Linköping Studies in Science and Technology
Dissertations No. 1146

Ultra-wideband Antenna and Radio Front-end Systems

Magnus Karlsson

Linköping University
INSTITUTE OF TECHNOLOGY

Department of Science and Technology
Linköping University, SE-601 74 Norrköping, Sweden

Norrköping 2007

Ultra-wideband Antenna and Radio Front-end Systems

A dissertation submitted to the Institute of Technology, Linköping University, Sweden for the degree of Doctor of Technology.

ISBN: 978-91-85895-36-6
ISSN: 0345-7524
http://urn.kb.se/resolve?urn=urn:nbn:se:liu:diva-10338

Linköping University
Department of Science and Technology
SE-601 74 Norrköping
Sweden
Printed by LiU-Tryck, Linköping 2007

Abstract

The number of wireless communication applications increase steadily, leading to the competition for currently allocated frequency bands. Pressure on authorities around the world to permit communications in higher and wider frequency ranges to achieve higher wireless capacity than those existed in the past has resulted in several new specifications. The federal communication commission (FCC) in USA has unleashed the band 3.1-10.6 GHz for ultra-wideband radio (UWB) communications. The release has triggered a worldwide interest for UWB. Other regulatory authorities throughout the world have issued use of UWB techniques as well. Capacity issues in form of data rate and latency have always been a bottleneck for broadened wireless-communication usages. New communication systems like UWB require larger bandwidth than what is normally utilized with traditional antenna techniques. The interest for compact consumer electronics is growing in the meantime, creating a demand on efficient and low profile antennas which can be integrated on a printed circuit board. In this thesis, some methods to extend the bandwidth and other antenna parameters associated with wideband usages are studied. Furthermore, methods on how to enhance the performance when one antenna-element is not enough are studied as well.

The principle of antenna parallelism is demonstrated using both microstrip patch antennas and inverted-F antennas. Several techniques to combine the antennas in parallel have been evaluated. Firstly, a solution using power-splitters to form sub-arrays that covers one 500-MHz multi-band orthogonal frequency division multiplexing (OFDM) UWB is shown in *Paper I*. It is then proposed that the sub-bands are selected with a switch network. A more convenient method is to use the later developed frequency multiplexing technique as described in *Papers V* and *VIII*. Using the frequency multiplexing technique, selective connection of a number of antennas to a common junction is possible. The characteristic impedance is chosen freely, typically using a 50-Ω feed-line. Secondly, in *Paper VIII* a frequency-triplexed inverted-F antenna system is investigated to cover the Mode 1 multi-band UWB bandwidth 3.1-4.8 GHz. The antenna system is composed of three inverted-F antennas and a frequency triplexer including three 5th order bandpass filters. In *Papers VI* and *X* a triplexer without and with low noise amplifier (LNA) integrated in a printed circuit board for multi-band UWB radio is presented. The triplexer utilizes a microstrip network and three combined broadside- and edge-coupled filters. The triplexer is fully integrated in a four metal-layer printed circuit board with the minimum requirement on process tolerances. Furthermore, the system is built completely with distributed microstrips, i.e., no discrete component except the LNA. Using the proposed solution an equal performance in the sub-bands is obtained. Finally suitable monopoles and dipoles are discussed and evaluated for UWB. In *Paper XI* circular monopole and dipole antennas for UWB utilizing the flex-rigid concept are proposed. The flex-rigid concept combines flexible polyimide material with the regular printed circuit board material. The antennas are placed entirely on the flexible part while the antenna ground plane and the dipole antenna balun are placed in the rigid part.

