Design of miniaturized monopole antennas fed by gap source

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DESIGN OF MINIATURIZED MONOPOLE ANTENNAS FED BY GAP SOURCE

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A thesis submitted to Nanyang Technological University
in fulfillment of the requirements for the degree of
Master of Engineering

2009

STATEMENT OF ORIGINALITY

I hereby certify that the work embodied in this thesis is the result of original research and has not been submitted for a higher degree to any other University or Institution.

10 October 2009

Date

Alireza Akbarzadeh

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SUMMARY

Antenna miniaturization is one of the most challenging tasks that antenna designers are encountering nowadays. Having a versatile electrically small antenna with a satisfactory radiation efficiency along with other components in a transceiver circuit has engendered a necessity for antenna designers to deal with this task prominently. Antenna parameters (e.g. radiated power, frequency bandwidth and radiation efficiency) are dependant upon its physical dimensions. Reducing the physical size of an antenna degrades its performance in a great deal. Fundamental limits have been determined as upper bounds on the performance of an electrically small antenna and it has been proved that one can slightly approach these limits in an ideal case.

All the proposed methods to overcome the prevailing challenges in antenna miniaturization are divided into two general ways: 1) using novel structures, 2) loading the antenna with materials. The idea behind the first method is to increase the effective volume of the antenna while its physical volume is unchanged. Increasing the effective volume of the antenna, its current distribution would be improved and the near field storage would be reduced. In the second method, it is aimed to load the antenna with dielectric materials to reduce its resonance frequencies. If we assume an antenna as a resonator and load it with a dielectric material properly, we can create a dielectric resonator with improved resonance properties over the unloaded resonator (antenna). Different types of materials such as magneto-dielectrics, meta-materials, anisotropic materials and electromagnetic band gap materials can be invoked to load the antenna for different applications.

In this thesis, I am to study both the miniaturization techniques. Specifically, I am going to load a vertical monopole with dielectric materials. The monopole is fed with a small gap source over an infinitely large ground plane. At the first step, the gap source model will be examined against the coaxial feed model for monopole antennas with the use of the modal expansion method. Two approximate closed-form expressions for the input admittance of

the monopole and also the optimum values for the gap-height will be given. Numerical results will be provided to show the equivalence between two models.

In the remaining part of this thesis, I am going to compare meta-materials with usual dielectric materials in monopole antenna miniaturization. Employing the eigenmode analysis, I can obtain the radiation parameters of the monopole loaded with meta-materials and normal dielectrics. Then I will give different results and distinguish their pitfalls and benefits. I am also going to examine the structure bending method (first miniaturization method) combined with material loading to achieve a highly miniaturized antenna. Hence, I will study the material loaded top-hat monopole antenna and provide various results to demonstrate how antenna elements and material parameters can affect the radiation performance. Finally, conclusions and recommendations for future work will come at the end.

CHAPTER 1

Introduction

1.1 Motivation and Background

Centuries before, man was busy with gods and religious affairs. Every thing was in the handle of churches and he had no freedom to seek for unknowns in nature. Every idea came from holy books and holy fathers were the only scientists in that time. Affluence was to be found in ecclesiastical matters and man had no safe haven but church. As a consequence, emperors approached churches to justify their governments and to guarantee their territories. It was the time of abuse! Little by little, man found out that ecclesiastical affairs were not able to satisfy all of his instincts and also he could not obtain every thing in churches. So man came down from sky to earth four centuries ago and tried to extract all his needs from nature on the basis of his own endeavor and respond to his endless curiosity. He stood against churches' kingdom to control his life by himself and dispose of any corruption or penetration. He conquered churches and empires and revolutionized his lifestyle. He became the center of universe and rushed to broaden his horizons in science and also to bring more comfort to his life. It was the birth of a career named "modernity". This effort has been carried on in so many different styles till now. In a word, modernity is nothing but a revolution in ideologies, techniques and instruments. These three aspects are fruits of reformation in man's thinking and also his drastic development in science.

Ultra fast jets, huge spacecrafts, guided missiles, nuclear power stations, speedy trains, cellular hand phones, computer networks, structural biology, optical fibers, nanotechnologies, all and all, have come to existence in the shadow of modernity. Intuitively, modernity and all it brings behind alter human's society and the ways he communicates with. It means that modernity

Chapter 1

Introduction

has increased man's demands. Satisfying these demands, he needs to communicate with all over the world in an efficient way to transfer a huge amount of information. It seems very unlikely for him to live an easy life without a professional communication system which is able to handle data transformation in a great deal. On the other hand, modernity has made the earth small for man and he is thinking of aiming at other points of the universe. As a result, he is supposed to enhance his communication systems.

At the present time, people are going on the edge of technology and are hungrier than before for novel and versatile devices. Rapid changes in consumers' interests are blowing up, which lead to high volatility of technology marketing. On the other hand, various protocols are emerging everyday to enhance the quality of computer networks. New standards are coming out to increase the data rate in Wireless Local Area Networks (WLAN). Modern packaging technologies are applied to reduce the size of communication circuits. Modified semiconductors and other materials along with reliable Integrated Circuit (IC) design methods are used to increase the fidelity of transceivers. New generation of Digital Signal Processors (DSP) and communicative microcontrollers are coming to existence to provide a more compatible communication system for humankind. Complex coding and cryptography systems are released to improve the security of data transformation. Quantum computation theories are being studied to break the data rate limits in networks.

Shown in Fig. 1.1 are the general block diagrams of a transmitter and a receiver in a communication system. It is seen that every communication system consists of various parts like amplifiers, filters, modulators, demodulators, mixers and oscillators. Recent technologies and researches in radio frequency circuit design and also in optics have done great jobs in the dramatic miniaturization of such elements with keeping their efficiencies. Man's achievements in lithography and ultra high frequency IC design are

examples of such capabilities in acquiring highly miniaturized circuits and systems.

In transceiver circuits, antenna is one of the largest elements, which possesses a very crucial role in converting electric energy into electromagnetic energy. Antenna dimensions are inversely dependant on its operating frequency. As a consequence, antenna size reduction deteriorates its radiation efficiency and also its frequency bandwidth. As I mentioned before, a modern transceiver is supposed to have an efficient and packed antenna. Since size reduction in antenna causes poor radiation efficiency, a novel technique must be utilized to design an efficient but electrically small antenna (ESA). This novel technique is called “antenna miniaturization”.

Fig. 1.1. Block diagram of a typical communication system

As I pointed out earlier, so many highly developed ideas are coming along with various parts of a transceiver. But it seems that antenna miniaturization is the most challenging part in a transceiver design. Thus an efficient miniaturization technique must be utilized to design an ESA, which is operating over a desired frequency band (wideband or multi-band) with an acceptable throughput.

1.2 Objectives

My major objective in this dissertation is to study the antenna miniaturization process. Nowadays, there exists a great tendency to miniaturize

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practical antennas in a proficient way along with other miniaturized elements in transceiver circuits. But antenna performance is directly related to its physical dimensions. Reducing the size of an antenna depletes its radiation efficiency and also deteriorates its radiation performance. Researchers have investigated antenna miniaturization from different points of view for decades and determined several fundamental limitations on the ESA performances to which we can only approach in the most ideal case.

Generally speaking, two main miniaturization techniques are proposed: 1) using novel structures, 2) loading the antenna with materials. In this thesis, I am concerned with implementing both methods, though my main emphasis is on the usage of materials in antenna miniaturization. Specifically, the so-called meta-materials (MTM), which have attracted much attention nowadays, can be invoked to load an ESA. Theoretically speaking, juxtaposition of materials having negative permittivity and/or negative permeability (MTMs) with usual dielectric materials leads to interesting phenomena due to symmetry and reciprocity of Maxwell's equations. The physics behind this comes from the resonances taking place at the interface between the materials. Later I shall deal with such phenomena in more details and try to explain how the effective use of anomalous MTMs would engender antenna size reduction.

As a matter of fact, I am motivated to study the effect of MTMs on the radiation properties of a wire monopole antenna in an analytical way to examine the capability of MTMs in comparison with normal dielectric materials in antenna size reduction. So the classical problem of the material coated monopole over an infinite ground plane must be solved and the radiation properties of the structure such as radiation pattern, input admittance and current distribution along the antenna must be obtained. Then considering the constitutive parameters of the coating, the effects of material coating on the monopole performance should be examined and benefits along with shortcomings of different materials should be notified.

It seems more advantageous to combine both the mentioned miniaturization methods. So another objective of this dissertation is to make a change in the topology of the monopole along with its material loading. A detailed study should be done on different parameters and several guidelines and compromises should be suggested in order to miniaturize the monopole antenna.

1.3 Contributions

Several contributions from this dissertation can be mentioned. Two approximate closed-form expressions for the input admittance of the monopole over an infinite ground plane driven by the coaxial feed line and the gap source model are presented, respectively. A good estimation of the gap height is presented as well. To my knowledge this estimation has been done for the first time. Additionally, a comprehensive comparison between the anomalous MTMs and normal dielectric materials (high permittivity and/or high permeability) are made and their challenges and superiorities in antenna miniaturization are presented in an analytical way. In fact, it is indicated that coating a normal monopole or loading a top-hat monopole with the MTMs or usual dielectric materials can lead to a significant antenna size reduction, though the bandwidth of operation would be so narrow and the radiation efficiency drop due to the conductive loss, is to some extent high. Some guidelines for the design of the miniaturized material loaded monopole antenna are also provided in this thesis.

1.4 Organization of the Thesis

Design of an efficient ESA with a flexible bandwidth (wide/narrow/multi-band) is one of the most important tasks for antenna designers and has been debated for decades. Fundamental limits on the performance of an ESA have been found in literature and it has been stated that these limits can not be

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violated. In chapter two, a brief survey on the ESA and fundamental limitations on its performance will be conducted and two general methods in antenna miniaturization will be discussed. Additionally, several notes on the classical monopole antenna and its excitation will be given in this chapter. An introduction on MTMs and their applications will be provided at the end of this chapter.

As it will be noted in the literature review, monopole antenna over a ground plane has been extensively studied in various reports due to its simplicity and its omni-directional radiation. There are actually two models to excite a monopole antenna: the coaxial feed line which is practical and also the finite gap source model which is more theoretical. In Chapter 3, applying an eigenmode analysis, both of these models will be studied and their equivalence will be verified. Also given in Chapter 3 is the optimum value for the gap height in the gap source model, which is obtained through the use of numerical approximation. On the basis of this derived optimum value, the input admittance of the monopole fed by a finite gap will be derived and also will be compared with the measured data.

In Chapter 4, a material coated monopole fed by a gap source over a large ground plane will be considered. Writing the electromagnetic fields in terms of modal functions, an analytical study will be made on the monopole with an infinitely long coating. Invoking the obtained optimum value from the previous chapter, the input admittance of the monopole will be calculated. The results show that the resonance of the monopole coated with Double Positive Materials (DPS) or Epsilon Negative Materials (ENG) would occur at lower frequencies than that of the unloaded one. To examine a more realistic case, I then resort to a monopole with a coating of limited height. Employing the modal functions of a partially filled cylindrical parallel plate waveguide in the coating region, the input admittance of the monopole will be achieved. Similarly, it will be shown

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that the first resonance of the monopole occurs at lower frequencies due to use of dielectric materials.

Taking advantage of both the miniaturization methods, it is aimed to combine the structure bending and material loading methods in Chapter 5. Combining the miniaturization methods, I consider a top-hat monopole loaded with a dielectric material. It is known that adding a monopole would make the current distribution along the monopole more uniform, which leads to increase of the antenna radiated power and its radiation efficiency. In another word, the first resonance of the antenna would occur at lower frequency due to the capacitive loading (top-hat) and hence the size of the antenna could be reduced. Further reduction in the antenna size would be obtained through the use of material loading. Loading the antenna with ENG materials or high permittivity/permeability materials would provide the antenna with more channels to radiate into the far zone at low frequencies. Both the materials (ENG and DPS) would actually reduce the high capacitive reactance of the vertical monopole in two different ways. So these two different ways would have several advantages and shortcomings, which should be compared with each other. All these facts with more details and explanations will be presented in chapter five.

And finally, conclusions on the obtained results in the previous chapters will be given in chapter six. Several recommendations on the usage of MTMs in antenna design and other applications will also be provided in chapter six.

CHAPTER 2

Literature Review

2.1 Electrically Small Antenna

Antenna is an element that accepts electrical energy from the circuits behind and then converts it into electromagnetic energy, which is propagating through free space. Antenna performance can be studied by considering several factors as input impedance, frequency bandwidth, radiation pattern and efficiency. It is well-known that such parameters are dependant on antenna size. As a matter of fact, any usual antenna with smaller dimensions presents poor performance indices. Before going further, it seems necessary to distinguish between small antenna and ESA. Small antenna is an antenna, which possesses physically small dimensions but ESA is a radiator with dimensions smaller than the wavelength of operation. Immediately, a very fundamental question appears and this is how small an antenna must be to be called ESA?

An antenna is called ESA if it can be surrounded by a sphere named “radian sphere”. Radian sphere is originated from Wheeler’s work [1-3] and is referring to an imaginary sphere of radius $\lambda/2\pi$. Hence any antenna of maximum size smaller than $\lambda/2\pi$, is called ESA.

So ESA is an antenna whose maximum size is a fraction of wavelength. As I mentioned before, the first person who mostly studied ESAs was Harold Wheeler. He studied the ESAs in a more qualitative way than quantitative. Defining the concept "radian sphere" on the basis of the effective area of an antenna, he determined a factor to separate ESAs from others [1]. In his later work [2], Wheeler regarded an ESA as a capacitor or inductor, which can be tuned to resonance by an opposite kind of reactive element. But this resulted resonance peak that is very sharp subjected to a fundamental limitation measured by antenna “radiation power factor (RPF)”. He considered this RPF

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proportional to antenna “effective volume”. It is therefore seen that ESA antenna performance is restricted by some kinds of fundamental limitations. It means that antenna size reduction causes an exponential increase in cost parameters [4].

Wheeler tried to explain the fundamental limitations of ESA in terms of RPF. But all the authors after him studied the fundamental limitations on ESAs in terms of another parameter called “radiation quality factor” which seems to be inversely proportional to the RPF. Considering an antenna as a resonant circuit, the radiation quality factor at resonance frequency is defined as

$$Q = \frac{\omega W_{total}}{P} \quad (2.1)$$

where W_{total} is the time-average energy stored in the network, P is the power dissipated in the network and ω is the radian frequency. If we take the input impedance of an antenna purely real at its resonance frequency, then the radiation quality factor can be defined as [5],

$$Q = \frac{2\omega W}{P} \quad (2.2)$$

where W is the larger of W_m (the time-average magnetic stored energy) or W_e (the time-average electric stored energy). The above definition for the antenna Q is just at the resonant frequency. Since the ESA frequency bandwidth is narrow, it can be reasonable to take the radiation quality factor equal to fractional bandwidth. But this equality for large frequency bands is questionable. Collin and Rothschild [5] and later Fante [6] proposed general methods to calculate the antenna radiation quality factor.

As I pointed out above, antenna size reduction is restricted by fundamental limitations. Thus there must be a minimum limit for the antenna radiation quality factor for an ESA. Having the knowledge of minimum possible value for radiation quality factor of an ESA is extremely important from the design point of view. In 1948, L. J. Chu published his classical paper

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[7], which dealt with the minimum limit of the omni-directional antenna radiation quality factor. He considered a general purpose antenna that is encircled in the smallest possible sphere. He took all the electromagnetic fields external to the surrounding sphere and wrote them in terms of weighted spherical waves. In order to separate the non-propagating energy from the total energy, he considered an equivalent ladder network for each mode. Then he derived an equivalent second-order RLC circuit to obtain the non-propagating energy by calculating the stored energy in the equivalent inductor and capacitor. Chu's lower limit on the radiation quality factor is

$$Q_{\min} = \frac{1}{k^3 a^3} + \frac{1}{ka} \quad (2.3)$$

with $k=2\pi/\lambda$, where λ is the free-space wavelength and a is the radius of the enclosing sphere. It should be mentioned that Chu's work was based on two significant approximations, which made his work to some extent questionable [7]. But afterward, his theoretical limit was used by other authors in different applications. Harrington did a similar work to find the lower limit on the radiation quality factor of a circularly polarized antenna [8].

In 1996, McLean re-examined the fundamental limits on the radiation quality factor of an ESA [9]. In spite of Chu's approach, he derived the radiation quality factor of the TM_{01} mode in a more direct way. His lower limit was slightly bigger than Chu's limit, while their agreement for $ka \ll 1$ was excellent. One of the efforts in order to approach this limit is presented in [10]. However, because of its importance, limitations on the performance of the ESAs have been debated in so many literatures in previous years. Traditional complex Poynting theorem is re-examined to find the radiation quality factor and also the bandwidth of antennas radiating both TE and TM modes in [11]. A time domain technique was applied in [12] to study the reactive energy and radiation quality factor in short-pulse antennas. Introducing complex frequency in Laplace domain and taking the antenna impedance as an analytic function in

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the complex frequency plane, the authors in [13] showed that Foster reactance theorem is valid for antennas and then employed it to find their radiation quality factor. A very detailed work on the input impedance of an antenna, radiation quality factor and its relation to the operating bandwidth was carried out in [14]. Small self-resonant wire antennas which possess total lengths larger than their overall lengths was discussed in [15]. Dividing a small linear antenna located in any orientation into two orthogonal vertical and horizontal dipoles (neglecting the multi-poles), in which any linear orientation can be constructed, Sten *et al* followed the same way as Mclean and also borrowed the Stratton's formulation to calculate the minimum quality factor of a small linear antenna close to the ground plane [16]. In [17], the superconductivity concept was proposed to revise McLean's lower bound of the radiation quality factor for ESAs [9]. Geyi re-examined the Fante's work [6] in his paper "physical limitations of antennas" [18]. He believes that Fante did not consider any proper constraint on the process of minimizing the radiation quality factor (Q) or maximizing the ratio of antenna gain to Q [18]. Imposing some constraints and using the Schwartz inequality, Geyi found new lower limits on $Q_{\min}$ and $(G / Q)_{\max}$ for both directional and omni-directional linear antennas [18]. Utilizing Poynting theorem in both frequency and time domains, he has also suggested a new method for evaluation of small antenna quality factor in [19]. Expressions for the Q of an electrically small antenna surrounded by a spheroidal surface, which can be identical to that of the spherical coordinate was obtained in [20], as well. In addition to all these mentioned references, many other efforts on the electrically small antenna and its miniaturization can be found in literatures such as [21-26].

