM) (Chapter 2) and the finite-difference time-domain method (FDTDM) (Chapter 3). For this, in Chapter 2, a series of integral equations (*N*-series integral equations of N1–N5) for the MoM are discussed. It is noted that an arbitrarily shaped conductor (of one, two, or three dimensions) is modeled as an aggregate of conducting wire cells (subdivided elements), and the periphery of each cell is approximated by straight wires or curved wires, to which the *N*-series integral equations are applied. In Chapter 3, the basic FDTDM is summarized and the locally one-dimensional (LOD) FDTDM is presented.

Part II is composed of 15 chapters (Chapters 4–18) and discusses low-profile natural antennas that are classified into four groups: base station antennas (Part II-1: Chapters 4–7), card antennas for mobile equipment (Part II-2: Chapters 8–10), beamforming antennas (Part II-3: Chapters 11–15), and earth–satellite and satellite–satellite communications antennas (Part II-4: Chapters 16–18).

The first part of low-profile natural antennas covers base station antennas (II-1), where wideband operation for inverted-F antennas and multiband operation for multiloop antennas are discussed. In addition, the realization of ultra-wideband (UWB) operation for a fan-shaped antenna and of a body-of-revolution antenna with a shorted parasitic ring (BOR–SPR) is described.

The second part (II-2) presents topics on low-profile card antennas, including the realization of multiband operation for inverted-LFL card antennas, together with the realization of UWB operation for fan-shaped card antennas and planar monopole card antennas. These card antennas are designed to fit into the limited space in mobile equipment, such as portable telephones and personal computers.

The beamforming antenna discussion in Part II-3 starts with the realization of a reconfigurable antenna, where inverted-F elements above an electromagnetic bandgap reflector are used. Next, reconfigurability for a bent two-leaf (BeToL) antenna and a bent four-leaf (BeFoL) antenna is discussed. Using switching circuits, these antennas can be reconfigured to radiate a beam in one of several directions at a specific frequency, while maintaining the same radiation characteristics. It is emphasized that the horizontal area of the BeToL and BeFoL antennas is much smaller than that of corresponding reconfigurable antennas that use patches.

The discussion is extended to beamforming based on the Fabry–Pérot principle. It is shown that a single low-gain feed patch forms a high-gain tilted LP beam in a specific direction, by placing a parasitic layer consisting of loop elements above the feed patch. This antenna structure is simple and differs from conventional tilted-beam array antenna structures, where the arrayed radiation elements are connected to a main feed source by transmission lines through phase shifters and attenuators to form the tilted beam.

The discussion in II-3 finishes with two grid array antennas: one is a linearly polarized (LP) rhombic grid array antenna and the other is a circularly polarized (CP) loop grid array antenna. These antennas have a radiation beam that scans from the broadside direction to the forward direction, with a high gain for both LP and CP waves.

In LP-wave communication systems, the receiving antenna must be aligned with the polarization direction of the transmitting antenna; for instance, if a vertically polarized antenna is used for a transmitting antenna, then a vertically polarized receiving antenna is needed to maximize the reception of the transmitted power. In other words, the transmitting and receiving antennas need to be aligned such that the polarization directions are the same. On the other hand, a communication system where a CP wave is used does not need such alignment; for example, if a right-handed CP antenna is used as a transmitting antenna, then it is enough to point a right-handed CP antenna at the transmitting antenna.

It is from this point of view that the fourth part, II-4, which discusses earth–satellite and satellite–satellite communications antennas, treats low-profile CP antennas. Arrays of spiral, helical, and curl antennas that obtain very high aperture efficiency are presented. In addition, a low-profile composite spiral and helical antenna array, which forms a high-gain tilted beam that can be aimed toward a satellite, is discussed.

Part III is composed of five chapters (Chapters 19–23) and discusses low-profile metamaterial-based antennas. Chapter 19 reveals that a metamaterial-based straight-line antenna (metaline antenna) radiates an LP backward beam, which cannot be achieved with a corresponding natural straight-line antenna having a right-handed property. Chapter 20 presents a new finding that a metamaterial-based loop antenna (metaloop) radiates single- and dual-peak LP beams at frequencies below a specific frequency (transition frequency) in addition to radiating the same beams at frequencies above the transition frequency.