Populärvetenskaplig sammanfattning

Antalet trådlösa radioapplikationer bara ökar, vilket medför att konkurrensen om tillgängliga frekvensband hårdnar. År 2002 öppnade federal communication commission (FCC) i USA frekvensbandet 3.1-10.6 GHz för ultra bredbandskommunikation. Det här frisläppandet av bandbredd utlöste en våg av förnyat intresse för bredbandiga radiosystem bland forskare världen över. Konstruktion av ultra bredbandsradio (UWB) ställer nya hårdare krav på kompetens och designmetodik. Kapacitetsbegränsningar i form av fördröjning och datakapacitet har alltid varit ett problem för trådlösa överföringstekniker. Bandbreddskravet för ett typiskt UWB-system är mycket högre än vad som traditionellt har krävts av en antenn i ett kommunikationssystem. Detta samtidigt som man eftersträvar att bygga små, kompakta system som kan integreras i alla möjliga sammanhang. I den här avhandlingen har olika metoder för att kontrollera viktiga parametrar i bredbandskommunikation, som till exempel att öka antennens impedansbandbredd. Förutom det har olika lösningar för att kombinera flera antenner studerats, det vill säga att effektivt koppla samman flera radiatorer parallellt när en antenn inte räcker till.

Principen för antennparallellism demonstreras både med microstrip patch-antenner och med inverterade-F antenner. En teknisk lösning baserad på switchar och delare har utvärderats. Ytterligare en teknisk lösning som studerats ingående för antennparallellism är frekvensmultiplexning. Med frekvensmultiplexning kan varje antenn täcka en del av det totala frekvensbandet, vilket kan vara ett 500 MHz frekvensband i UWB sammanhang. Förutom multiantennlösningar har även bredbandiga antennelement och möjliga integrationslösningar studerats. Speciellt intressant är antennintegration på så kallat ”flex-rigid” substrat som är en kombination av traditionellt (icke böjbart) och mjukt böjbart mönsterkortsmaterial. Det flexibla substratet är större och sticker ut från det hårda, således får man en fast (eng. rigid) och en flexibel (eng. flex) del. Den flexibla delen lämpar sig för att placera själva antennen som sen kan böjas åt olika håll. I rigid delen kan man placera övriga transceiverkomponenter som till exempel en balun.

Acknowledgements

First of all I would like to express my gratitude to my supervisor, Professor Shaofang Gong, for providing guidance through the years, and not least the opportunity of doing research in this interesting and challenging field. Furthermore, I want to thank all the personell at the Department of Science and Technology who in various ways have supported me in my work.

Many thanks are to the teachers in various postgraduate and graduate courses that have given me the foundation to do this work. Furthermore I want to thank remaining teachers in my previous educations that have improved my theoretical knowledge and technological skills.

Moreover, I would like to thank Acreo AB for support with measurements, especially Patrick Blomqvist and Magnus Svensson. Furthermore, Ericsson AB in Sweden is acknowledged for financial support.

Last but not least, I would like to express my deepest gratitude to my parents Bo and Gunilla Karlsson. Without their support, it would be impossible for me to achieve what I have done.

List of Publications

Papers included in the thesis:

[I] **M. Karlsson**, and S. Gong, "Wideband patch antenna array for multi-band UWB," *Proc. IEEE 11th Symp. on Communications and Vehicular Tech.*, Ghent, Belgium, Nov. 2004.

[II] **M. Karlsson**, and S. Gong, "An integrated spiral antenna system for UWB," *Proc. IEEE 35th European Microwave Conf.*, Paris, France, Oct. 2005, pp 2007-2010.

[III] **M. Karlsson**, and S. Gong, "Monofilar spiral antennas for UWB with and without air core," *ISAST Transactions on Electronics and Signal Processing*, 2007.

[IV] **M. Karlsson**, and S. Gong, "Air core patch antennas suitable for multi-band UWB," *Proc. GigaHertz 2005*, Uppsala, Sweden, Nov. 2005.

[V] **M. Karlsson**, P. Håkansson, A. Huynh, and S. Gong, "Frequency-multiplexed Inverted-F Antennas for Multi-band UWB," *IEEE Wireless and Microwave Technology Conf. WAMICON 2006*, pp. FF-2-1 - FF-2-3, Dec. 2006.

[VI] A. Serban, **M. Karlsson**, and S. Gong, "All-Microstrip Design of Three Multiplexed Antennas and LNA for UWB Systems," *Proc. Asia-Pacific Microwave Conf.*, 2006, pp. 1109-1112, Dec. 2006.