The fundamental idea behind the antenna miniaturization is originated from Wheeler's work [1], where he related small radiation power factor (or large radiation quality factor) of an ESA to its diminutive effective volume. Hence reducing the radiation quality factor of an ESA, one needs to increase its

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effective volume. In another word, the volume of the imaginary sphere circumscribing ESA must be used effectively in order to reduce its radiation quality factor. Consequently, antenna miniaturization methods, in a general view, can be classified into two categories: a) Antenna miniaturization with the use of novel topologies and b) Antenna miniaturization with the use of material loading.

The first method is a more innovative way, which tries to occupy the volume of the mentioned sphere as much as possible. Top-hat monopoles, folded monopoles and thick monopoles are simple types of miniaturized antenna occupying more space. Recently, S. R. Best has studied the performance of the folded spherical dipoles in more details [27, 28] and shown that increasing the effective volume of an ESA can enhance its performance. A similar principle has been employed to approach the lower bounds of radiation quality factor of an ESA in [10, 29]. On the other hand, fractal shapes have been attracted attentions in antenna miniaturization because of their self-similarity and also space filling properties in recent years [30-36]. Their space filling properties make their total dimensions so large while their overall dimension is fixed. This idea can be used to reduce the radiation quality factor. Slot antenna and patch antenna are other two large groups of antenna, which have presented great benefits in antenna miniaturization due to their flexibilities and radiation mechanisms though they are not that useful for high power applications [37-44].

Loading antenna by dielectric materials with constitutive parameters larger than one is another method to reduce the size of the antenna. Various types of such antennas can be found in [45]. Using a material with high permittivity and/or high permeability causes a reduction of $\lambda_g = \lambda / \sqrt{\mu_r \epsilon_r}$ in which λ is the free-space wavelength. Hence, we have a decrease in the radian length $\lambda_g / 2\pi$ which leads to diminishing the volume of radian sphere. In

another word, effective volume of ESA approaches to that of the radian sphere. As a result, we can have a smaller radiation quality factor. From the other point of view, an ESA antenna is a resonator with a high quality factor. So we have

$$\frac{1}{Q_1} = \frac{1}{Q_A} \quad (2.4)$$

Loading the same antenna by a dielectric adds another term to the right side of equation (2.4),

$$\frac{1}{Q_2} = \frac{1}{Q_A} + \frac{1}{Q_D} \quad (2.5)$$

If we have $\sqrt{\mu_r \epsilon_r} > 1$, then we will get $Q_D > 1$. And it leads to $1/Q_2 > 1/Q_1$ or $Q_2 < Q_1$. We see that loading an ESA by a dielectric material can reduce the radiation quality factor of the whole system. More details can be found in [46]. The idea of antenna size reduction with the use of magneto-dielectric materials has been applied in [47-53].

2.2 Gap Source Model for Wire Antennas

From the first days when Hertz tried to verify Maxwell's equations till now, the gap source excitation has been widely implemented in wire antenna feeding. In his experiment, Hertz utilized a spark gap to derive a wire antenna of length 60cm terminated by a metallic plate of 40cm square at each side [54]. After that, authors and researchers have invoked the gap source to model the feedings of wire antennas for more than a century [55-59].

Obtaining the current distribution along the wire antenna is the key point to achieve different characteristics of the antenna such as radiated fields in near and far zones, directivity, radiation efficiency and input impedance of the antenna in the deriving point. The gap source excitation has been used in so many literatures due to its simplicity to explain the sinusoidal current distribution along the wire antennas. However, the accuracy and validity of the

gap source model are more questionable than that of a more feasible excitation such as coaxial feed line.

The delta gap source model has been extensively used in the wire antenna analyses based on the method of moments (MoM). However, the poor convergence of the current distribution around the deriving point is a drawback about this model [60]. To avoid this problem, two solutions have been proposed. The first was the use of a gap source model with a Gaussian field distribution and fixed length [60]. The second method was to employ a magnetic frill source which was of limited use due to its complexity in finding the near fields [61, 62]. Then a point matching was used to study the gap excited wire antenna [62] and also it was suggested that the height of the gap source should be fixed [63]. A moderately thick cylindrical wire antenna with a fixed gap source was examined in [64].

The finite difference time domain (FDTD) technique also has been employed in literature to model the gap source model in wire antennas and examine its validity in computing radiation properties of such antennas. A monopole fed by a simple delta-gap FDTD source model over a conducting box was studied in [65]. However, the simulation in [65] was not able to truly explain the field distribution in the practical excitation zone consisting of a coaxial cable with a proper length. In [66], a quasi-static electric-field FDTD source model, that replaces the physical coaxial excitation with a finite air gap over a ground plane [67, 68], was used to study a monopole antenna mounted on a cylinder. The field distribution of the quasi-static finite gap in [66] was similar to that of [68].

2.3 Anomalous Meta-Materials

Meta-material (MTM) is a “macroscopic composite having a manmade, three dimensional, periodic cellular architecture designed to produce an optimized combination, not available in nature, of two or more responses to

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specific excitation” [69]. In fact, MTMs are structures with unusual properties, which acquire their uncommon behavior from their structure rather than their compositions. Inclusion of electric and/or magnetic dipolar elements periodically or even randomly in a medium offers effective permittivity and/or effective permeability with unusual behaviors to that medium. Such behaviors have been shown in [70-72].

The concept of left-handed media and negative refraction was initiated by Schuster in 1904 [73] where he showed that an incident electromagnetic wave upon a medium in which the group and phase velocities are in reverse directions (back-ward wave) would be bent in an unusual way in comparison with normal media. Afterward similar concepts on such a negative refraction, the back-ward wave propagation, dispersion behavior in such media and their circuit models and even the usage of negative refraction in parallel plane lenses were studied in [73-76]. But the first person who considered negative values for both the constitutive parameters and also classified their unusual behaviors was Veselago [77]. Assuming negative values for permittivity and also permeability of a medium, Veselago showed the negative refraction for an incident wave from a normal medium upon the mentioned medium and concluded negative values for their refractive indices [77]. He called these media the negative refractive index media or left-handed media due to the left-handness of three vectors, electric field, magnetic field and wave vector in such media [77]. He also predicted other dramatic anomalies in left-handed materials such as reversal of Doppler shift and also the Vavilov-Cherencov radiation, reversal of Goos-Hanchen effect, transformation of a point source to a point image due to a left-handed parallel plane lens [77].

The thing that diminished the interests in left-handed materials in the time of publishing of Veselago’s paper was the unavailability of materials with both negative permittivity and permeability in nature though epsilon-negative (ENG) or mu-negative (MNG) media were observed in different experiments [70, 71,

78, 79]. So the negative refraction idea did not attract much interest till recently that the idea of periodically combining arrays of rods (negative permittivity [72]) and split ring resonators (negative permeability [80]) with periodicities and element dimensions much smaller than the wavelength was employed to observe a composite medium with both negative permittivity and permeability in [81]. As it is shown in Fig. 2.1, the possible values of permittivity and permeability can be divided into four quadrants: Double Positive (DPS), Epsilon Negative (ENG), Mu Negative (MNG) and Double Negative (DNG).

Fig. 2.1. Four quadrants of possible values for permittivity and permeability.

Utilizing a similar structure as a sample, Shelby *et al* [82] assembled a setup as shown in Fig. 2.2. They showed that the normalized detected power for the Teflon sample is at a positive angle from normal (usual refraction) and on the contrary the refracted power is detected at a negative angle for the left-handed sample [82]. As a matter of fact, they could verify the negative refraction in left-handed media in an experimental way [82].

Pendry investigated Veselago's theory on parallel plane lens made up of left-handed materials in [83]. He showed that imaging of an infinitesimal dipole in front of an ordinary lens is not perfect due to the exponential decaying of evanescent waves and there is a limit on the resolution of the lens [83]. Then letting the refractive index of the lens equal to -1, Pendry showed that the propagating and evanescent waves would contribute in imaging of the

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infinitesimal dipole by the lens and hence there would be no limit on the resolution of the lens which is called perfect lens by Pendry [83]. However several articles like [84, 85] questioned Pendry's prediction on perfect imaging. And also in [86] it was proved that Pendry's suggestion on the possibility of perfect lens with only one negative parameter among constitutive parameters is not correct.

Fig. 2.2. The proposed 2D negative refractive index structure in [80]

After such reports and also because of the unusual physics, MTMs attracted an extensive amount of attention and became one of the hottest area of research in electromagnetic wave theory and its applications. A very wide range of publications can be found in literature on the usage of MTMs in different parts of science. Among all these applications, I can point to the application of MTMs in super lenses and near field imaging [87], photonic crystals [88], filters and other radio frequency circuit components [89, 90], electrically small antennas [91, 92], directive antennas [93, 94], leaky wave antennas [95], surface wave suppression or propagation [96, 97] and pairing epsilon negative (ENG) and mu-negative (MNG) slabs to have unusual features like perfect tunneling [98]. The recently proposed idea of cloaking objects and making them invisible [99-101] has triggered the imaginations to have invisible men or

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invisible clothes which are attributed to fantasy movies rather than the real world. Groups of researchers are busy with fabrication and testing MTMs in different regimes and in different sizes. Examples of such efforts are [82, 102, 103].

The disappointing point about MTMs is their highly frequency dispersive behaviors. As it was shown in [81], periodic inclusion of conductive elements with small periodicity in comparison with the wavelength would lead to an effective permittivity and/or permeability, which are highly frequency dependant. The general form of $\varepsilon(\omega)$ and $\mu(\omega)$ can be expressed as [81],

$$\varepsilon(\omega) = 1 - \frac{\omega_p^2}{\omega(\omega + i\gamma)} \quad (2.6)$$

$$\mu(\omega) = 1 - \frac{F\omega^2}{\omega^2 - \omega_0^2 + i\omega\Gamma} \quad (2.7)$$

The parameters in the above equations were defined in [81] with more details. It is seen that both the constitutive parameters have resonant frequency dependant behaviors, which restrict the usage of MTMs in different areas of electromagnetics.

2.4 Top-Hat Monopole Antennas

One of the methods to improve the radiation performance of electrically small antennas is the design of structures that occupies volume of the radian sphere more effectively [1-3]. Monopole antenna is the simplest form of antenna that has been studied extensively. It has been shown that the current distribution along the vertical monopole over an infinitely large ground plane is a sinusoidal function, which vanishes at the end of monopole. As a consequence, for electrically short monopoles the current curve at low frequencies is a sharp triangular curve and the integral area under that is so small. It is known that the area under such a curve is related directly to the

radiated power of the antenna. Therefore one needs a solution to avoid this null in the current distribution and make the current distribution more uniform.

Fig. 2.3. Examples of the top-hat monopole; a) horizontal hat, b) inclined hat

Intuitively speaking, we can lengthen the antenna horizontally as a top-hat to get the current to flow further and avoid the null at the end of monopole. In other word, we have enlarged the effective volume of the antenna without exceeding the circumscribing sphere around it. We can make this effective volume larger and larger till the top-hat radius becomes larger than the length of the monopole antenna. Adding such a hat, it seems that the performance of the short monopole improves and the radiated power is increased due to the more uniform current distribution along the monopole. This top-hat can be several rods or a cylindrical disc which are oriented parallel or inclined toward the ground plane. The experimental radiation characteristics of electrically short umbrella top-loaded monopole were reported in [104]. Employing an eigenmode approach, the dielectric loaded top-hat monopole was studied in [105]. Utilizing a similar formulation to [105], Francavilla *et al* examined the properties of a top-hat monopole loaded with radially layered dielectric materials [106]. In [107], the generalized form of the top-hat monopole antenna loaded with homogeneous or inhomogeneous layered materials over an infinite or finite ground plane was analyzed on the basis of FDTD method. Making a small change in the geometry of usual top-hat monopole, a broad band top-cap

monopole was studied in [108]. The geometry of proposed antenna in [108] was optimized with the genetic algorithm. The relation between the top-hat monopole antenna, the disc resonator diode mounts and the coaxial to waveguide transitions were discussed in [109].

2.5 Summary

In this chapter, I conducted a brief literature study on the ESA and managed to figure out the challenges in front of antenna miniaturization in a classical way. The quantitative definition of an ESA was given and the fundamental limitations on the performance of a general purpose ESA reviewed. Two main miniaturization techniques and their effects on the operation of an ESA were introduced.

Since I am to combine the mentioned techniques (shape bending and material loading) to achieve a highly miniaturized antenna and also make a comprehensive comparison between ENG and DPS materials in vertical monopole miniaturization, a brief introduction into the MTMs and their properties and also a short description on the top-hat monopole antenna were provided. On the other hand, as I am going to utilize the gap source excitation in my formulation, several literatures on the usage of gap source model in electromagnetic wave theory were reviewed.

CHAPTER 3

The Gap Source Model for Monopoles

3.1 Fundamentals

Before I deal with the miniaturization problem and the possibilities of loading the antenna with MTMs for the purpose of size reduction, I review a fundamental problem in wire monopole antennas. Validating the equivalence between the gap source model and the coaxial feed model for monopoles and finding the optimum value for gap height, I shall use the gap model in the next chapters to study the effect of the MTM loading on the monopole radiation properties.

Monopole antenna has been extensively studied due to its structural simplicity and omni-directional radiation properties in various literatures [110-113]. Different methods have been utilized to study the radiation performance of the monopole antenna over an infinite or finite ground plane. In [114], an integral equation method was employed to study the performance of a thick monopole antenna fed by a finite gap source over an infinite ground plane and its current distribution. The Finite Difference Time Domain (FDTD) method [115], the conservation of complex power method [116] and also the modal expansion method [105, 117-119] are other techniques which have been applied as well.

In general, two feed models have been proposed in modeling a monopole antenna excitation. The first one is the coaxial feed model, which is a very realistic feed though it is complicated to be studied [116, 118, 119]. The second one is the voltage drop across a finite gap over the ground plane, which is simpler than the first one but not physically feasible [105, 114, 117]. In this chapter, I am going to set up the equivalence between these two feed models. A modal analysis will be used to find the input admittances of both models. In

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order to find the gap height in the second model, the obtained input admittances will be set equal to each other. Numerical results will be presented to verify the accuracy of derived equations and also prove validity of the suggested equivalence between the two models.

3.2 The Gap Source Model

3.2.1 Formulation

Consider a monopole antenna of radius a and height h . As it can be seen in Fig. 3.1, this monopole is fed by a voltage drop across a finite gap with the height of Δ over an infinite ground plane. In this axially symmetric structure, I introduce a perfect electric conductor (PEC) at a distance D from the ground plane. This assumption does not affect the calculation of the monopole antenna's input admittance or its current distribution, since the radiation in the vertical direction is null and the major part of the radiation occurs horizontally. As it will be shown later, changes in D will not affect the input admittance of the monopole, if it is sufficiently large. It should be mentioned that a circular disc has been postulated in the PEC above the monopole (Fig. 3.1) to facilitate the mode-expansion analysis due to the weak field coupling in the z direction [118, 119]. As a result, approximate single-mode expression can be obtained in the small region above the monopole.

Dividing the domain of interest into several regions, as shown in Fig. 3.1, I shall write the electromagnetic fields in terms of eigenmodes that can exist in each region, weighted with unknown coefficients. It is noted that I just consider a uniform E field across the gap, since its height is much smaller than the monopole radius. So in region I, we have

$$E_z^I = \frac{1}{j\omega\epsilon_0} A^I \quad (3.1)$$

which leads to

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$$V_{in} = \frac{j\Delta}{\omega\epsilon_0} A' \rightarrow A' = \frac{\omega\epsilon_0}{j\Delta} V_{in} \quad (3.2)$$

where ω is the angular frequency, ϵ_0 is the free space permittivity and A' is an unknown coefficient. The electromagnetic field expressions in region II (at $\rho=a$) can be written as

$$E_z^{II} = \frac{1}{j\omega\epsilon_0} \sum_{m=0}^{\infty} A_{2m} Z_{2m} \cos\left[\frac{m\pi(z-h)}{d}\right] \quad (3.3)$$

$$H_\phi^{II} = \sum_{m=0}^{\infty} A_{2m} \cos\left[\frac{m\pi(z-h)}{d}\right] \frac{\epsilon_m}{d} \quad (3.4)$$

Fig. 3.1. Geometry of a monopole fed by a gap source

where

$$Z_{2m} = \frac{\epsilon_m}{d} \frac{\gamma_{2m} J_0(\gamma_{2m} a)}{J_1(\gamma_{2m} a)}, \gamma_{2m}^2 = k_0^2 - \left(\frac{m\pi}{d}\right)^2, m = 0, 1, 2, \dots, \epsilon_m = \begin{cases} 1 & m = 0 \\ 2 & m \neq 0 \end{cases},$$

$k_0 = \omega\sqrt{\mu_0\epsilon_0}$, μ_0 is the free space permeability, J_0 and J_1 are the first kind Bessel functions of order zero and one, respectively. In region III (at $\rho=a$), we have

$$E_z^{III} = \frac{1}{j\omega\epsilon_0} \sum_{n=0}^{\infty} A_{3n} \cos\left(\frac{n\pi z}{D}\right) \frac{\epsilon_n}{D} \quad (3.5)$$

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$$H_{\phi}^{III} = \sum_{n=0}^{\infty} A_{3n} Y_{3n} \cos\left(\frac{n\pi z}{D}\right) \quad (3.6)$$

where $Y_{3n} = \frac{\epsilon_n H_1^{(2)}(\gamma_{3n} a)}{D \gamma_{3n} H_0^{(2)}(\gamma_{3n} a)}$, $\gamma_{3n}^2 = k_0^2 - \left(\frac{n\pi}{D}\right)^2$, $n = 0, 1, 2, \dots$, $H_0^{(2)}$ and $H_1^{(2)}$ are

the second kind Hankel functions of order zero and one, respectively.