Subsequently, Chapter 21 presents an open metaloop antenna that radiates a left-handed CP beam across a specific frequency band and a right-handed CP beam across a different frequency band (dual-band counter CP radiation). It is noted that such dual-band counter CP radiation cannot be obtained using a corresponding natural loop antenna that has a fixed single feed point. This is also true for both *natural* spiral and *natural* helical antennas.

However, the metamaterial-based spiral (metaspiral) antenna presented in Chapter 22 and the metamaterial-based helical (metahelical) antenna presented in Chapter 23 are shown to create dual-band counter CP radiation, which solves this issue. It should be emphasized that the antenna height for the metaspiral antenna is approximately 1/100 of the wavelength at the lowest operating frequency, in contrast to the 1/4 wavelength antenna height of conventional antennas backed by a conducting plate (reflector).

It is hoped that the antennas presented in this book will give readers supplemental knowledge that will be useful for practical antenna applications.

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Contents

Part I

Introduction

Chapter 1

Categorization of Natural Materials and Metamaterials

An electromagnetic material is categorized by its constitutional parameters, permittivity ε and permeability μ. A double-positive (DPS) material ($\varepsilon > 0$ and $\mu > 0$) is defined as a right-handed (RH) material. The phase constant of wave propagation within the RH material exhibits a positive value ($\beta > 0$). A double-negative (DNG) material ($\varepsilon < 0$ and $\mu < 0$) is defined as a left-handed (LH) material. The phase constant of wave propagation within the LH material exhibits a negative value ($\beta < 0$). Note that β within a mu-negative (MNG) material ($\varepsilon > 0$ and $\mu < 0$) and an epsilon-negative (ENG) material ($\varepsilon < 0$ and $\mu > 0$) is zero (i.e., evanescent).

A DPS material is a material found easily in nature and called a natural material, while a DNG, MNG, or ENG material is an artificial material and called a metamaterial (MTM) [1].

1.1 NATURAL AND METAMATERIAL ANTENNAS DISCUSSED IN THIS BOOK

Most antennas are made of natural materials. Antennas based on metamaterials are new, and some examples are found in Refs [1–3]. The categorization of natural and metamaterial antennas presented *in this book* is in reference to β, the propagation phase constant of the *current* flowing along a *fed* element.

Figure 1.1a shows a fed antenna where the out-going current from the feed point F toward the antenna element ends flows with a positive phase constant ($\beta > 0$). This means that the phase distribution takes a regressive form, that is, the phase is delayed from point F toward the antenna element ends. This type of antenna is categorized as a *natural* (NTR) *antenna*.

Figure 1.1b shows a fed antenna where the propagation phase constant of the out-going current can be either negative within a specific frequency band ($\beta < 0$) or zero at a nonzero frequency ($\beta = 0$). This type of antenna is categorized as a *metamaterial-based antenna* (simply referred to as a metamaterial antenna). The phase

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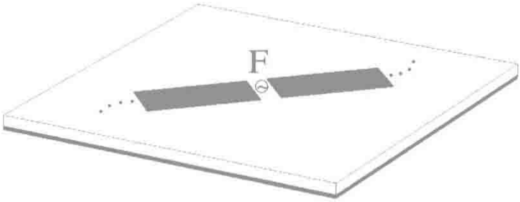
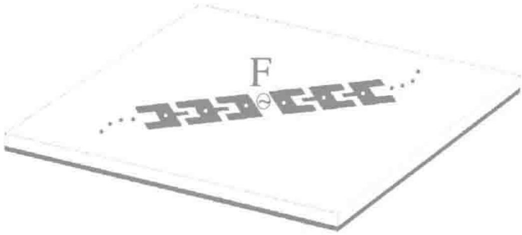


(a)



(b)

Figure 1.1 Antenna definition. (a) Natural antenna. The propagation phase constant is positive. (b) Metamaterial-based antenna (metamaterial antenna). The propagation phase constant is either negative within a specific frequency band or zero at a nonzero specific frequency.

distribution for $\beta < 0$ takes a progressive form from point F to the antenna element ends. A phase constant of zero ($\beta = 0$) means that the wavelength is infinitely long.