[VII] **M. Karlsson**, P. Håkansson, S. Gong, "A Frequency Triplexer for Ultra-wideband Systems Utilizing Combined Broadside- and Edge-coupled Filters," *Manuscript submitted to IEEE Transactions on Advanced Packaging*, 2007.

[VIII] **M. Karlsson**, and S. Gong, "A Frequency-Triplexed Inverted-F Antenna System for Ultra-wide Multi-band Systems 3.1-4.8 GHz," *ISAST Transactions on Electronics and Signal Processing*, 2007.

[IX] A. Serban, **M. Karlsson**, and S. Gong, "Microstrip Bias Networks for Ultra-Wideband Systems," *ISAST Transactions on Electronics and Signal Processing*, 2007.

[X] A. Serban, **M. Karlsson**, and S. Gong, "A Frequency-triplexed RF Front-end for Ultra-wideband Systems 3.1-4.8 GHz," *Manuscript, submitted to ISAST Transactions on Electronics and Signal Processing*, 2007.

[XI] **M. Karlsson**, S. Gong, "Mono- and Di-pole Antennas for UWB Utilizing Flex-rigid Technology," *ISAST Transactions on Electronics and Signal Processing*, 2007.

Contents

Preface:

Chapters:

Appended papers:

Due to copyright restrictions the articles have been remov

List of Abbreviations

2.5D	2.5 dimensional
3D	3 dimensional
ADS	Advanced design system
BOK	bi-orthogonal keying
BPSK	binary phase shift keying
CAD	computer-aided design
CAT	computed axial tomography
CEPT	Conferance of Postal and Telecommunications Administrations
CF	center frequency
CW	continuous wave
DoD	U.S. Department of Defense
DS	direct sequence
DSSS	direct sequence spread spectrum
EC	European commission
ECC	(European) Electronic Communications Committee
EIRP	equivalent isotropically-radiated power
EM	electromagnetic
ERO	European Radio communications Office
ETSI	European Telecommunication Standard Institute
FCC	federal communication commission
IP	internet protocol
LNA	low noise amplifier
OFDM	orthogonal frequency division multiplexing
PCB	printed circuit board
PSD	power spectral density
R&O	report and order

RF	radio frequency
SNR	signal to noise ratio
SWR	standing wave ratio
TFC	time-frequency code
UWB	ultra wideband radio
VSWR	voltage standing wave ratio
WPAN	wireless personal area network

1. Introduction

Wireless data transfer capacity is a returning issue in modern wireless communications. To make a rough simplification there are two variables to work with: information density per frequency-unit and the occupied frequency bandwidth. To use a wide bandwidth enables more data transfer capacity but imposes great challenges to the transceiver front-end. The analog front-end including the antenna has been an important part ever since the first wireless communication system was invented. It has been subjects for a lot of research throughout the years, with various parameters in focus in order to further enhance system performance. When the idea of ultra wideband radio (UWB) entered the scene, the interest for radio frequency front-ends with very large bandwidth has got a rebirth.

1.1. Background and Motivation

Most of today's radio systems operate above 1 GHz, which are traditionally known as pure microwave frequencies. To build radio systems with both high frequencies and wide bandwidths demands new knowledge and approaches. First of all, high frequency designs cannot be treated the same way as for low frequency electronics, i.e., all signals must be treated as waves according to the transmission line theory. Traditional electronics design rules and techniques must be complemented with new rules and tools. Secondly a wide band operation has its own challenges with many complex questions. Classical radio systems utilize bandwidths that are within a couple of percents of the center frequency (CF). A narrow bandwidth provides large freedom for other antenna parameters. However, new technologies like UWB may utilize a relative bandwidth even larger than the CF. To build antennas with acceptable performance variation within such a wide bandwidth requires a thorough control of several antenna variables. Match techniques for narrow band circuits used in classical radio frequency (RF) systems is well known among RF designers, but designing wideband circuits and systems needs a new approach. When discussing bandwidth at high frequencies the definition fractional bandwidth or relative bandwidth is commonly used, which is the ratio of bandwidth over the CF.