Satisfying the boundary conditions across the interfaces between the regions and with some mathematical manipulations, we can obtain

$$A_{3n} = A^I F_{In} + \sum_{m=0}^{\infty} A_{2m} Z_{2m} F_{IImn} \quad (3.7)$$

$$A_{2m} = \sum_{n=0}^{\infty} A_{3n} Y_{3n} F_{IImn} \quad (3.8)$$

where

$$F_{IImn} = \frac{nd^2 D}{\pi(n^2 d^2 - m^2 D^2)} \left\{ (-1)^m \sin\left[\frac{n\pi(h+d)}{D}\right] - \sin\left(\frac{n\pi h}{D}\right) \right\}, n, m = 0, 1, \dots,$$

$$F_{In} = \Delta \text{sinc}\left(\frac{n\pi\Delta}{D}\right) \text{ and } \text{sinc}(x) = \frac{\sin(x)}{x}.$$

Or in the matrix form

$$A_3 = A^I F_I + F_{II} Z_2 A_2 \quad (3.9)$$

$$A_2 = F_{II}^T Y_3 A_3 \quad (3.10)$$

where the superscript “T” indicates the matrix transpose.

Solving these matrix equations, we can find the unknown coefficients with respect to A^I . And then we are able to obtain different characteristics of the antenna such as input admittance, current distribution and directivity pattern. Specifically for the current distribution, we have

$$I(z) = 2\pi a \left(\hat{\mathbf{a}}_{\rho} \times \hat{\mathbf{a}}_{\phi} H_{\phi}^{III} \right) = 2\pi a \sum_{n=0}^{\infty} A_{3n} Y_{3n} \cos\left(\frac{n\pi z}{D}\right) \quad (3.11)$$

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Having the current at the excitation point ($z=A$), we can obtain the input admittance as follows

$$Y_{in} = \frac{I_{in}}{V_{in}} = G_{in} + jB_{in} = \frac{\omega\epsilon_0}{j\Delta A^I} 2\pi a \sum_{n=0}^{\infty} A_{3n} Y_{3n} \cos\left(\frac{n\pi\Delta}{D}\right) \quad (3.12)$$

And also the radiated electromagnetic fields at the far zone can be obtained through the following far-field integration [120],

$$\vec{E} = \frac{-j\omega\mu_0}{4\pi} \iint_s \vec{J}(r') \frac{e^{-jk_0 R}}{R} ds' \quad (3.13)$$

$$\vec{H} = \frac{1}{\eta} \vec{E} \quad (3.14)$$

where $\eta \approx 120\pi$ is the free space wave impedance. After some straight manipulations, we have

$$\vec{E} = \hat{a}_\theta E_\theta = \hat{a}_\theta \frac{j\mu_0 a \omega e^{-jk_0 r}}{8\pi^2 r} \sin(\theta) J_0(k_0 a \sin(\theta)) \sum_{n=0}^{\infty} A_{3n} Y_{3n} \times P_n(\theta) \quad (3.15)$$

where

$$P_n(\theta) = \frac{1}{\left(\frac{n\pi}{D}\right)^2 - k_0^2 \cos^2 \theta} \left\{ jk_0 \cos \theta \left[e^{jk_0 h \cos \theta} \cos\left(\frac{n\pi h}{D}\right) - 1 \right] + \frac{n\pi}{D} e^{jk_0 h \cos \theta} \sin\left(\frac{n\pi h}{D}\right) \right\}.$$

The directivity of the monopole antenna can be calculated as

$$D(\theta) = \frac{2[F^2(\theta)]}{\int_0^\pi F^2(\theta) \sin(\theta) d\theta} \quad (3.16)$$

where $F(\theta) = \sin(\theta) J_0(k_0 a \sin(\theta)) \sum_{n=0}^{\infty} A_{3n} Y_{3n} P_n(\theta)$ and $[F^2(\theta)]_{\max}$ or $D_{\max}$

occurs at $\theta=0$.

3.2.2 The One-Mode Approximation

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In order to find an approximate closed-form expression for the input admittance of the monopole antenna, we need to make an approximation. It sounds meaningful, specifically for thin monopoles with small values for d , to assume just one mode in region II. As a matter of fact, this assumption does not disturb the input admittance or the current distribution of the monopole significantly, since the radiation in the vertical direction is relatively weak. Invoking this approximation and after some manipulations, we obtain

$$Y_{in} = -j\omega\epsilon_0 \frac{2\pi a}{\beta} \sum_{n=0}^{\infty} \sum_{k=0}^{\infty} Y_{3n} (\beta\delta_{nk} + F_n F_k Y_{3k}) \text{sinc}\left(\frac{\pi k \Delta}{D}\right) \cos\left(\frac{n\pi \Delta}{D}\right) \quad (3.17)$$

where

$$F_n = \frac{D}{n\pi} \left\{ \sin\left[\frac{n\pi(h+d)}{D}\right] - \sin\left(\frac{n\pi h}{D}\right) \right\}, \beta = Z_2 - \sum_{k=0}^{\infty} F_k^2 Y_{3k}, Z_2 = \frac{dJ_1(k_0 a)}{k_0 J_0(k_0 a)},$$

$\delta_{nk} = 1$ when $n = k$ and also $\delta_{nk} = 0$ when $n \neq k$.

3.3 The Coaxial Feed Model

3.3.1 Formulation

In this part, I am to find the input admittance of the same monopole antenna fed by a coaxial transmission line with inner and outer radii a and b , respectively, as shown in Fig. 3.2. It should be pointed out that, I consider the same circular disc loaded PEC above the monopole as the previous case. Dividing the domain of interest into several regions and following a similar approach as the gap source model we can write the fields in each region. Considering only one mode in region I, the transverse electromagnetic fields in the coaxial line [121] simplifies to

$$E_{\rho}^I = (1 + R) \frac{1}{\sqrt{2\pi \ln \frac{b}{a}}} \frac{1}{\rho} \quad (3.18)$$

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$$H_{\phi}^I = (1 - R) \frac{\sqrt{\epsilon_r}}{\eta_0} \frac{1}{\sqrt{2\pi \ln \frac{b}{a}}} \frac{1}{\rho} \quad (3.19)$$

where $\eta_0 \approx 120\pi$, ϵ_r is permittivity of the dielectric between the inner and outer conductors in the coaxial transmission line and R is the unknown reflection coefficient of the TEM wave propagating through the guide. It should be mentioned that I considered only one mode in the coaxial feed line, since the ratio of outer radius to the inner radius in the coaxial transmission line is not that big. For large values for this ratio, more modes should be considered to have more accurate results. In region II and at $\rho=a$, we have

$$E_z^II = \frac{1}{j\omega\epsilon_0} \sum_{m=0}^{\infty} A_{2m} Z_{2m} \cos \left[\frac{m\pi(z-h)}{d} \right] \quad (3.20)$$

$$H_{\phi}^II = \sum_{m=0}^{\infty} A_{2m} \cos \left[\frac{m\pi(z-h)}{d} \right] \frac{\epsilon_m}{d} \quad (3.21)$$

where Z_{2m} and ϵ_m have the same definitions as those of the previous section.

The electromagnetic field expressions in region III can be written by applying the resonator method [118] as follows

$$E_z^{III} = \frac{1}{j\omega\epsilon_0} \sum_{n=0}^{\infty} \{A_{3n} U_{3n}(\rho) + B_{3n} V_{3n}(\rho)\} \cos \left(\frac{n\pi z}{D} \right) \frac{\epsilon_n}{D} \quad (3.22)$$

$$E_{\rho}^{III} = C_3 \frac{\sin k_0(D-z)}{\sin(k_0 D)} \frac{1}{\sqrt{2\pi \ln \frac{b}{a}}} \frac{1}{\rho} \quad (3.23)$$

$$H_{\phi}^{III} = \sum_{n=0}^{\infty} \frac{-1}{\gamma_{3n}^2} \{A_{3n} U'_{3n}(\rho) + B_{3n} V'_{3n}(\rho)\} \cos \left(\frac{n\pi z}{D} \right) \frac{\epsilon_n}{D} + C_3 \frac{\cos k_0(D-z)}{j\eta_0 \sin(k_0 D)} \frac{1}{\sqrt{2\pi \ln \frac{b}{a}}} \frac{1}{\rho} \quad (3.24)$$

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Fig. 3.2. Geometry of the monopole fed by the coaxial feed line

where

$$\gamma_{3n}^2 = k_0^2 - \left(\frac{n\pi}{D} \right)^2, n = 0, 1, \dots, U_{3n}(\rho) = \frac{J_0(\gamma_{3n}\rho)Y_0(\gamma_{3n}a) - Y_0(\gamma_{3n}\rho)J_0(\gamma_{3n}a)}{J_0(\gamma_{3n}b)Y_0(\gamma_{3n}a) - Y_0(\gamma_{3n}b)J_0(\gamma_{3n}a)}$$

$$\text{and } V_{3n}(\rho) = \frac{J_0(\gamma_{3n}b)Y_0(\gamma_{3n}\rho) - Y_0(\gamma_{3n}b)J_0(\gamma_{3n}\rho)}{J_0(\gamma_{3n}b)Y_0(\gamma_{3n}a) - Y_0(\gamma_{3n}b)J_0(\gamma_{3n}a)} \text{ with } Y_0 \text{ being the second}$$

kind Bessel's function of order zero.

The field expressions in region IV and at $\rho=b$ can be written in a similar way to that of the gap source model as

$$E_z^{IV} = \frac{1}{j\omega\epsilon_0} \sum_{n=0}^{\infty} A_{4n} \cos\left(\frac{n\pi z}{D}\right) \frac{\epsilon_n}{D} \quad (3.25)$$

$$H_\phi^{IV} = \sum_{n=0}^{\infty} A_{4n} Y_{4n} \cos\left(\frac{n\pi z}{D}\right) \quad (3.26)$$

$$\text{where } Y_{4n} = \frac{\epsilon_n}{D} \frac{H_1^{(2)}(\gamma_{4n}b)}{\gamma_{4n} H_0^{(2)}(\gamma_{4n}b)} \text{ and } \gamma_{4n}^2 = k_0^2 - \left(\frac{n\pi}{D} \right)^2, n = 0, 1, \dots$$

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Applying the boundary conditions across the regional interfaces, we can obtain

$$A_3 = P_3 B_3 + Q_3 C_3 \quad (3.27)$$

$$B_3 = FZ_2 A_2 \quad (3.28)$$

$$A_2 = F^T U_3 A_3 + F^T U_4 B_3 + V_2 C_3 \quad (3.29)$$

$$\text{where } P_{3n} = \frac{H_0^{(2)}(\gamma_{3n} b)}{H_0^{(2)}(\gamma_{3n} a)}, U_{3n} = \frac{\epsilon_n}{D} \left(\frac{-1}{\gamma_{3n}^2} U_{3n}'(a) \right), U_{4n} = \frac{\epsilon_n}{D} \left(\frac{-1}{\gamma_{3n}^3} V_{3n}'(a) \right),$$

$$Q_{3n} = j\omega\pi\epsilon_0 \frac{H_0^{(2)}(\gamma_{3n} b) \{J_0(\gamma_{3n} a) Y_0(\gamma_{3n} a) - Y_0(\gamma_{3n} b) J_0(\gamma_{3n} a)\}}{\sqrt{2\pi \ln \frac{b}{a}} H_0^{(2)}(\gamma_{3n} a)},$$

$$F_{nm} = \frac{nd^2 D}{\pi(n^2 d^2 - m^2 D^2)} \left\{ (-1)^m \sin \left[\frac{n\pi(h+d)}{D} \right] - \sin \left(\frac{n\pi h}{D} \right) \right\},$$

$$V_{2m} = \frac{k_0}{a \sqrt{2\pi \ln \frac{b}{a}}} \frac{\sin[k_0(D-h)] - (-1)^m \sin[k_0(D-h-d)]}{j\eta_0 \gamma_{2m}^2 \sin(k_0 D)}.$$

Solving the above matrix equations, we have

$$A_2 = T_1 C_3 \quad (3.30)$$

$$B_3 = T_2 C_3 \quad (3.31)$$

$$A_3 = T_3 C_3 \quad (3.32)$$

$$\text{where } T_1 = (I - F^T Y_4 F Z_2)^{-1} V_C, T_2 = F Z_2 T_1, T_3 = P_3 T_2 + Q_3 \quad \text{and}$$

$$V_C = F^T U_3 Q_3 + V_2. \text{ It also can be seen from the boundary conditions that}$$

$$(1 - R) = \frac{\eta_0}{\sqrt{\epsilon_r}} W C_3 \quad (3.33)$$

$$1 + R = C_3 \quad (3.34)$$

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where $W = \frac{1}{j\eta_0 \tan(k_0 D)} + W_1(T_3 - T_2)$ and $W_{1n} = \frac{-\epsilon_n}{\gamma_{3n}^2 D} \sqrt{\frac{2\pi}{\ln \frac{b}{a}}}$. Finally we

can find the input admittance of the monopole fed by the coaxial feed line as

$$Y_{in} = Y_0 \frac{1-R}{1+R} = \frac{2\pi}{\ln\left(\frac{b}{a}\right)} W \quad (3.35)$$

$$\text{where } Y_0 = \frac{\sqrt{\epsilon_r}}{60 \ln\left(\frac{b}{a}\right)}.$$

3.3.2 The One-Mode Approximation

In order to derive an approximate closed-form expression for the input admittance of the monopole fed by a coaxial line, I invoke the same single-mode approximation as what I did for the gap model. Again, as the radiation in the vertical direction is a null, my approximation seems reasonable. Hence, assuming a single mode in region II, with some kinds of algebraic manipulations we have

$$Y_{in} = \frac{j}{60[\ln(b/a)]^{1.5}} \left\{ \frac{-\sqrt{\ln(b/a)}}{\tan(k_0 D)} + \frac{k_0}{\sqrt{2\pi}} \sum_{n=0}^{\infty} W_{1n} \left[\frac{V_2 F_n}{a\beta} (P_{3n} - 1) + Q_{3n} \right] \right\} \quad (3.36)$$

$$\text{where } P_{3n} = \frac{H_0^{(2)}(\gamma_{3n} b)}{H_0^{(2)}(\gamma_{3n} a)}, Q_{3n} = \frac{\pi P_{3n}}{2} [J_0(\gamma_{3n} b) Y_0(\gamma_{3n} a) - Y_0(\gamma_{3n} b) J_0(\gamma_{3n} a)]$$

$$V_2 = \frac{\sin[k_0(D-h-d)] - \sin[k_0(D-h)]}{k_0^2 \sin(k_0 D)} + \sum_{k=0}^{\infty} \frac{\epsilon_k F_k P_{3k}}{D \gamma_{3k}^2}, \beta = Z_2 - \sum_{k=0}^{\infty} F_k^2 Y_{3k} \text{ and}$$

$$Z_2 = \frac{dJ_1(k_0 a)}{k_0 J_0(k_0 a)}.$$

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3.4 Finding the Gap Δ

In previous sections, I derived approximate closed-form expressions for the input admittance of a monopole antenna fed by the gap source and the coaxial line, respectively. In this part, I am going to find the best value for Δ , which provides equivalence between two analyzed models. In order to estimate the value of Δ , which makes the gap model equivalent to the coaxial model, I let the approximate closed-form expressions from equations (2.17) and (2.36) equal to each other.

$$Y_{in_{Gap}} = Y_{in_{Coax}} \quad (3.37)$$

or,

$$\begin{aligned} f(\Delta) = Y_{in_{Gap}} - Y_{in_{Coax}} = \sum_{n=0}^{\infty} \left(Y_{3n} \cos\left(\frac{n\pi\Delta}{D}\right) \left[\sum_{k=0}^{\infty} (\beta\delta_{nk} + F_n F_k Y_{3k}) \operatorname{sinc}\left(\frac{k\Delta}{D}\right) \right] \right. \\ \left. + \frac{W_{1n}}{a^2 \sqrt{2\pi [\ln(b/a)]^3}} \left[V_2 F_n (P_{3n} - 1) + a\beta Q_{3n} \right] \right) - C_0 = 0 \end{aligned} \quad (3.38)$$

where $C_0 = \frac{\beta}{k_0 a \ln(b/a) \tan(k_0 D)}$. Solving the equation (2.38), we can calculate the optimum value for Δ , which provides the equivalence between the two models.

3.5 Numerical Results

I applied the modal expansion analysis in the previous sections to estimate the value of Δ , which makes two described models equivalent. As a result, I derived equation (2.38) that estimates this value. But as it can be seen from equation (2.38), $f(\Delta)$ is a complex function, which possesses real and imaginary parts both containing Δ . Thus in order to find the unknown Δ , both

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parts must be vanished. But it appears unlikely to find an analytical value for Δ from equation (2.38) as it is too complicated. Consequently, I resort to a numerical technique [122] to obtain an approximate Δ , which is sufficiently close to the precise one. In this numerical method, I am going to minimize $|f(\Delta)|$.

It should be pointed out that, I have to truncate the number of modes considered in regions III (in Figs. 3.1 and 3.2) and IV (in Fig. 3.2) so that the corresponding summations have finite terms. But the truncated numbers must be large enough to have good convergences in the summations [123].

Fig. 3.3 shows the calculated Δ for a monopole excited by a gap source that is equivalent to the same monopole fed by a coaxial line. It is seen that we have similar values for thin monopoles, which means that b does not have a considerable effect for thin monopoles. It is also observed that the Δ value is approximately proportional to the inner radius of the monopole while the proportional constant is related to the outer radius of the coaxial feed.