Exercise

Figure 1.2a shows a plane wave traveling within a lossless medium (permittivity ε and permeability μ) in the z-direction. Figure 1.2b shows a lossless transmission line characterized by distributed circuit parameters [$C'(F/m)$,$L'(H/m)$,$C'_Z(F \bullet m)$, and $L'_Y(H \bullet m)$]. Discuss the correspondence between the medium constitutional parameters (ε and μ) and the circuit parameters.

Answer Plane wave propagation within a lossless medium is specified by the following characteristic impedance Z_C and propagation constant γ:

$$Z_C = \sqrt{\frac{\mu}{\epsilon}} \tag{1.1}$$

$$\gamma = j\omega\sqrt{\mu\epsilon} \tag{1.2}$$

The propagation of voltage and current in a lossless transmission line is specified by the following characteristic impedance Z_{MTM} and propagation constant γ_{MTM}:

$$Z_{MTM} = \sqrt{\frac{Z'}{Y'}} \tag{1.3}$$

$$\gamma_{MTM} = \sqrt{Z'\,Y'} \tag{1.4}$$

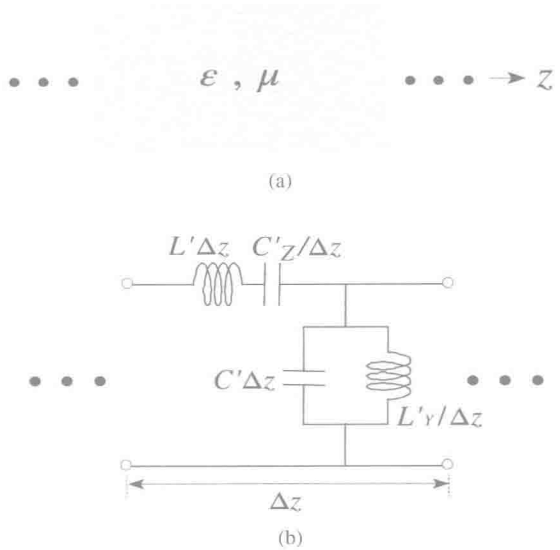


Figure 1.2 Equivalence. (a) Plane wave propagation within a lossless medium. (b) Lossless transmission line.

where

$$Z' = j\omega\left(L' - \frac{1}{\omega^2 C'_Z}\right) \equiv j\omega\,\mu_{TL} \tag{1.5}$$

$$Y' = j\omega\left(C' - \frac{1}{\omega^2 L'_Y}\right) \equiv j\omega\,\epsilon_{TL} \tag{1.6}$$

Then, Eqs. (1.3) and (1.4) are given by

$$Z_{MTM} = \sqrt{\frac{\mu_{TL}}{\epsilon_{TL}}} \tag{1.7}$$

$$\gamma_{MTM} = j\omega\sqrt{\mu_{TL}\epsilon_{TL}} \tag{1.8}$$

It is concluded that the medium constitutional parameters μ and ε correspond to the circuit parameters μ_{TL} and ϵ_{TL}, respectively,

$$\mu = \mu_{TL} = L' - \frac{1}{\omega^2 C'_Z} \tag{1.9}$$

$$\epsilon = \epsilon_{TL} = C' - \frac{1}{\omega^2 L'_Y} \tag{1.10}$$

Note that μ_{TL} and ϵ_{TL} can both be negative across a specific frequency region. In such a situation (simultaneously $\mu_{TL} < 0$ and $\epsilon_{TL} < 0$), the transmission line in Fig. 1.2b has a negative phase constant of $\beta = -\omega\sqrt{|\epsilon_{TL}||\mu_{TL}|}$. ■