1.2. Objectives

Emerging new wideband standards like UWB imposes new demands for wideband front-end systems. First of all, understanding of the specifications is needed to determine system requirements. Secondly, to design a front-end with a high fractional bandwidth requires a good knowledge about underlying theories and techniques. Finally, based on those studies different approaches for wideband antenna and front-end systems for the UWB bandwidth 3.1-10.6 GHz have been studied in this work. The goal is not only to find an antenna that has enough fractional bandwidth, but also to take the specifications of respective UWB services into account so that more efficient antenna solutions can be developed.

1.3. Outline of the Thesis

The thesis is of summary style, which presents the background of the work, the underlying technologies to be studied, and the finished scientific papers that suit the scope of the thesis. In Chapter 2 antenna theory and wideband antenna technologies are discussed. Chapter 4 presents UWB and the driving force for broad-band antennas. Chapter 5 gives detailed information about the author's contribution in the scientific papers.

In chapter 2 various theories and approaches to extend the bandwidth of wideband antennas are discussed. Their suitability for various UWB services is commented, supported by listed references.

In chapter 3 the types of antennas used in this work are presented. The microstrip patch, monofilar spiral, inverted-F, monopole, and dipole antennas are described.

In chapter 4 the UWB technology, potential, and various usage examples are summarized. The focus is on the applications, implementations, and areas most relevant to the scope of this thesis.

2. Ultra-wideband Radio Antennas

For over a century mankind has explored the electromagnetic phenomena that transmit signals through the air, well known as radio waves. The link between the electric circuitry and the air has always been a mysterious thing to many people. In fact, as all other technologies to build antennas starts from fundamental theories combined with the modern science. The bandwidth utilization of communication systems has increased throughout the years. Some of the very first implementations were very bandwidth consuming. However, it is only in modern times when antennas possess intentional wideband capabilities. In most consumer applications, space is an issue therefore planar or preferably integrated antennas on a printed circuit board are of interest for investigation.

2.1. Antenna History

Since there were lack of knowledge how to generate continuous waves, the radio pioneers had to rely on resonant circuits to feed the antenna. Such resonant circuitries generate damped sinusoidal impulses. In the time domain the transmission is seen as a train of impulses. Naturally when transferred in the frequency domain such a kind of pulse transmission consumes a very wide frequency band around the impulse center frequency. However, the fundamentals of electromagnetic wave propagation through the medium were known by the scientists a century ago. The relation among frequency, phase velocity, and wavelength had been established long time ago. This knowledge soon led to the idea of wavelength or frequency selected communication channels [1]-[3], the route towards the frequency-selective narrowband systems seen in many wireless applications today. A variety of resonant antenna technologies like the patch antenna have been presented since then. Parameters such as directivity, gain, and polarization often received more attention than the bandwidth extension, since with a fractional bandwidth of only a few percents the impedance-bandwidth performance of antennas has not been a issue before [2]-[5]. However, with today's UWB requirement the antenna bandwidth can be a bottle neck of a wireless communication system.

2.2. Theory and Techniques

Maxwell's well known equations lay the foundation of all electromagnetism and therefore the foundation of antenna technologies. Originally Maxwell's equations consisted of a total of 20 equations describing the behavior of electric and magnetic fields. More commonly used is the differential form of Maxwell's equations that describes the fields at a particular point in space, also known as the four fundamental equations of electromagnetism, shown in Eqs. 2.1-2.4 [1], [2]. Were E is the electric field intensity in volts per meter and ρ_t is the total electric charge density in coulombs per cubic meter. As seen in Eq. 2.5 the total electric charge density is dependant of the free charge density ρ_f, and the bound charge density ρ_b. Furthermore, the bound charge density depends of the electric polarization in coulombs per square meter P, see Eq. 2.6. B is the magnetic induction in teslas, J_m is the current density resulting from the flow of charges in matter, in amperes per square meter, J_f is the current density of free charges, dP/dt is the polarization current density, $\nabla \times M$ is the equivalent current density in magnetized matter, and M is the magnetization in amperes per meter. ε_0 is the permittivity of free-space, and μ_0 is the permeability of free-space. The equations are partial differential equations involving space and time derivatives of the electric (E) and magnetic (B) field vectors, the total charge density (ρ_t), and the current density (J_m). However, the equations do not yield any values of E and B directly before integration with proper boundary conditions applied. Eq. 2.1 is also known as Gauss' law and describes how charges produce electric fields. Eq. 2.2 yields magnetic flux conservation, i.e., there are no magnetic charges. Eq. 2.3 shows Faraday's law, showing that a change in the magnetic environment will induce a voltage, i.e., typically in a coil of conducting wire. Eq. 2.4 is known as the Generalized Ampere's law or Maxwell-Ampere's law. It shows the relation between the magnetic field, current flow and electric field [1], [2].