In order to gain a better understanding of the effect of monopole height on the optimum value of Δ , I illustrate the variation of Δ with respect to a for different monopole heights in Fig. 3.4. As it is expected, monopole height does not have a considerable effect on the optimum value of Δ for thin monopoles. However for a thick monopole, its height can change the optimum value of Δ .

TABLE 3.1. Input admittance of a monopole with gap feed for different values of D ($a=0.0509\lambda$, $b=1.189a$, $h=\lambda/4$, $\Delta=0.005\lambda$)

D	G_{in} (mhos)	B_{in} (mhos)
1.5λ	0.0191	0.0048
2λ	0.0190	0.0048
2.5λ	0.0189	0.0050
3λ	0.0189	0.0050

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Fig. 3.3. Optimum values for Δ which make the gap model equivalent to the coaxial model ($h=\lambda/4$)

Fig. 3.4. Optimum values of Δ which make the gap model equivalent to the coaxial model for different monopole heights ($b=2.301a$)

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During my formulation, I assumed a disc loaded PEC above the monopole at the distance of D from the ground plane and I stated that this PEC wall will not influence the calculation of the input admittance of the monopole if D is large enough (say $D > 1.5\lambda$) [105, 118, 119]. This insignificant effect of large D is presented in Table 3.1 where it is seen that for different values of D , we have almost identical values for the input admittance of the monopole antenna.

Fig. 3.5. Comparison between approximate and accurate input admittances of the monopole fed by the gap source ($a=0.00635\lambda$, $b=1.189a$, $d=0.0006\lambda$)

On the other hand, I took a single mode in region II for a small d , in both of the structures. Based on this approximation, I derived an approximate closed-form expression for the input admittance of the monopole. In order to assess this approximation, I compare the input admittance results of a monopole with small and large values of d in Fig. 3.5. It is observed that my assumption is very accurate, especially for thin monopoles.

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The input conductance and susceptance of a monopole fed by a gap source are indicated in Figs. 3.6 (a) and (b), respectively. Additionally, the input conductance and susceptance of the coaxial model are presented in the same figures too. It is observed that we have equivalence between two models only with a specific value for Δ , which is obtained from equation (2.38). It is also noticed that the input conductance of the gap model is not sensitive to the gap height, while its input susceptance changes noticeably when Δ changes.

Finally, I compare my calculated results for the input admittance of a monopole fed by a gap source with the measured data in [124] for a thin and a thick monopole in Figs. 3.7 and 3.8, respectively. In these figures, I have obtained the gap height Δ from equation (2.38). It is observed that, the calculated and measured data are in an excellent agreement. It should be mentioned that I have already published the obtained results of this chapter in [125].

3.6 Summary

In this chapter, I utilized the modal expansion method to assess the accuracy of the gap source model in comparison with the practical coaxial feed model for monopole antennas. I acquired two approximate closed-form expressions for the input admittances of monopole antenna in both the models. Employing the obtained approximate closed-form expressions, I derived a closed-form equation to estimate the gap height, which makes two feed models equivalent. Numerical results were presented to demonstrate the equivalence and also the accuracy of obtained equations.

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(a)

(b)

Fig. 3.6. Computed input admittance of a monopole antenna fed by a gap source with different values of Δ and a coaxial line ($a=0.0509\lambda$, $b=1.189a$)

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Fig. 3.7. Computed input admittance of a thin monopole antenna fed by a gap source and a coaxial line ($a=0.001588\lambda$, $b=1.189a$, $\Delta=0.0001\lambda$)

Fig. 3.8. Computed input admittance of a thick monopole antenna fed by a gap source and a coaxial line ($a=0.0726\lambda$, $b=1.189a$, $\Delta=0.0072\lambda$)

CHAPTER 4

Monopole Antenna with Dielectric Coating

4.1 Introduction

All the electromagnetic phenomena can be explained on the basis of Maxwell's equations, which collectively offer a rich physics with a high degree of flexibility. Due to such flexibility, many interesting phenomena specifically in the interactions between materials and electromagnetic waves and sources, could occur. Refraction of electromagnetic waves and their propagation in high permittivity and/or high permeability media, anomalous properties of MTMs and the Electromagnetic Band Gap (EBG) structures and unexpected behavior of anisotropic structures are results of such interactions.

Accordingly, the radiation performance of the antenna as a kind of electromagnetic source beside dielectric materials could be significantly affected due to the microscopic polarization of such materials and one can readily take advantage of such effects to enhance the radiation efficiency of an ESA at low frequencies. In another word, an ESA is an RLC circuit with high resonant frequency. Combining the ESA with a dielectric material of specific electromagnetic properties and physical dimensions would create a resonator which possesses a resonance at the desired frequency band.

This chapter is concerned with the coating of an electrically short monopole antenna over an infinitely large ground plane. It should be noted that the monopole is fed by a small gap source, which should have an optimum height as discussed in the previous chapter. Invoking the modal analysis and far-field approximations, we can find out the radiation properties of the antenna coated with high permittivity or permeability and epsilon negative (ENG) materials. Two types of coating will be studied and it will be shown that both

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materials (DPS and ENG) could be employed to reduce the resonance frequency of the short monopole and enhance its properties at low frequencies.

4.2 Fully Coated Monopole Antenna

4.2.1 Formulation

Consider a dielectric (ϵ_r, μ_r) coated monopole antenna fed by a gap source over an infinitely large ground plane, as shown in Fig. 4.1. The monopole height is h and its radius is a . Due to the weak vertical radiation of the monopole, I have assumed a disc loaded PEC above the monopole at the distance D from the ground plane. Dividing the domain of interest into several regions, we can write the electromagnetic fields in each region. We can consider just a uniform electric field in region I, since the height of the gap is small compared to the wavelength of operation. So we have

$$E_z^I = \frac{1}{j\omega\epsilon_0} A' \quad (4.1)$$

Fig. 4.1. Schematic of the fully coated monopole fed by a gap source over an infinitely large ground

In region II and at the interface $\rho=a$, we have

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$$E_z^{II} = \frac{1}{j\omega\epsilon_0} \sum_{m=0}^{\infty} A_{2m} Z_{2m} \cos\left[\frac{m\pi(z-h)}{d}\right] \quad (4.2)$$

$$H_{\phi}^{II} = \sum_{m=0}^{\infty} A_{2m} \cos\left[\frac{m\pi(z-h)}{d}\right] \frac{\epsilon_m}{d} \quad (4.3)$$

where Z_{2m} and ϵ_m are the same functions as what mentioned in the previous chapter. Combining the Bessel's functions linearly and normalizing them in region III, we have

$$E_z^{III} = \frac{1}{j\omega\epsilon} \sum_{n=0}^{\infty} \{A_{3n} U_{3n}(\rho) + B_{3n} V_{3n}(\rho)\} \cos\left(\frac{n\pi z}{D}\right) \frac{\epsilon_n}{D} \quad (4.4)$$

$$H_{\phi}^{III} = \sum_{n=0}^{\infty} \frac{-1}{\gamma_{3n}^2} \{A_{3n} U_{3n}'(\rho) + B_{3n} V_{3n}'(\rho)\} \cos\left(\frac{n\pi z}{D}\right) \frac{\epsilon_n}{D} \quad (4.5)$$

where $U_{3n}(\rho) = \frac{J_0(\gamma_{3n}\rho)Y_0(\gamma_{3n}a) - Y_0(\gamma_{3n}\rho)J_0(\gamma_{3n}a)}{J_0(\gamma_{3n}b)Y_0(\gamma_{3n}a) - Y_0(\gamma_{3n}b)J_0(\gamma_{3n}a)}$, $\gamma_{3n}^2 = k_0^2\mu_r\epsilon_r - \left(\frac{n\pi}{D}\right)^2$

and

$$V_{3n}(\rho) = \frac{J_0(\gamma_{3n}b)Y_0(\gamma_{3n}\rho) - Y_0(\gamma_{3n}b)J_0(\gamma_{3n}\rho)}{J_0(\gamma_{3n}b)Y_0(\gamma_{3n}a) - Y_0(\gamma_{3n}b)J_0(\gamma_{3n}a)}. \text{ And finally at the interface } \rho=b$$

in region IV, we have

$$E_z^{IV} = \frac{1}{j\omega\epsilon_0} \sum_{n=0}^{\infty} A_{4n} \cos\left(\frac{n\pi z}{D}\right) \frac{\epsilon_n}{D} \quad (4.6)$$

$$H_{\phi}^{IV} = \sum_{n=0}^{\infty} A_{4n} Y_{4n} \cos\left(\frac{n\pi z}{D}\right) \frac{\epsilon_n}{D} \quad (4.7)$$

where $Y_{4n} = \frac{H_1^{(2)}(\gamma_{4n}b)}{\gamma_{4n}H_0^{(2)}(\gamma_{4n}b)}$ and $\gamma_{4n}^2 = k_0^2 - \left(\frac{n\pi}{D}\right)^2$.

Satisfying the continuity of tangential components of the electromagnetic fields across the regional interfaces, we can obtain

$$A_3 = \epsilon_r A_4 \quad (4.8)$$

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$$A_3 = P_3 B_3 \quad (4.9)$$

$$B_3 = F_I A^I + FZ_2 A_2 \quad (4.10)$$

$$A_2 = F^T (W_3 A_3 + W_4 B_3) \quad (4.11)$$

$$\text{where } F_{nm} = \frac{nd^2 D}{\pi(n^2 d^2 - m^2 D^2)} \left\{ (-1)^m \sin \left[\frac{n\pi(h+d)}{D} \right] - \sin \left[\frac{n\pi h}{D} \right] \right\},$$

$$F_{In} = \epsilon_r \Delta \text{sinc} \left(\frac{n\pi\Delta}{D} \right), W_{3n} = \frac{2\epsilon_n}{\pi\gamma_{3n}^2 a D \left\{ Y_0(\gamma_{3n}a)J_0(\gamma_{3n}b) - Y_0(\gamma_{3n}b)J_0(\gamma_{3n}a) \right\}}$$

$$W_{4n} = \frac{\epsilon_n \left\{ Y_1(\gamma_{3n}a)J_0(\gamma_{3n}b) - J_1(\gamma_{3n}a)Y_0(\gamma_{3n}b) \right\}}{\gamma_{3n} D \left\{ Y_0(\gamma_{3n}b)J_0(\gamma_{3n}a) - Y_0(\gamma_{3n}b)J_0(\gamma_{3n}a) \right\}}$$

$$\text{and } P_{3n} = \frac{\frac{-\epsilon_r}{\gamma_{3n}^2} V'_{3n}(b)}{Y_{4n} + \left(\frac{\epsilon_r}{\gamma_{3n}^2} U'_{3n}(b) \right)}. \text{ The superscript } T \text{ shows the transpose of the}$$

matrix. Solving the above system of linear equations, we obtain

$$A_2 = F^T Y_{eq} B_3 \quad (4.12)$$

where $Y_{eq} = U_3 P_3 + U_4$. And then,

$$A_2 = Q_1 F_I A^I \quad (4.13)$$

$$B_3 = Q_2 F_I A^I \quad (4.14)$$

where $Q_1 = (I - F^T Y_{eq} FZ_2)^{-1} F^T Y_{eq}$, $Q_2 = FZ_2 Q_1 + I$ and I is the unitary matrix.

A_3 and A_4 can be found simply from (4.9) and (4.8), respectively. Finally, we have

$$Y_{in} = G_{in} + jB_{in} = \frac{2\pi a \omega \epsilon_0}{j\Delta A^I} \sum_{n=0}^{\infty} \frac{-\epsilon_n}{D\gamma_{3n}^2} \left(U'_{3n}(a)A_{3n} + V'_{3n}(a)B_{3n} \right) \cos \left(\frac{n\pi\Delta}{D} \right) \quad (4.15)$$

and also,

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$$I(z) = 2\pi a \sum_{n=0}^{\infty} \frac{-\mathcal{E}_n}{D \gamma_{3n}^2} \left\{ A_{3n} U'_{3n}(\rho) + B_{3n} V'_{3n}(\rho) \right\} \cos\left(\frac{n\pi z}{D}\right) \quad (4.16)$$

After finding the input admittance and current distribution along the monopole antenna, we can now find its directivity and radiation pattern. As it is known [126], in the far field we have

$$E_r \cong 0 \quad (4.17)$$

$$E_\theta \cong -\frac{j\beta e^{-j\beta r}}{4\pi r} (L_\phi + \eta N_\theta) \quad (4.18)$$

$$E_\phi \cong \frac{j\beta e^{-j\beta r}}{4\pi r} (L_\theta - \eta N_\phi) \quad (4.19)$$

$$H_r \cong 0 \quad (4.20)$$

$$H_\theta = -\frac{1}{\eta} E_\phi \quad (4.21)$$

$$H_\phi = \frac{1}{\eta} E_\theta \quad (4.22)$$

where,

$$N_\theta = \iint_{S'} \left[J_\rho \cos \theta \cos(\phi - \phi') + J_\phi \cos \theta \sin(\phi - \phi') - J_z \sin \theta \right] e^{j\beta \vec{r} \cdot \vec{a}_r} ds',$$

$$L_\theta = \iint_{S'} \left[M_\rho \cos \theta \cos(\phi - \phi') + M_\phi \cos \theta \sin(\phi - \phi') - M_z \sin \theta \right] e^{j\beta \vec{r} \cdot \vec{a}_r} ds',$$

$$N_\phi = \iint_{S'} \left[-J_\rho \sin(\phi - \phi') + J_\phi \cos(\phi - \phi') \right] e^{j\beta \vec{r} \cdot \vec{a}_r} ds',$$

$$L_\phi = \iint_{S'} \left[-M_\rho \sin(\phi - \phi') + M_\phi \cos(\phi - \phi') \right] e^{j\beta \vec{r} \cdot \vec{a}_r} ds'.$$

$\vec{J}$ and $\vec{M}$ are the equivalent electric and magnetic current densities, respectively. It should be noted that the prime terms are due to the source points. Solving the above integrals, we have

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$$\vec{\mathbf{E}}_{far\ field} = E_{\theta} \mathbf{a}_{\theta} \cong -\frac{j\beta e^{-j\beta r}}{4\pi r} (L_{\phi} + \eta N_{\theta}) \quad (4.23)$$

$$\vec{\mathbf{H}}_{far\ field} = H_{\phi} \mathbf{a}_{\phi} = \frac{1}{\eta} E_{\theta} \mathbf{a}_{\phi} \quad (4.24)$$

where

$$L_{\phi} = \frac{-4\pi b\beta}{\omega\epsilon_0} \cos\theta \sin(\beta D \cos\theta) J_1(\beta b \sin\theta) \sum_{n=0}^{\infty} A_{4n} \frac{\epsilon_n}{D} \frac{(-1)^n}{\left(\frac{n\pi}{D}\right)^2 - (\beta \cos\theta)^2}$$

and

$$N_{\theta} = 4\pi b\beta \sin\theta \cos\theta \sin(\beta D \cos\theta) J_0(\beta b \sin\theta) \sum_{n=0}^{\infty} A_{4n} Y_{4n} \frac{\epsilon_n}{D} \frac{(-1)^n}{\left(\frac{n\pi}{D}\right)^2 - (\beta \cos\theta)^2}$$

Then we can find the directivity as

$$D(\theta) = \frac{2|F(\theta)|_{\max}^2}{\int_0^{\frac{\pi}{2}} |F(\theta)|^2 \sin(\theta) d\theta} \quad (4.25)$$

$$\text{where } F(\theta) = \frac{1}{4\pi\beta b D} (L_{\phi} + \eta N_{\theta}) \text{ and } \{F(\theta)\}_{\max} = F\left(\theta = \frac{\pi}{2}\right).$$

4.2.2 Numerical Results

In this part, I present numerical results based on the modal analysis to see how materials can affect the monopole antenna's performance. First of all, in order to show the validity of my analysis, I need to examine its convergence behavior and verify its accuracy in comparison with the measured data. As it was noted in the previous chapter, the number of the modes in each region should be chosen properly in order to ensure the summations convergent [123]. The convergence of the antenna input admittance for a particular frequency with respect to the number of the modes in region III (N) is illustrated in Fig. 4.2. It is seen that the coated monopole input admittance will be properly

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convergent, if the number of modes is large enough. After the convergence is studied, my calculated results for the input admittance of a non-coated and a coated monopole is illustrated and compared with the measured data in [124] and [127] through Figs. 4.3 and 4.4, respectively. It is observed that my calculated results are in good agreement with the measured data, which validates the accuracy of my eigenmode analysis.

Fig. 4.2. Convergence of the antenna input admittance with respect to the number modes in region III N for (M (number of the modes in region) II = 80, $\lambda=600\text{mm}$, $a=0.001588\lambda$, $b=3.5a$, $\Delta=0.1a$ and $h=100\text{mm}$)

After the validation of my modal technique, I am going to study the effects of different coating materials on the performance of the monopole and show the antenna miniaturization by material coating.

The input admittance of the coated monopole for different positive permittivity and permeability values is shown in Figs. 4.5 and 4.6, respectively. It is seen that coating the antenna with high permittivity and/or permeability materials would reduce the resonance frequency of the antenna. This is due to the polarization of the coating material and effective distribution of multi-poles

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in the coating medium that make them contribute in radiation process effectively and enhance the radiation of the antenna into the far zone at lower frequencies. However, loading the antenna with high permittivity and/or permeability materials would lead to sharp resonances and mismatches to the desired transmission line characteristic impedance (say 50Ω) and hence narrow bandwidths. It is also seen that the high permeability materials are more effective than the high permittivity materials to reduce the resonance frequency, though they lead to sharper resonances and hence bigger mismatches to the feed line.

The effects of negative permittivity and permeability coating on the input admittance of the shown monopole are studied through Figs. 4.7 and 4.8. Due to their negative permittivity, ENG shells act like a capacitor with negative dielectric constant. Therefore, the ENG shells would offer inductive properties rather than capacitive. It is known that a short vertical monopole over an infinitely large ground plane exhibits a large capacitive input impedance. Accordingly, putting the ENG shell and the electrically small monopole together would create an LC combination, which produces resonances at low frequencies and compensates for the high capacitive impedance of the short monopole at low frequencies. And obviously, the smaller the absolute value of the negative epsilon or the bigger the radius of the shell (coating) is, the lower will be the resonance frequency of the structure combined of the ENG shell and the short monopole. This is in fact shown in Fig. 4.7.

On the other hand, a mu-negative shell would act as an inductor with negative permeability and hence it would offer capacitive properties. This is quite contrary to the ENG shell. Therefore, the magnetic permeability shell would increase the resonance frequency of the short monopole with no radiation improvement at low frequencies. As it can be observed in Fig. 4.8, the input susceptance of the mu-negative coated monopole would cross the frequency axis at frequencies higher than 2 GHz and hence the resonance frequency of the

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antenna would be increased. Obviously, it can be interpreted that mu-negative shells (MNG) would enhance the radiation properties of small loop antennas, which exhibits inductive properties at low frequencies. To have a better insight into the antenna coating, the current distribution of the coated monopole antenna for different coating materials are illustrated in Figs. 4.9-11.

During the previous paragraphs, it was shown that ENG and DPS materials are able to reduce the resonance frequency of the monopole antenna. However, both the materials lead to sharp resonances, big mismatches to the feed line and are applicable only in narrow band operations. The other thing about the ENG materials is their frequency dispersion, which can affect the performance of the antenna. Different models such as Drude model can be employed to take this fact into account and study its corresponding effects on the coated monopole performance. In this respect, a comprehensive study will be conducted in the next chapter and the comparison between the ENG and DPS loading will be made in more details. But before ending this section, I present three examples of highly reduced resonance frequency (from 1.77GHz to 385MHz) of a monopole through the usage of ENG and high permeability materials in Figs. 4.12-17.

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Fig. 4.3. Input admittance of the non-coated monopole antenna for ($a=0.9528\text{mm}$, $\Delta=0.1a$ and $h=150\text{mm}$)

Fig. 4.4. Input admittance of the coated monopole antenna for ($\lambda=500\text{mm}$, $a=4.35\text{mm}$, $b=6.35\text{mm}$, $\Delta=0.1a$, $\epsilon_r=15$ and $\mu_r=1$)

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(a)

(b)

Fig. 4.5. Input admittance of the fully coated monopole for different permittivity values ($a=0.9528\text{mm}$, $b=2.2560\text{mm}$, $\Delta=0.1a$, $h=42.3333\text{mm}$ and $\mu_r=1$)

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(a)

(b)

Fig. 4.6. Input admittance of the fully coated monopole for different permeability values ($a=0.9528\text{mm}$, $b=2.2560\text{mm}$, $\Delta=0.1a$, $h=42.3333\text{mm}$ and $\epsilon_r=1$)

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(a)

(b)

Fig. 4.7. Input admittance of the fully coated monopole for different negative permittivity values ($a=0.9528\text{mm}$, $b=2.2560\text{mm}$, $\Delta=0.1a$, $h=42.3333\text{mm}$ and $\mu_r=1$)

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(a)

(b)

Fig. 4.8. Input admittance of the fully coated monopole for different negative permeability values ($a=0.9528\text{mm}$, $b=2.2560\text{mm}$, $\Delta=0.1a$, $h=42.3333\text{mm}$ and $\epsilon_r=1$)

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Fig. 4.9. Normalized current distribution of the fully coated monopole for different permittivity values ($\lambda=600\text{mm}$, $a=0.9528\text{mm}$, $b=2.2560\text{mm}$, $\Delta=0.1a$, $h=42.3333\text{mm}$ and $\mu_r=1$)

Fig. 4.10. Normalized current distribution of the fully coated monopole for different permeability values ($\lambda=600\text{mm}$, $a=0.9528\text{mm}$, $b=2.2560\text{mm}$, $\Delta=0.1a$, $h=42.3333\text{mm}$ and $\epsilon_r=1$)

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Fig. 4.11. Normalized current distribution of the fully coated monopole for different negative permittivity values ($\lambda=600\text{mm}$, $a=0.9528\text{mm}$, $b=2.2560\text{mm}$, $\Delta=0.1a$, $h=42.3333\text{mm}$ and $\mu_r=1$)

Fig. 4.12. Input admittance of the monopole coated with a high permeability material ($a=0.9528\text{mm}$, $b=8.76\text{mm}$, $\Delta=0.1a$, $h=42.3333\text{mm}$, $\mu_r=25.28$ and $\epsilon_r=1$)

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Fig. 4.13. Normalized current distribution of the monopole coated with a high permeability material ($f=385\text{MHz}$, $a=0.9528\text{mm}$, $b=8.76\text{mm}$, $\Delta=0.1a$, $h=42.3333\text{mm}$ $\mu_r=25.28$ and $\epsilon_r=1$)

Fig. 4.14. Radiation pattern (directivity) of the monopole coated with a high permeability material ($f=385\text{MHz}$, $a=0.9528\text{mm}$, $b=8.76\text{mm}$, $\Delta=0.1a$, $h=42.3333\text{mm}$ $\mu_r=25.28$ and $\epsilon_r=1$)

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Fig. 4.15. Input admittance of the monopole coated with a negative permittivity material ($a=0.9528\text{mm}$, $b=8.76\text{mm}$, $\Delta=0.1a$, $h=42.3333\text{mm}$ $\mu_r=1$ and $\epsilon_r=-1.8$)

Fig. 4.16. Normalized current distribution of the monopole coated with a negative permittivity material ($f=385\text{MHz}$, $a=0.9528\text{mm}$, $b=8.76\text{mm}$, $\Delta=0.1a$, $h=42.3333\text{mm}$ $\mu_r=1$ and $\epsilon_r=-1.8$)

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Fig. 4.17. Radiation pattern (directivity) of the monopole coated with a negative permittivity material ($f=385\text{MHz}$, $a=0.9528\text{mm}$, $b=8.76\text{mm}$, $\Delta=0.1a$, $h=42.3333\text{mm}$, $\mu_r=1$ and $\epsilon_r=-1.8$)

It is seen through Figs. 4.12-4.17 that both the high permittivity and ENG materials are able to decrease the resonance frequency of the monopole antenna in a great deal (from 1.77GHz to 385MHz). The current distribution along the monopole in both cases is sinusoidal and the pattern of the antenna is remained unchanged in comparison with the normal quarter wavelength monopole antenna. However, the conductance curves are pretty sharp which may cause mismatches between antenna and the common feed lines.

4.3 Partially Coated Monopole Antenna

4.3.1 Formulation

Another geometry that I examine in this chapter is shown in Fig. 4.18. It is seen that this geometry is more practical in antenna miniaturization. This geometry is actually indicating a monopole antenna coated with a dielectric

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layer (ϵ_r, μ_r) of radius “ $b-a$ ” and limited height to l . The structure is divided into four regions among which region III is broken into two sub-regions. I use the same approach as before to study this structure and find its radiation parameters. However, the formulation will be, to some extent, different from that of the previous section.

Dividing the structure into several regions, I write the electromagnetic fields in terms of modal functions. The electromagnetic fields in regions I and II ($\rho=a$) are exactly of the same form as that of the fully coated monopole antenna.

Solving the Maxwell's equations for a partially filled cylindrical shield with two parallel PEC plates at its top and bottom, we can write the electromagnetic fields in region III as

Fig. 4.18. Geometry of the partially coated monopole fed by a gap source over infinite ground plane

$$E_z^{III} = \frac{1}{j\omega\epsilon_0} \sum_{n=0}^{\infty} [A_{3n} U_{3n}(\rho) + B_{3n} V_{3n}(\rho)] W_{3n}(z) \quad (4.26)$$

$$H_{\phi}^{III} = \sum_{n=0}^{\infty} \frac{-1}{\gamma_{3n}^2} [A_{3n} U'_{3n}(\rho) + B_{3n} V'_{3n}(\rho)] W_{3n}(z) Q_{3n}(z) \quad (4.27)$$

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$$\begin{cases} \left(\gamma_{3n}\right)^2 = k_0^2 - \left(k_{3n}^B\right)^2 = k_0^2 \mu_r \epsilon_r - \left(k_{3n}^A\right)^2 \\ k_{3n}^B \tan\left(k_{3n}^B t\right) = -\frac{k_{3n}^A}{\epsilon_r} \tan\left(k_{3n}^A l\right) \end{cases} \quad (4.28)$$

We can write the equation (4.31) in another form as follows,

$$\begin{cases} \left(k_{3n}^A\right)^2 = k_0^2 \left(\mu_r \epsilon_r - 1\right) + \left(k_{3n}^B\right)^2 \\ \epsilon_r k_{3n}^B \sin\left(k_{3n}^B t\right) \cos\left(k_{3n}^A l\right) + k_{3n}^A \sin\left(k_{3n}^A l\right) \cos\left(k_{3n}^B t\right) = 0 \end{cases} \quad (4.29)$$

If we let $x = k_{3n}^B$ and $ct = k_0^2 \left(\mu_r \epsilon_r - 1\right)$, we have

$$func = \epsilon_r \left(SRB \times CRA\right) + \left(CRB \times SRA\right) = 0 \quad (4.30)$$

$$\text{where, } CRB = \begin{cases} \cos(tx) & x^2 \geq 0 \\ \cosh(t|x|) & x^2 \leq 0 \end{cases}, SRB = \begin{cases} x \sin(tx) & x^2 \geq 0 \\ -x \sinh(t|x|) & x^2 \leq 0 \end{cases}$$

$$CRA = \begin{cases} \cos\left(l\sqrt{ct+x^2}\right) & ct+x^2 \geq 0 \\ \cosh\left(l\sqrt{ct+x^2}\right) & ct+x^2 \leq 0 \end{cases} \text{ and}$$

$$SRA = \begin{cases} \sqrt{ct+x^2} \sin\left(l\sqrt{ct+x^2}\right) & ct+x^2 \geq 0 \\ -\sqrt{ct+x^2} \sinh\left(l\sqrt{ct+x^2}\right) & ct+x^2 \leq 0 \end{cases}. \text{ The modal fields in region}$$

IV ($\rho=b$) is similar to that of the fully coated monopole, as

$$E_z^{IV} = \frac{1}{j\omega\epsilon_0} \sum_{q=0}^{\infty} A_{4q} \cos\left(\frac{q\pi z}{D}\right) \frac{\epsilon_q}{D} \quad (4.31)$$

$$H_\phi^{IV} = \sum_{q=0}^{\infty} A_{4q} Y_{4q} \cos\left(\frac{q\pi z}{D}\right) \quad (4.32)$$

$$\text{where } Y_{4q} = \frac{\epsilon_q H_1^{(2)}(\gamma_{4q} b)}{D \gamma_{4q} H_0^{(2)}(\gamma_{4q} b)} \quad \text{and} \quad \gamma_{4q}^2 = k_0^2 - \left(\frac{q\pi}{D}\right)^2. \quad \text{Applying the}$$

electromagnetic field boundary conditions, we have

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I) At $\rho=a$, $0 < z < D$

$$E_z^{III} = \begin{cases} E_z^I & 0 < z < \Delta \\ E_z^{II} & h < z < h + d \\ 0 & elsewhere \end{cases} \quad (4.33)$$

or,

$$B_{3n} \int_0^D W_{3n}(z) W_{3n}(z) Q_{3n}(z) dz = A^I \int_0^\Delta W_{3n}(z) Q_{3n}(z) dz + \times \sum_{m=0}^{\infty} A_{2m} Z_{2m} \int_h^{h+d} W_{3n}(z) Q_{3n}(z) \cos \left[\frac{m\pi(z-h)}{d} \right] dz \quad (4.34)$$

which leads to,

$$B_{3n} = C_n G_n A^I + C_n \sum_{m=0}^{\infty} A_{2m} Z_{2m} P_{nm} \quad (4.35)$$

$$\text{where } G_n = \Delta a_n \text{sinc}(k_{3n}^A \Delta), \frac{1}{C_n} = \frac{a_n^2 l}{2\epsilon_r} \left[\text{sinc}(2k_{3n}^A l) + 1 \right] + \frac{b_n^2 t}{2} \left[\text{sinc}(2k_{3n}^B t) + 1 \right]$$

$$a_n = \frac{1}{\cos(k_{3n}^A l)}, \text{ and also } P_{nm} = S_1 + S_2 \text{ with}$$

$$S_1 = \begin{cases} \frac{a_n}{(k_{3n}^A)^2 - (\frac{m\pi}{d})^2} \vartheta_{nm}^I & k_{3n}^A \neq \frac{m\pi}{d} \\ \frac{a_n}{2} \vartheta_{nm}^{II} & k_{3n}^A = \frac{m\pi}{d} \neq 0 \\ a_n \xi & k_{3n}^A = \frac{m\pi}{d} = 0 \end{cases}, S_2 = \begin{cases} \frac{b_n}{(k_{3n}^B)^2 - (\frac{m\pi}{d})^2} \vartheta_{nm}^{III} & k_{3n}^B \neq \frac{m\pi}{d} \\ \frac{b_n}{2} \vartheta_{nm}^{IV} & k_{3n}^B = \frac{m\pi}{d} \neq 0 \\ b_n (d - \xi) & k_{3n}^B = \frac{m\pi}{d} = 0 \end{cases}$$

$$\vartheta_{nm}^I = \left\{ \cos(k_{3n}^A l) \cos\left(\frac{m\pi}{d} \xi\right) \left(k_{3n}^A \tan(k_{3n}^A l) - \frac{m\pi}{d} \tan\left(\frac{m\pi}{d} \xi\right) \right) - k_{3n}^A \sin(k_{3n}^A h) \right\},$$

$$\vartheta_{nm}^{II} = \frac{\sin\left[(m\pi/d)(h + 2\xi)\right] - \sin(m\pi h/d)}{(2m\pi/d)} + \xi \cos(m\pi h/d),$$

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$$\begin{aligned} \vartheta_{nm}^{III} &= (-1)^m k_{3n}^B \sin \left[k_{3n}^B (h + d - D) \right] + \cos(k_{3n}^B t) \cos\left(\frac{m\pi}{d} \xi\right) \\ &\quad \times \left[\frac{m\pi}{d} \tan\left(\frac{m\pi}{d} \xi\right) + k_{3n}^B \tan(k_{3n}^B t) \right], \\ \vartheta_{nm}^{IV} &= \frac{\sin \left[\frac{m\pi(h + 2d - D)}{d} \right] - \sin \left[\frac{m\pi(h + 2\xi - D)}{d} \right]}{\frac{2m\pi}{d}} + (d - \xi) \cos \left[\frac{m\pi(D - h)}{d} \right] \end{aligned}$$

II) At $\rho=b$, $0 < z < D$

$$E_z^{IV} = E_z^{III} \quad (4.36)$$

Then

$$A_{4q} \int_0^D \cos\left(\frac{q\pi z}{D}\right) \cos\left(\frac{q\pi z}{D}\right) \frac{\varepsilon_q}{D} dz = \sum_{n=0}^{\infty} A_{3n} \int_0^D W_{3n}(z) \cos\left[\frac{q\pi z}{D}\right] dz \quad (4.37)$$

which gives

$$A_{4q} = \sum_{n=0}^{\infty} A_{3n} F_{qn} \quad (4.38)$$

where $F_{nq} = S_3 + S_4$,

$$S_3 = \begin{cases} \frac{a_n l}{2\varepsilon_r} \left\{ \operatorname{sinc} \left[\left(\frac{q\pi}{D} + k_{3n}^A \right) l \right] + \operatorname{sinc} \left[\left(\frac{q\pi}{D} - k_{3n}^A \right) l \right] \right\} & k_{3n}^A \neq \frac{q\pi}{D} \\ \frac{a_n l}{2\varepsilon_r} \left\{ \operatorname{sinc} \left(\frac{2q\pi l}{D} \right) + 1 \right\} & k_{3n}^A = \frac{q\pi}{D} \neq 0 \\ \frac{a_n l}{\varepsilon_r} & k_{3n}^A = \frac{q\pi}{D} = 0 \end{cases} \quad \text{and}$$

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$$S_4 = \begin{cases} \frac{(-1)^q b_n t}{2} \left\{ \text{sinc} \left[\left(\frac{q\pi}{D} + k_{3n}^B \right) t \right] + \text{sinc} \left[\left(\frac{q\pi}{D} - k_{3n}^B \right) t \right] \right\} & k_{3n}^B \neq \frac{q\pi}{D} \\ \frac{(-1)^q b_n t}{2} \left\{ 1 + \text{sinc} \left(\frac{2q\pi t}{D} \right) \right\} & k_{3n}^B = \frac{q\pi}{D} \neq 0 \\ b_n t & k_{3n}^B = \frac{q\pi}{D} = 0 \end{cases}.$$

III) At $\rho=b$, $h < z < h+d$

$$H_\phi^{II} = H_\phi^{III} \quad (4.39)$$

So we have,

$$A_{2m} \frac{\epsilon_m}{d} \int_h^{h+d} \cos^2 \left[\frac{m\pi(z-h)}{d} \right] dz = \sum_{n=0}^{\infty} \frac{-1}{\gamma_{3n}^2} \left[A_{3n} U_{3n}'(a) + B_{3n} V_{3n}'(a) \right] \times \int_h^{h+d} W_{3n}(z) Q_{3n}(z) \cos \left[\frac{m\pi(z-h)}{d} \right] dz \quad (4.40)$$

It gives,

$$A_{2m} = \sum_{n=0}^{\infty} \frac{-1}{\gamma_{3n}^2} \left[A_{3n} U_{3n}'(a) + B_{3n} V_{3n}'(a) \right] D_{mn} \quad (4.41)$$

IV) At $\rho=b$, $0 < z < D$

$$H_\phi^{III} = H_\phi^{IV} \quad (4.42).$$

Similarly, it leads to

$$\frac{-1}{\gamma_{3n}^2} \left[A_{3n} U_{3n}'(b) + B_{3n} V_{3n}'(b) \right] = C_n \sum_{q=0}^{\infty} A_{4q} Y_{4q} F_{nq} \quad (4.43)$$

We can briefly express the above equations in matrix forms,

$$B_3 = CGA^I + CPZ_2 A_2 \quad (4.44)$$

$$A_4 = FA_3 \quad (4.45)$$

$$A_2 = P^T (W_1 A_3 + W_2 B_3) \quad (4.46)$$

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$$W_3 A_3 + W_4 B_3 = CF^T Y_4 A_4 \quad (4.47)$$

where $W_{1n} = \frac{-1}{\gamma_{3n}^2} U'_{3n}(a)$, $W_{2n} = \frac{-1}{\gamma_{3n}^2} V'_{3n}(a)$, $W_{3n} = \frac{-1}{\gamma_{3n}^2} U'_{3n}(b)$ and

$W_{4n} = \frac{-1}{\gamma_{3n}^2} V'_{3n}(b)$. These equations lead to

$$A_3 = S^{-1} C G A^I \quad (4.48)$$

$$B_3 = T_1 A_3 \quad (4.49)$$

where $S = T_1 - C P Z_2 T_2$, $T_2 = P^T (W_1 + W_2 T_1)$ and $T_1 = W_4^{-1} (C F^T Y_4 F - W_3)$.

We also know that,

$$V_{in} = \frac{j\Delta}{\omega\epsilon_0} A^I \rightarrow A^I = \frac{\omega\epsilon_0}{j\Delta} V_{in} \quad (4.50)$$

Now, we are able to find the input admittance and the current distribution of the antenna;

$$I_{in} = 2\pi a H_\phi^{III}(z = \Delta) = 2\pi a \sum_{n=0}^{\infty} \frac{-1}{\gamma_{3n}^2} [A_{3n} U'_{3n}(a) + B_{3n} V'_{3n}(a)] W_{3n}(\Delta) Q_{3n}(\Delta) \quad (4.51)$$

$$Y_{in} = G_{in} + jB_{in} = \frac{I_{in}}{V_{in}} \quad (4.52)$$

4.3.2 Numerical Results

In this section, I am to present my results for the input admittance and the current distribution of the coated monopole shown in Fig. 4.18. At the first step, I illustrate my computed results for the input admittance of the monopole coated with Teflon ($\epsilon_r = 2.08$, $\mu_r = 1$) and compare it with the results obtained from the High Frequency Structure Simulator (HFSS). It is shown in Fig. 4.19.

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It is seen that my results and HFSS results for the Teflon coated monopole antenna are in good agreement, though this agreement for the input susceptance is reduced at higher frequencies. The causes of this reduced agreement at high frequencies are the finite size of the ground plane, existence of objects with very different sizes, limited amount of memory in HFSS simulation and also sensitivity of input susceptance of the monopole to the limited number of modes.

As I mentioned before, we need to truncate the number of the modes in regions II, III and IV, so that the summations in the above formulation have finite terms. It should be noted that these truncated numbers must be sufficiently large to make the summations convergent [123]. This fact is evidently examined in Fig. 4.20, where the convergence of the input admittance of the monopole is illustrated versus the number of the modes in region III (which is equal to the number of the modes in region IV). It is seen that the input admittance of the monopole is pretty convergent for the values of $N \geq 200$.

In the previous section, it was shown that coating the monopole antenna with DPS or ENG materials would lead the coating material to participate in radiation and enhances the antenna performance at low frequencies. It seems that the performance of partially coated monopole shown in Fig. 4.18 is similar to that of the fully coated monopole antenna as long as $a, b \ll h$ and $l \geq h$. This is because the coating layer acts like a dielectric resonator that its radius is more dominant than its height to determine the resonance frequency. Avoiding the repeating results and discussions, I just present an example of miniaturized monopole antenna (similar to what was shown at the end of the previous section) coated with a high permeability material. As it was mentioned, it is seen that the miniaturized monopole would exhibit a very similar input admittance and current distribution to that of the miniaturized monopole studied through Figs. 4.12-14.

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Fig. 4.19. Input admittance of the shown monopole in Fig. 4.18 with a Teflon coating layer ($\epsilon_r=2.08$, $\mu_r=1$, $a=1.588\text{mm}$, $b=2.588\text{mm}$, $h=150\text{mm}$ and $l=2h$)

Fig. 4.20. Convergence of the antenna input admittance with respect to the number modes in region III N (=number of the modes in region IV) for (M (number of the modes in region II) =80, $\lambda=750\text{mm}$, $a=1.588\text{mm}$, $b=2.588\text{mm}$, $l=2h$, $\Delta=0.1a$ and $h=150\text{mm}$)

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Fig. 4.21. Input admittance of the shown monopole in Fig. 4.18 with a high permeability coating layer ($\epsilon_r=1$, $\mu_r=25.28$, $a=0.9528\text{mm}$, $b=8.76\text{mm}$, $h=42.33\text{mm}$ and $l=1.1h$)

Fig. 4.22. Current distribution of the shown monopole in Fig. 4.18 with a high permeability coating layer ($\epsilon_r=1$, $\mu_r=25.28$, $a=0.9528\text{mm}$, $b=8.76\text{mm}$, $h=42.33\text{mm}$ and $l=1.1h$)

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4.4 Conclusion

Employing modal analysis, two types of coated monopoles fed by the gap source were studied and it was shown that coating the vertical monopole with DPS or ENG materials would lead the electrically short monopole to offer lower resonance frequencies and improved radiation into the far zone at lower frequencies. It was shown that both the structures (fully and partially coated monopoles) would have similar properties due to the dominance of the coating layer radius in determining the resonance frequency rather than its height as long as $a, b \ll h$ and $l \geq h$.

CHAPTER 5

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5.1 Introduction

As it was mentioned in the literature review, we can divide the miniaturization techniques into two major categories. The first one is designing physically small structures, which can exhibit large effective areas [1]. The second one is loading the antenna with materials that can effectively contribute in radiation process [46].

It is known that a single short vertical monopole has a highly triangular current distribution that causes a strongly stored near field around the antenna and a very poor radiation into the far-zone. In the equivalent circuit model, we can model the antenna with a very huge reactance and a small resistance in which the delivered power to the resistance is poor at low frequencies. So the radiation quality factor of the antenna jumps up to big numbers and also its radiation efficiency collapses down at radio frequencies. And this is just due to the high resonance frequency of the antenna.

Reducing the resonance frequency, we need to add inductive or capacitive elements in series or parallel with the antenna, respectively. Adding inductive coils to a dipole arms [128] and using end wires [104] are examples to improve the input impedance characteristics of an electrically short dipole or monopole.

As I mentioned before, an electrically small monopole has a triangular current distribution. Making the current distribution along the monopole more uniform, we increase the area under the current distribution curve and hence increase the radiated power at low frequencies. We can make the current distribution more uniform by ending the monopole with a capacitive loading such as a top-hat or radial rods. That is because the current distribution does not need to be zero at the end of monopole and is able to flow along the radial arms

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or disks. So loading the monopole with a radial disk at its end, we can produce resonances at lower frequencies and improve the radiation characteristics of the antenna at such frequencies.

Loading the antenna with novel materials such as MTMs or magneto-dielectrics is another method to reduce its resonance frequency and hence decrease its physical sizes. In fact, effective material loading provides the antenna with supplementary channels to radiate energy [46]. In [46], several general methods to utilize the material loading proficiently have been concluded. Using dielectric shells of resonant dimensions around the antenna is one of the methods in [46], which can provide an efficient coupling between the guided modes and spherical radiating modes. The dielectric resonant shells are even of more interests while they are composed of new artificial materials such as negative permittivity and/or negative permeability materials.

In this chapter, taking advantage of both mentioned miniaturization techniques, I am to combine the structure bending method with the material loading procedure to significantly miniaturize a monopole antenna over an infinitely large ground plane. I am actually going to observe the resonance behavior of a loaded top-hat monopole antenna and obtain lower frequency resonances for a short monopole. I first invoke a modal approach to find the input admittance or impedance of a loaded top-hat monopole fed by a small gap source over an infinitely large ground plane. After ascertaining the convergence behavior of the numerical approach and also verifying its accuracy by comparison with available experimental results, I study the effect of material loading in addition to the disc loading on the resonance behavior of the monopole antenna and its radiation pattern. As it was mentioned in the literature review, MTM is nowadays one of the hottest topics in electromagnetic field theory which is finding applications in various areas of science and engineering such as antenna miniaturization. As a matter of fact, I specifically emphasize on the comparison of ENG materials with high positive permittivity and/or permeability to examine their promising capabilities in the antenna size

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reduction and observe how they may affect the far-field radiation of the antenna. I end with a couple of explanations to provide the reader with a physical insight into the phenomena and a more clear understanding on the related facts. Finally, a few conclusions will be drawn.

5.2 Modal Analysis

5.2.1 Input Admittance and Current Distribution

Consider a top-hat monopole of radius a and height h , as shown in Fig. 5.1. The radius of the top-hat is b and its thickness is t . This monopole is fed by a small gap source over an infinitely large ground plane. I have assumed the same artificial perfect conductor as the previous chapters above the antenna at a distance D above the ground plane. The monopole is loaded with a dielectric material of relative permittivity ϵ_r and relative permeability μ_r .

Then, we can divide the structure of interest into several regions and write the electromagnetic fields in terms of time-harmonic modal functions weighted with unknown coefficients (A^I, A_2, A_3, B_3 and A_4). Following the same procedure as that of the previous chapter, we can obtain the input admittance and current distribution of the antenna. Avoiding repeating formulation, I just write the final results for the mentioned parameters. The input admittance of the antenna would be

$$Y_{in} = G_{in} + jB_{in} = \frac{2\pi a \omega \epsilon_0}{j\Delta A^I} \sum_{n=0}^N \frac{-1}{\gamma_{3n}^2} \{U'_{3n}(a)A_{3n} + V'_{3n}(a)B_{3n}\} \cos\left(\frac{n\pi\Delta}{h}\right) \frac{\epsilon_n}{h} \quad (5.1)$$

where U_{3n} and V_{3n} were defined in the previous chapter.

The current distribution of the antenna would be obtained from the tangential components of the H -fields. Thus the current distribution along the vertical stem of the antenna is

$$I(z) = 2\pi a H_\phi^{III}(a) \quad (5.2)$$

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Fig. 5.1. Geometry of a loaded top-hat monopole.

which gives

$$I(z) = 2\pi a \sum_{n=0}^N \frac{-1}{\gamma_{3n}^2} \{U_{3n}'(a)A_{3n} + V_{3n}'(a)B_{3n}\} \cos\left(\frac{n\pi z}{h}\right) \frac{\epsilon_n}{h} \quad (5.3)$$

Similarly, the radial current along the upper and lower surfaces of the top-hat would be

$$I_{upper}(\rho) = \sum_{m=0}^M A_{2m} \frac{(-1)^m \epsilon_m \gamma_{2m} J_1(\gamma_{2m} \rho)}{dJ_1(\gamma_{2m} b)} \quad (5.4)$$

$$I_{lower}(\rho) = \sum_{n=0}^N \frac{(-1)^{n+1}}{\gamma_{3n}^2} \{A_{3n} U_{3n}'(\rho) + B_{3n} V_{3n}'(\rho)\} \frac{\epsilon_n}{h} \quad (5.5)$$

5.2.2 Radiation Pattern

In order to find the far-zone radiated electromagnetic fields, we need to enclose the loaded top-hat monopole and its image into an artificial volume. This volume is shown in Fig. 5.2. Considering physical and equivalent electric and magnetic currents over the outer surfaces of the shown volume, we can obtain the far-zone electromagnetic fields being radiated into the unbounded free space. It should be mentioned that we follow the same approach as that of [126] to find the far-zone radiated fields due to an electromagnetic source.

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Fig. 5.2. Schematic of the artificial enclosing volume around the antenna.

I am to derive the far-fields in three steps. First, based on the induction theorem [126], I consider the equivalent electric and magnetic currents over the boundary between regions III and IV.

$$\overline{\mathbf{M}}_{eq} = -\hat{\mathbf{a}}_\rho \times \hat{\mathbf{a}}_z E_z^{III} = \hat{\mathbf{a}}_\phi E_z^{III} \quad (5.6)$$

$$\overline{\mathbf{J}}_{eq} = \hat{\mathbf{a}}_\rho \times \hat{\mathbf{a}}_\phi H_\phi^{III} = \hat{\mathbf{a}}_z H_\phi^{III} \quad (5.7)$$

Then on the basis of equations (4.18-22), we have

$$N_{\theta 1} = -4\pi b k_0 \cos \theta \sin \theta \sin(k_0 h \cos \theta) J_0(k_0 b \sin \theta) \times \sum_{n=0}^N (U_{3n} A_{3n} + V_{3n} B_{3n}) \frac{\epsilon_n}{h} \frac{(-1)^n}{(k_0 \cos \theta)^2 - \left(\frac{n\pi}{h}\right)^2} \quad (5.8)$$

$$L_{\phi 1} = \frac{4\pi b}{\omega \epsilon} k_0 \cos \theta \sin(k_0 h \cos \theta) J_1(k_0 b \sin \theta) \times \sum_{n=0}^N A_{3n} \frac{\epsilon_n}{h} \frac{(-1)^n}{(k_0 \cos \theta)^2 - \left(\frac{n\pi}{h}\right)^2} \quad (5.9)$$

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Hence, we obtain

$$E_{\theta 1} \cong \frac{-jk_0 e^{-jk_0 r}}{4\pi r} (L_{\phi 1} + \eta N_{\theta 1}) \quad (5.10)$$

$$H_{\phi 1} \cong \frac{1}{\eta} E_{\theta 1} \quad (5.11)$$

where $\eta = \sqrt{\frac{\mu_0}{\epsilon_0}} \cong 120\pi$. Other components of the far-zone radiated fields are approximately zero.

In the second step, I consider the physical electric current flowing on the perimeter of the top-hat. This physical current can be found as

$$\vec{\mathbf{J}} = \vec{\mathbf{a}}_\rho \times \vec{\mathbf{a}}_\phi H_\phi^{IV} = \vec{\mathbf{a}}_z H_\phi^{IV} \Big|_{h < z < h+t} \quad (5.12)$$

Then, taking the same integration done before, we have

$$\begin{aligned} N_{\theta 2} = & -2\pi b \sin \theta J_0(k_0 b \sin \theta) \sum_{p=0}^P A_{4p} Y_{4p} \times \\ & \left\{ (h+t) \left(\text{sinc} \left[\left(\frac{p\pi}{D} + k_0 \cos \theta \right) (h+t) \right] + \text{sinc} \left[\left(\frac{p\pi}{D} - k_0 \cos \theta \right) (h+t) \right] \right) - \right. \\ & \left. h \left(\text{sinc} \left[\left(\frac{p\pi}{D} + k_0 \cos \theta \right) h \right] + \text{sinc} \left[\left(\frac{p\pi}{D} - k_0 \cos \theta \right) h \right] \right) \right\} \end{aligned} \quad (5.13)$$

where Y_{4p} is similar to that defined in chapter four. And similarly, we have

$$E_{\theta 2} \cong \frac{-jk_0 e^{-jk_0 r}}{4\pi r} \eta N_{\theta 2} \quad (5.14)$$

$$H_{\phi 2} \cong \frac{1}{\eta} E_{\theta 2} \quad (5.15)$$

Finally, I consider the physical electric current flowing radially over the outer surface of the top-hat. With regard to the tangential component of the H -field in region II, we know

$$\vec{\mathbf{J}} = \vec{\mathbf{a}}_z \times \vec{\mathbf{a}}_\phi H_\phi^{II} = \vec{\mathbf{a}}_\rho H_\phi^{II} \Big|_{z=h+t} \quad (5.16)$$

Hence we have

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$$N_{\theta 3} = 4\pi \cos \theta \sin [k_0 (h+t) \cos \theta] \sum_{m=0}^M \frac{\varepsilon_m}{dJ_1(\gamma_{2m}b)} A_{2m} \times \int_0^b J_1(k_0 \rho' \sin \theta) J_1(\gamma_{2m} \rho') \rho' d\rho' \quad (5.17)$$

And then,

$$E_{\theta 3} \cong \frac{-jk_0 e^{-jk_0 r}}{4\pi r} \eta N_{\theta 3} \quad (5.18)$$

$$H_{\phi 3} \cong \frac{1}{\eta} E_{\theta 3} \quad (5.19)$$

Summation of the obtained fields gives the total far-zone radiated field by the loaded top-hat monopole antenna. So we have

$$E_{\theta} = E_{\theta 1} + E_{\theta 2} + E_{\theta 3} = \frac{-jk_0 e^{-jk_0 r}}{4\pi r} [L_{\phi 1} + \eta (N_{\theta 1} + N_{\theta 2} + N_{\theta 3})] \quad (5.20)$$

$$H_{\phi} = H_{\phi 1} + H_{\phi 2} + H_{\phi 3} = \frac{-jk_0 e^{-jk_0 r}}{4\pi r} (\eta^{-1} L_{\phi 1} + N_{\theta 1} + N_{\theta 2} + N_{\theta 3}) \quad (5.21)$$

5.3 Numerical Results and Discussions

5.3.1 Convergence and Validity of the Modal Analysis

First of all, I examine the convergence behavior of the modal analysis. It is well understood that in order to achieve an acceptable convergence, we need to truncate number of the modes in each region to a proper value [123]. Based on the criterion for the modal analysis [123], we have

$$\frac{P}{D} = \frac{N}{h} = \frac{M}{d} \quad (5.22)$$

And then, we can write

$$N + M = \left(\frac{h}{D} + \frac{d}{D} \right) P \cong P \quad (5.23)$$

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It is noted that the previous equation would be valid as long as t is negligible in comparison with h . Table 5.1 shows the convergence of input admittance of the unloaded top-hat monopole ($\mu_r = \epsilon_r = 1$) with respect to the number of modes considered in regions II and III.

TABLE 5.1. Convergence of input admittance of the unloaded top-hat monopole for ($a=0.9528$ mm, $b=2.7534$ mm, $h=150$ mm, $t=0.9623$ mm, $\Delta=0.09528$ mm and $\mu_r = \epsilon_r = 1$)

M	N	Input conductance (mhos)	Input susceptance (mhos)
80	10	0.001367	0.001089
80	20	0.001376	0.001248
80	40	0.001373	0.001487
80	80	0.001370	0.001733
80	160	0.001366	0.001948
80	300	0.001363	0.002097
160	300	0.001372	0.002042

It is observed that for $M = 80$ and $N = 300$, we have an acceptable convergence for the input admittance of the top-hat monopole.

As it was mentioned in the previous chapters, the assumed artificial conductive wall should be far enough from the ground plane, though the radiation of the top-hat monopole is null in the vertical direction. During my computations, it was noticed that $D = 3.24\lambda$ is quite enough to have a good convergence, where λ is the wavelength in free space.

To verify the accuracy of my approach, I compare my results for the input admittance of the unloaded top-hat monopole with those of [105, 124]. Shown in Fig. 5.3 is the input admittance of an unloaded monopole with a very small top-hat. This antenna can actually be compared with a common monopole without a top-hat, which was measured by King [124]. A good agreement between my results and the measured ones is observed in this figure.

Other comparisons are made in Figs. 5.4 and 5.5, where my results for the input impedance of unloaded and loaded top-hat monopoles are illustrated against the experimental results reported in [105], respectively. The agreement between my computed results and measured ones is good, which validates the

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accuracy of my modal analysis approach. Finally, my computed results for $|rE_\theta|$ pattern and radiation efficiency are compared with the measured data in [105] and shown in Fig. 5.6 and 5.7, respectively. The good agreement between the shown results certifies my far-field formulation.

5.3.2 Unloaded Top-Hat Monopole

As it was mentioned earlier, a short monopole antenna has a sharp triangular current distribution that leads to poor radiation into the far zone. By adding capacitive elements such as ending discs, this current distribution would become more uniform. Acquiring a more uniform current distribution, the short monopole would be resonant at lower frequencies and hence radiate more energy. This fact is simply shown in Fig. 5.8, where the values of top-hat radius b that make the antenna of height h and radius a resonant at the wavelength λ are depicted. It is observed that by increasing the top-hat radius, the resonance frequency of the antenna would decrease.

Fig. 5.3. The input admittance of an unloaded monopole with a very small top-hat ($a=0.9528$ mm, $b=0.9534$ mm, $h=150$ mm, $t=0.9623$ mm, $\Delta=0.09528$ mm and $\mu_r = \epsilon_r = 1$)

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Fig. 5.4. The input impedance of an unloaded top-hat monopole with ($a=1.19$ mm, $b=3.11$ mm, $h=31.75$ mm, $t=0.0119$ mm, $\Delta=0.119$ mm and $\mu_r = \epsilon_r = 1$)

Fig. 5.5. The input impedance of a loaded top-hat monopole with ($a=1.19$ mm, $b=38.7$ mm, $h=31.75$ mm, $t=0.0119$ mm, $\Delta=0.119$ mm, $\sigma=0.01$ S/m, $\epsilon_r=2.68$ and $\mu_r=1$)

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Fig. 5.6. $|rE_\theta|$ of an unloaded top-hat monopole with ($a=1.19\text{mm}$, $b=38.7\text{mm}$, $h=31.75\text{mm}$, $t=0.0119\text{mm}$, $\Delta=0.119\text{mm}$, $f=4\text{GHz}$ and $\epsilon_r, \mu_r=1$)

Fig. 5.7. The radiation efficiency of a loaded top-hat monopole with ($a=1.19\text{mm}$, $b=38.7\text{mm}$, $h=31.75\text{mm}$, $t=0.0119\text{mm}$, $\Delta=0.119\text{mm}$, $\epsilon_r=2.68$ and $\mu_r=1$)

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Fig. 5.8. Optimum values for b to make the unloaded top-hat monopole of height h resonant ($t=0.9623$ mm and $\lambda=0.09528$ mm)

Current distribution of the monopole is the next parameter that I examine. As I mentioned before, the major idea behind the top-hat loading is to produce a capacitive loading, which reduces the capacitive reactance of a short monopole. At low frequencies, a short vertical monopole has a triangular current distribution with small area under its curve that is related to the radiated power. Adding a radial disc to the top end of the monopole, we can make the current distribution along the vertical stem more uniform, since the current does not need to disappear at the end of monopole and can flow along the radial disc. The larger is the top-hat, the further would be the null of current distribution from the end of monopole and hence the more uniform would be the current distribution along the vertical stem. Making the current distribution more uniform, we would increase the area under its curve and hence we would increase the radiated power of the antenna. This fact is illustrated in Fig. 5.9, where the current distribution of an unloaded top-hat monopole antenna is illustrated versus its vertical stem height for different b .

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It is observed that the current distribution becomes more uniform as we increase the radius of the top-hat. It seems that the magnitude of current at the end of the vertical stem increases as we widen the radial disk. This fact can also be notified in Fig. 5.10, where the current distribution at the top of vertical stem versus the ratio of the top-hat radius to the stem radius (b/a) is shown. We see that increasing the radius of top-hat of a special height, the current magnitude at $z = h$ increases. It is also observed that this increase is faster for higher monopoles. The interesting point that should be noticed is the independence of the mentioned current upon the wavelength of operation.

The next thing that I take into account is the radiation pattern and also the radiated power of the unloaded top-hat monopole. Shown in Figs. 5.11 and 5.12 are the directivity patterns of the antenna with regard to its height and top-hat radius, respectively. It is seen that increasing radius of the top-hat does not have any significant effect on the directivity pattern of the antenna. To have a more clear idea on the radiated power, I indicate this parameter in Fig. 5.13 where $|rE_\theta|$ is given for different top-hat radiuses. It is observed that $|rE_\theta|$ increases as the top-hat gets wider and so does the radiated power density.

Finally, based on the data given in Fig. 5.8, I compute the input admittance of an unloaded top-hat monopole of height 10 mm and $a=0.9528$ mm which is desired to be resonating at $f=2.5$ GHz ($h/\lambda=0.083$, $\lambda=120$ mm). It is observed that if $b=8.2a$, we will have the resonance at that frequency. Using these values, I compute the input admittance of the mentioned monopole and indicate it in Fig. 5.14. As we see, adding such a top-hat to a short monopole decreases the resonance frequency from 7.68 GHz to about 2.5 GHz.

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Fig. 5.9. Current distribution along the vertical stem of the unloaded top-hat monopole for different b ($a=0.9528$ mm, $\lambda=100$ mm, $h=\lambda/20$, $t \cong 0$ and $\Delta=0.09528$ mm)

Fig. 5.10. Magnitude of current at $z=h$ with respect to b/a for different heights ($a=0.001588\lambda$, $t \cong 0$ and $\Delta=0.1a$)

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(a) $h = \lambda / 20$

(b) $h = \lambda / 10$

(c) $h = \lambda / 4$

(d) $h = \lambda / 2$

(e) $h = \lambda$

(f) $h = 2\lambda$

Fig. 5.11. Directivity patterns of the unloaded top-hat monopole for different heights ($\lambda=100$ mm, $a=0.001588 \lambda$, $b=0.4h$, $t \cong a$ and $\Delta=0.1a$)

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(a) $b = 0.1h$

(b) $b = 0.2h$

(c) $b = 0.4h$

(d) $b = 0.8h$

(e) $b = 1.6h$

(f) $b = 3.2h$

Fig. 5.12. Directivity patterns of the unloaded top-hat monopole for different top-hat radii ($\lambda=100$ mm, $a=0.001588 \lambda$, $h=0.1 \lambda$, $t \cong a$ and $\Delta=0.1a$)

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Fig. 5.13. $|rE_\theta|$ with respect to θ for ($\lambda=300$ mm, $h=\lambda/30$, $a=0.9528$ mm, $t \cong a$ and $\Delta=0.1a$)

Fig. 5.14. Input admittance of the unloaded top-hat monopole antenna for ($h=10$ mm, $a=0.9528$ mm, $b=8.2a$, $t \cong a$ and $\Delta=0.1a$)

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Fig. 5.15. Input admittance of the loaded top-hat monopole with ($a=0.9528$ mm, $b=8.2a$, $h=10$ mm, $t \approx a$, $\Delta=0.1a$, $\epsilon_r=45.35$ and $\mu_r=1$)

5.3.3 Material Loaded Top-Hat Monopole

It is very well known that loading an antenna with dielectric materials of high permittivity and/or permeability would reduce the resonance frequency and hence would increase the radiation efficiency at lower frequencies. This fact is due to polarization of the loading material, which leads to radiate more energy into the far-zone and reduce the near field storage. As an example, it can be seen through Figs. 5.15 and 5.16 that loading a top-hat monopole of height $\lambda_0/30$ with a material of $\epsilon_r=45.35, \mu_r=1$ or $\epsilon_r=1, \mu_r=7.64$ would make the monopole be resonant at $\lambda_0=300$ mm.

By increasing the permittivity and/or permeability of the loading material, the resonance frequency of the antenna clearly would decrease more and the conductance curve would become sharper, which leads to a narrower bandwidth and bigger mismatch to the feed line.

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It has recently been suggested that negative index materials can be employed to reduce the resonance frequency of the short antenna and hence increase its radiation efficiency at lower frequencies [91, 92]. However, it seems that epsilon negative materials can do the job and there is no need to invoke negative permeability [91]. And this is specifically of more interest, since fabrication of ENG materials is easier and also they suffer from smaller amount of loss and dispersion. Therefore, as it can be seen through Fig. 5.17, reducing the absolute value of the negative permittivity of the loading ENG material, the resonance frequency of the antenna becomes smaller regardless how small the antenna is.

Fig. 5.16. Input admittance of the loaded top-hat monopole with ($a=0.9528$ mm, $b=8.2a$, $h=10$ mm, $t \approx a$, $\Delta=0.1a$, $\epsilon_r=1$ and $\mu_r=7.64$)

Satisfying the second thermodynamic principle, a composite medium such as MTM with small inclusions arranged in a low periodicity must be frequency dispersive [129]. On the other hand, practical MTMs are lossy media due to their conductive inclusions such as rings or wires. Considering both properties (dispersion and loss), different models for MTMs were proposed in

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literature [130]. Drude model which is a simplified form of Lorentz model can be employed to take the MTM dispersion and loss into account. This model is actually coming from the polarization of the bounded charges due to impinging electromagnetic plane waves. Considering the microscopic forces and oscillations in such case and writing the second Newton's law, this simple model would be obtained [125]. Since I am interested in the ENG materials, I just consider the frequency dispersion and loss in the relative permittivity of the medium;

$$\varepsilon_r(\omega) = 1 - \frac{\omega_p^2}{\omega(\omega - j\Gamma)} = \left(1 - \frac{\omega_p^2}{\omega^2 + \Gamma^2}\right) - j \frac{\Gamma \omega_p^2}{\omega(\omega^2 + \Gamma^2)} \quad (5.24)$$

where ω_p is the plasma frequency and Γ is the collision frequency. It is noticed that the real part of relative permittivity would be negative as long as $\omega < \sqrt{\omega_p^2 - \Gamma^2}$.

Fig. 5.17. Input susceptance of the loaded top-hat loaded monopole with respect to its relative permittivity loading for different top-hat radiuses ($a=0.001588\lambda, h=\lambda/20, t\cong a, \Delta=0.1a$ and $\mu_r=1$)

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As it was mentioned before, it can be found a specific value for the real part of the ENG material permittivity to make the antenna resonant at the desired frequency. Therefore, on the basis of this value and also the desired tangent loss, we can find the plasma and collision frequency. As an example, we found that a permittivity like $\epsilon_r = -1.426$ can reduce the resonance frequency of a top-hat monopole of $h=10$ mm, $a=0.9528$ mm and $b=8.2a$ to $f=1$ GHz. It means that if we load the mentioned top-hat monopole antenna with an ENG material with $\epsilon_r = -1.431$, the imaginary part of the antenna input admittance would cross the frequency axis and the real part would possess a positive slop at $f=1$ GHz. However, ENG materials are dispersive and we need to find the plasma frequency and collision frequency. As I am going to compare the ENG and DPS loading in the antenna miniaturization, I take a very low tangent loss like $\tan \delta = 0.0008$. It should be mentioned that in reality, due to conductive inclusions, the tangent loss value is much higher than the assumed one for the ENG material. So finding the real and imaginary part of the permittivity at the resonance frequency, we are able to find the collision and plasma frequency of the ENG material as

$$\begin{cases} \Gamma = \frac{\epsilon_r''}{1 - \epsilon_r'} \omega_0 \\ \omega_p = \omega_0 \left[(1 - \epsilon_r') + \frac{(\epsilon_r'')^2}{(1 - \epsilon_r')} \right]^{0.5} \end{cases} \quad (5.25)$$

Utilizing (5.25), we find $\omega_p = 9.796$ GRad/sec and $\Gamma = 3$ MRad/sec. Therefore by loading the top-hat monopole of the same size as mentioned above ($h=\lambda/30$ and $\lambda=300$ mm) with an ENG material of plasma frequency 9.796 Grad/sec and collision frequency 3 MRad/sec, we can have resonance at 1 GHz. It is noted that the relative permittivity of the material at $\lambda=300$ mm is $-1.431 - j0.0012$ with tangent loss of 0.0008.

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Input admittance and directivity pattern of the ENG loaded top-hat monopole is shown in Figs. 5.18 and 5.19, respectively. It is seen that the radiation pattern of the antenna is similar to that of the conventional quarter wavelength monopole antenna. It can be shown very easily that increasing the collision frequency, the conductance curve would be dampened and the reflection coefficient would decrease and hence the bandwidth of the antenna would increase. However, increasing the collision frequency leads to greater conductive loss and smaller radiation efficiency. It might be expected that the conductance curve of the DPS loaded antenna would become wider if we consider loss in the DPS loading material. In order to have a fair comparison between the ENG and DPS loadings, I take lossy high permeability material into account. It should be mentioned that I am more interested in high permeability materials than the high permittivity materials here, since high permittivity materials lead to sharper resonances in comparison with high permeability materials which is due to dominance of the *TM* modes in the shown structure in Fig. 5.1. Then, loading the same monopole with a high permeability material of the same tangent loss as the ENG material has at the resonance frequency (1 GHz), the input admittance of the antenna is computed and shown in Fig. 5.20. It is seen that the conductance curve becomes wider as the tangent loss increases. However, due to small tangent loss, the input admittance curves (susceptance and conductance) of the lossy case would not be very different from the lossless one.

It is known that radiation resistance and radiation quality factor of an ESA are its two major factors to determine the quality of its radiation at low frequencies. As it was mentioned in the literature review, there are various ways to determine the radiation quality factor Q of an ESA. In this chapter, I use the formula for Q given in [14], which is a general expression using the derivative of the antenna input impedance at its resonance frequency;

$$Q \approx \frac{\omega_0}{2R_0} |Z'_m(\omega_0)| \quad (5.26)$$

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where ω_0 is the resonance angular frequency and R_0 is input impedance of the antenna at the resonance frequency. Utilizing the formula given above, it is found out that the radiation quality factor of the ENG loaded top-hat monopole is about 7350.474 while lossy DPS ($\mu_r = 7.64 - j0.0064$) loading would exhibit a Q around 214.266! It means the DPS loading would offer a much wider bandwidth than the ENG one provided the loaded antenna is matched to its feed-line.

Fig. 5.18. Input admittance of the ENG loaded top-hat monopole with ($\omega_p = 9.796$ GRad/sec, $\Gamma = 3$ MRad/sec, $a = 0.9528$ mm, $b = 8.2a$, $h = 10$ mm, $t \approx a$, $\Delta = 0.1a$ and $\mu_r = 1$)

Fig. 5.19. Radiation pattern of the ENG loaded top-hat monopole with ($\omega_p = 9.796$ GRad/sec, $\Gamma = 3$ MRad/sec, $a = 0.9528$ mm, $b = 8.2a$, $h = 10$ mm, $t \approx a$, $\Delta = 0.1a$ and $\mu_r = 1$)

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Radiation resistance of the antenna can simply be obtained from its radiated power and input current as

$$R_r = \frac{2P_r}{|I_{in}|^2} \quad (5.27)$$

where I_{in} is the input current, $P_r = \int_0^{2\pi} \int_0^{\pi/2} \frac{E_\theta \times H_\phi^*}{2} r^2 \sin \theta d\theta d\phi$, E_θ and H_ϕ were given in (5.25) and (5.26) and the superscript “*” stands for the complex conjugate of the function. The radiation resistances of the ENG and DPS loaded top-hat monopole antenna is shown in Fig. 5.21.

It is noticed that the radiation resistance of the DPS loaded monopole is higher than that of the ENG loaded antenna around 1 GHz. It is also seen that the radiation resistance of the DPS case is flat, while it is fluctuating a lot for the ENG loaded antenna. This is due to the loss tangent of the DPS loading material and also the frequency dispersion of the ENG medium. Consequently, the ratio of the radiated power to the input power of the DPS loaded monopole ($V_{in} = 1$ V) at the resonance frequency would be more than that of the ENG one as it is indicated in Fig. 5.22. This is a fact that can be seen from Fig. 5.24, where the current distributions of the DPS and ENG loaded top-hat monopole antenna are shown. It is seen that the current distribution of the DPS loaded antenna is more uniform than the ENG one.

Invoking the radiation efficiency of the antenna as

$$\eta_{rad} = \frac{P_r}{P_r + P_d} \quad (5.28)$$

where $P_d = \frac{\sigma}{2} \int_0^{2\pi} \int_0^h \int_a^b \left\{ |E_\rho^{III}(\rho, z)|^2 + |E_z^{III}(\rho, z)|^2 \right\} \rho d\rho dz d\phi$ is the dissipated power and P_r is the radiated , I can take all the parameters into account and

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have a more comprehensive comparison between the ENG and DPS loading. Shown in Fig. 5.25 is the radiation efficiency of the ENG and DPS loaded top-hat monopole antenna, respectively. It is observed through Figs. 5.22-23 that the radiation efficiency of the DPS loaded antenna at 1 GHz is much larger and flatter than that of the ENG loaded monopole which was obviously expected.

During my study in the previous subsection and present subsection, we observed that increasing radius of top-hat in the shown monopole and/or loading it with ENG materials; we can reduce the resonance frequencies and hence be able to miniaturize the antenna. As we mentioned before, improving the current distribution along the vertical stem of the top-hat monopole and increasing the capacitance of the antenna, increasing the radius of the top-hat can cause reduction in resonance frequencies and miniaturization. This is a well-understood fact. Because of the limitations on antenna miniaturization, one can not increase the top-hat radius as much as he desires. Resorting to the usage of ENG materials becomes inevitable if a highly miniaturized monopole is desired though the bandwidth of operation could not be that broad.

Fig. 5.20. Input admittance of the DPS loaded top-hat monopole with ($a=0.9528$ mm, $b=8.2a$, $h=10$ mm, $t \approx a$, $\Delta=0.1a$, $\epsilon_r=1$ and $\mu_r=7.64-j0.0064$)

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Fig. 5.21. Radiation resistance of the DPS ($\epsilon_r=1$ and $\mu_r=7.64-j0.0064$) and ENG ($\omega_p=9.796$ GRad/sec, $\Gamma=3$ MRad/sec) loaded top-hat monopole for ($a=0.9528$ mm, $b=8.2a$, $h=10$ mm, $t \approx a$ and $\Delta=0.1a$)

Fig. 5.22. $|P_r|/|P_{in}|$ of the DPS ($\epsilon_r=1$ and $\mu_r=7.64-j0.0064$) and ENG ($\omega_p=9.796$ GRad/sec, $\Gamma=3$ MRad/sec) loaded top-hat monopole for ($a=0.9528$ mm, $b=8.2a$, $h=10$ mm, $t \approx a$ and $\Delta=0.1a$)

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Fig. 5.23. $|rE_\theta|$ of the DPS ($\epsilon_r=1$ and $\mu_r=7.64-j0.0064$) and ENG ($\omega_p=9.796$ GRad/sec, $\Gamma=3$ MRad/sec) loaded top-hat monopole for ($a=0.9528$ mm, $b=8.2a$, $h=10$ mm, $t \approx a$ and $\Delta=0.1a$)

Fig. 5.24. Normalized current distribution of the DPS ($\epsilon_r=1$ and $\mu_r=7.64-j0.0064$) and ENG ($\omega_p=9.796$ GRad/sec, $\Gamma=3$ MRad/sec) loaded top-hat monopole for ($a=0.9528$ mm, $b=8.2a$, $h=10$ mm, $t \approx a$ and $\Delta=0.1a$)

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Fig. 5.25. Radiation efficiency of the DPS ($\epsilon_r=1$ and $\mu_r=7.64-j0.0064$) and ENG ($\omega_p=9.796$ GRad/sec, $\Gamma=3$ MRad/sec) loaded top-hat monopole for ($a=0.9528$ mm, $b=8.2a$, $h=10$ mm, $t \approx a$ and $\Delta=0.1a$)

It is known that one needs to couple the guided modes of the transmission feed lines to the radiating TM_{10} spherical modes in free space through the radiation of an antenna. The antenna actually works as a resonator. And also it is known that an electrically small antenna is not a good resonator at lower frequencies and hence can not provide an efficient coupling between the aforementioned modes. But it was notified in this chapter that using DPS (high positive permeability and/or permittivity) or ENG materials this difficulty disappears. Assume the electrically small top-hat monopole antenna as a very small cylindrical particle (resonator). The resonance of this small particle due to the incidence of electromagnetic plane wave would take place at very high frequencies. If we fill (load) the cylindrical resonator with DPS or ENG materials, as it was seen, we can reduce the resonance frequencies to lower frequencies regardless of how small the particle is. So if we juxtapose the electrically small antenna with a DPS or an ENG resonator “properly”, we can

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acquire an efficient coupling between the guided modes and radiated modes. In fact, the electrically small antenna plays a role the same as a probe to excite the resonator. The question is how we can juxtapose the antenna and the ENG resonator “properly”?

It is well known that an electrically small monopole is highly capacitive. In fact, the ESA can be modeled as a series combination of a capacitor, an inductor and a resistor which exhibits a very high capacitive reactance and a very poor resistance at low frequencies. In order to make this combination resonant and increase the delivered power to the modeled resistor (radiation resistance), we basically have two choices: increasing the capacitor or the inductor. By ending the antenna with the top-hat and loading it with DPS materials, we actually increase the capacitor value in the antenna circuit model and hence reduce the capacitive reactance at low frequencies. And also it has been shown that an ENG structure consists of small particles which can exhibit an inductive characteristic collectively [131]. So proper juxtaposition means we need to invoke an ENG material which has an opposite reactive property to that of the electrically small monopole. In other words, a monopole of a unique size can excite a resonator of a unique permittivity. Proper selection of parameters h , b and ϵ_r or μ_r in my formulated structure can lead us to miniaturize the electrically small monopole in a great deal.

Throughout this section, a comprehensive study made to compare ENG and DPS materials in miniaturization of a top-hat monopole antenna. It was shown that both the materials are able to reduce the resonance frequency of the antenna unlimitedly. However, as it was noted, ENG materials are suffering from higher amounts of dispersion and conductive loss rather than DPS (high permeability) materials. Accordingly, the radiation efficiency of the DPS loaded antenna would be higher and flatter than the ENG loaded one. The offered bandwidth of the DPS loaded antenna is larger than the ENG one due to higher radiation quality factor of the DPS loaded one. It was also found out that the current distribution of the DPS loaded monopole is more uniform than that of

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the ENG loaded top-hat monopole and hence the ratio P_r / P_{in} in the DPS case would be bigger than the ENG one. On the other hand the fabrication and engineering of ENG materials is more problematic than DPS materials. As a consequence of all these respects, DPS (high permeability) loaded top-hat monopole antenna would be preferred to the ENG loaded one, though both the materials would lead to the effective monopole antenna miniaturization.

As an example, finally, I show the radiation parameters of a DPS and an ENG loaded top-hat monopole antenna of height $\lambda_0 / 64$ while $\lambda_0 \approx 630$ mm through the following figures (Figs. 5.26-35). It is seen that utilizing both the structure bending and material loading can lead us to proficiently miniaturize the antenna to such an electrically small size. It is also seen that the antenna is resonant and the radiation pattern of the material loaded top-hat monopole antenna is similar to that of the quarter wave length monopole antenna over a large ground plane at $\lambda_0 \approx 630$ mm.

Fig. 5.26. Input admittance of the ENG loaded top-hat monopole with ($\omega_p = 4.58$ GRad/sec, $\Gamma = 1.5$ MRad/sec, $a = 0.9528$ mm, $b = 8.2a$, $h = 9.77$ mm, $t \approx a$, $\Delta = 0.1a$ and $\mu_r = 1$)

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Fig. 5.27. Input impedance of the ENG loaded top-hat monopole with ($\omega_p = 4.58$ GRad/sec,

$\Gamma = 1.5$ MRad/sec, $a = 0.9528$ mm, $b = 8.2a$, $h = 9.77$ mm, $t \approx a$, $\Delta = 0.1a$ and $\mu_r = 1$)

Fig. 5.28. Input admittance of the DPS loaded top-hat monopole with ($a = 0.9528$ mm, $b = 8.2a$,

$h = 9.77$ mm, $t \approx a$, $\Delta = 0.1a$ and $\mu_r = 34.42 - j0.03$ and $\epsilon_r = 1$)

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Fig. 5.29. Input impedance of the DPS loaded top-hat monopole with ($a=0.9528$ mm, $b=8.2a$, $h=9.77$ mm, $t \approx a$, $\Delta=0.1a$ and $\mu_r=34.42-j0.03$ and $\epsilon_r=1$)

Fig. 5.30. Radiation pattern of the ENG loaded top-hat monopole with ($\omega_p=4.58$ GRad/sec, $\Gamma=1.5$ MRad/sec, $a=0.9528$ mm, $b=8.2a$, $h=9.77$ mm, $t \approx a$, $\Delta=0.1a$ and $\mu_r=1$)

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Fig. 5.31. Radiation pattern of the DPS loaded top-hat monopole with ($a=0.9528$ mm, $b=8.2a$, $h=9.77$ mm, $t \approx a$, $\Delta=0.1a$ and $\mu_r=34.42-j0.03$ and $\epsilon_r=1$)

Fig. 5.32. Normalized current distribution of the ENG ($\omega_p=4.58$ GRad/sec, $\Gamma=1.5$ MRad/sec) and a DPS ($\mu_r=34.42-j0.03$ and $\epsilon_r=1$) loaded top-hat monopole with ($a=0.9528$ mm, $b=8.2a$, $h=9.77$ mm, $t \approx a$ and $\Delta=0.1a$)

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Fig. 5.33. Ratio of the radiated power of an ENG ($\omega_p=4.58$ GRad/sec, $\Gamma=1.5$ MRad/sec) and a DPS ($\mu_r=34.42-j0.03$ and $\epsilon_r=1$) loaded top-hat monopole to an unloaded monopole without top-hat for ($V_{in}=1$, $a=0.9528$ mm, $b=8.2a$, $h=9.77$ mm, $t \approx a$ and $\Delta=0.1a$)

Fig. 5.34. $|rE_\theta|$ of an ENG ($\omega_p=4.58$ GRad/sec, $\Gamma=1.5$ MRad/sec) and a DPS ($\mu_r=34.42-j0.03$ and $\epsilon_r=1$) loaded top-hat monopole for ($V_{in}=1$, $a=0.9528$ mm, $b=8.2a$, $h=9.77$ mm, $t \approx a$, $\Delta=0.1a$ and $f=476$ MHz)

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Fig. 5.35. $|P_r|/|P_{in}|$ of an ENG ($\omega_p=4.58$ GRad/sec, $\Gamma=1.5$ MRad/sec) and a DPS ($\mu_r=34.42-j0.03$ and $\epsilon_r=1$) loaded top-hat monopole for ($V_{in}=1$, $a=0.9528$ mm, $b=8.2a$, $h=9.77$ mm, $t \approx a$ and $\Delta=0.1a$)

Fig. 5.36. Radiation resistance of an ENG ($\omega_p=4.58$ GRad/sec, $\Gamma=1.5$ MRad/sec) and a DPS ($\mu_r=34.42-j0.03$ and $\epsilon_r=1$) loaded top-hat monopole for ($a=0.9528$ mm, $b=8.2a$, $h=9.77$ mm, $t \approx a$ and $\Delta=0.1a$)

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Fig. 5.37. Radiation efficiency of an ENG ($\omega_p=4.58$ GRad/sec, $\Gamma=1.5$ MRad/sec) and a DPS ($\mu_r=34.42-j0.03$ and $\epsilon_r=1$) loaded top-hat monopole for ($a=0.9528$ mm, $b=8.2a$, $h=9.77$ mm, $t \approx a$ and $\Delta=0.1a$)

5.4 Conclusions

Utilizing an eigenmode approach, a material loaded top-hat monopole was studied and its radiation parameters were derived. Comparing with measured results, the validity of the approach was approved. It was seen that adding a top-hat to a short monopole would make the highly triangular current distribution more uniform and hence reduce the resonance frequency. However, widening the top-hat unlimitedly is not desirable from the miniaturization point of view. In order to reduce the resonance frequency further, we need to resort to loading the antenna with materials. As it is known, due to their anomalous properties, nowadays MTMs are attracting lots of attentions in different areas of science such as microwave and antenna design. However it would be a very valuable study if a fair comparison between the DPS and MTMs are made in such areas of science and engineering.

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Taking DPS and ENG materials into account, it was shown that both the materials can reduce the resonance frequency of the antenna effectively and enhance its radiation into the far-zone at low frequencies. Deriving all the antenna parameters, a comprehensive comparison between the DPS and ENG loaded top-hat monopole antenna was made. It was shown that due to its higher radiation efficiency and radiation resistance, broader bandwidth and smaller radiation quality factor, easier fabrication and also less dependence upon frequency, DPS loaded top-hat monopole antenna would be preferred to the ENG loaded one.

CHAPTER 6

Conclusions and Future Work

6.1 Conclusions

The main scope of this thesis was to study the effects of loading homogenous, linear and isotropic dielectric materials with positive and/or negative constitutive parameters on the performance of the electrically small monopole antenna over an infinitely large ground plane in an analytic way. Noticeably, a considerable amount of attention has been paid to the anomalous MTMs (ENG, MNG and DNG) and their applications in various areas of physics and engineering recently. The electromagnetic properties of MTMs are really amazing and have led to a tremendous number of publications in recent years. One of the major areas, which have attracted lots of attention, is the antenna miniaturization with the aid of novel MTMs.

As I mentioned during my literature review, antenna miniaturization techniques are divided into two main groups: employing novel structures, loading with materials. Nowadays, so much effort is being dedicated to loading the ESA with different types of materials such as magneto-dielectrics, EBGs and MTMs to enhance its radiation performance at low frequencies. Researchers are to make use of material properties and their interactions with near fields to develop the radiation process of an ESA at low frequencies and create a versatile miniaturized antenna, which is a difficult task. Antenna is a sort of transition between the guided waves and spherical radiated waves and matches these two types effectively. The characteristics of this resonator such as resonance frequency and quality factor are directly proportional to its geometry and physical dimensions. Antenna beside the dielectric materials would construct a dielectric loaded resonator whose parameters could be totally different from the unloaded (single) antenna. As a matter of fact, a proper

loading with proper material of proper geometry could cause to betterment of the ESA operation at low frequencies.

During this thesis, I aimed to investigate the properties of a vertical monopole which is loaded with dielectric materials. The monopole was fed by a small gap source over an infinitely large ground plane. First, I validated the gap source model against the practical coaxial source and found the optimum value for the gap height, which is, to the best of my knowledge, reported for the first time in this dissertation. Taking advantage of the optimum value for the gap height, I studied the effects of dielectric loading on performance of the mentioned monopole in an analytical way. I also combined two main antenna miniaturization methods (shaping and loading) to obtain a highly miniaturized antenna with a reliable radiation efficiency at low frequencies. Referring to my formulation and numerical results, I managed to compare the MTMs with usual dielectric materials (of high permittivity and/or high permeability) in antenna miniaturization and distinguish between their corresponding pitfalls and advantages. However, it should be mentioned that this comparison between MTMs and DPS materials was made for the vertical monopole antenna and may lead to different conclusions for other types of antennas.

6.2 Recommendations for Future Work

As I mentioned before, MTMs have become one of the hottest topics in electromagnetic wave theory and led to a significant number of publications. However, MTMs are not available in nature and their synthesis in different applications might be costly. Consequently, conducting a comprehensive study on the reality of MTMs, their shortcomings and the benefits they may offer in different applications without any exaggeration is obviously needed. Specifically, a general purpose antenna loaded with a general type of material is suggested to be taken into account. The volume of the loading material could be a portion or the entire radian sphere around the antenna. Utilizing the spherical waves, which satisfy Maxwell's equations inside and outside the radian sphere

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and following a similar approach to Chu's approach, the fundamental limitations on the performance of a material loaded ESA would be redefined and fair judgment on different types of materials would be obtained.

Artificial Perfect Magnetic Conductor (PMC) could be achieved by the virtue of EBG materials. Therefore, it would be an interesting direction for the future work to invoke EBG materials in fabricating the reconfigurable antennas that could offer higher gain, multiple-band and/or wideband and circular polarization in comparison with the conventional antennas.

Another kind of newly proposed materials is the zero-index material, which could offer zero-phased electromagnetic wave propagation [132]. This zero-phased wave propagation could lead to squeeze of the electromagnetic beam in a particular direction, provided the zero-index medium has a boundary of appropriate geometry with the adjacent medium (air). In another word, when the electromagnetic wave is traveling from the zero-index medium to a DPS medium ($n > 1$) such as air, all the electromagnetic beams will be bent toward the line normal to the cross section between two media. This fact is due to Snell's law and shown in Fig. 6.1.

Fig. 6.1. Electromagnetic ray travelling from a zero-index medium into air

$$n_1 \sin(\theta_1) = n_2 \sin(\theta_2) \quad (6.1)$$

$$\lim_{n_1 \rightarrow 0} \theta_2 = 0 \quad (6.2)$$

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It is interesting to note that if the electromagnetic wave travels from air to the zero-index medium, all the rays will be reflected back, regardless of the incident angle. Both the mentioned phenomena (squeeze and total reflection) could be employed in a prospective research work to create highly directive antennas and also surface wave antennas and waveguides.

As it is known, all the structures that have been proposed as MTMs, all are anisotropic media, in which the anisotropy could be in one direction or more. Hence, it would be worthwhile studying the effects of anisotropic MTMs on the radiation of an ESA. Some elements of constitutive tensors of the anisotropic MTMs could be negative, while the others would be positive or zero. Accordingly, an analytical method such as the eigenmode analysis could be utilized to explore the properties of the ESA, say a vertical monopole, loaded with an anisotropic material. Controlling the element parameters of the constitutive tensors in anisotropic materials could bring us an extra flexibility to synthesize our desired ESA's radiation properties, such as radiation pattern and radiation beam-width.

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