$$\nabla \bullet E = \frac{\rho_t}{\varepsilon_0} \tag{2.1}$$

$$\nabla \bullet B = 0 \tag{2.2}$$

$$\nabla \times E + \frac{dB}{dt} = 0 \tag{2.3}$$

$$\nabla \times B - \varepsilon_0 \mu_0 \frac{dE}{dt} = \mu_0 J_m \tag{2.4}$$

Were:

$$\rho_t = \rho_f + \rho_b \tag{2.5}$$

$$\rho_b = -\nabla \bullet P \tag{2.6}$$

$$J_m = J_f + \frac{dP}{dt} + \nabla \times M \tag{2.7}$$

Based on the very principles of its operation an antenna has its typical characteristics or behavior. In order to break the performance boundaries of a certain technology, various techniques have been developed through the years. For instance, parasitic or resistive loading in various forms has been presented [2]. Antennas relying on smooth changing geometrical shapes has also been investigated [2], [6]. The geometry of the antenna is altered so that the physical fundamental in terms of resonance length is no longer defined by a single distinct geometric length. Furthermore, to extend the bandwidth by using more than one resonating radiating element is another method. It can either be a multi-element resonating antenna structure or an array of antennas combined.

2.2.1. Antenna Principles and Printed Circuit Board Integration

An electromagnetic wave originating from a point source propagating in free space will propagate uniformly in all directions. At a far distance the radiation from the antenna has plane wave properties, i.e., a plane wave propagating in the far-field region [1], [7]. The velocity when propagating in free-space (c) is given by Eq. 2.8, were ε_0 is the permittivity of free-space, and μ_0 is the permeability of free-space. Eq. 2.9 shows the ratio between magnetic and electrical properties η that defines the free-space wave impedance. Furthermore, for free-space the relation between frequency f_0 and wavelength λ_0 is then only dependant of c, see Eq. 2.10 [1], [7].

$$c = \frac{1}{\sqrt{\mu_0 \varepsilon_0}} \approx 3 * 10^8 \text{ m/s} \quad (2.8)$$

$$\eta = \sqrt{\mu_0 / \varepsilon_0} \approx 377\,\Omega \quad (2.9)$$

$$f_0 = \frac{c}{\lambda_0} \quad (2.10)$$

Meanwhile propagation in a medium is also dependant of the relative parameters, i.e., the phase velocity v in a non-conducting medium. The relative parameters for permittivity, and permeability are ε_r, and μ_r, respectively. In air the propagation is normally approximated to be equal to that in the free space, i.e., $\varepsilon_r = 1$ and $\mu_r = 1$. For non-conducting material the complete expression for permittivity is defined as $\varepsilon_r \varepsilon_0$, and the permeability are $\mu_r \mu_0$. The Maxwell's wave equations for non-conducting materials are therefore defined as in Eqs. 2.11 and 2.12.

$$\nabla^2 E = -\varepsilon_r \varepsilon_0 \mu_r \mu_0 \omega^2 E \quad (2.11)$$

$$\nabla^2 H = -\varepsilon_r \varepsilon_0 \mu_r \mu_0 \omega^2 H \quad (2.12)$$

The phase velocity v_p of non-conducting material is then defined as in Eq. 2.13 and phase velocity in relation to free space propagation in Eq. 2.14. It is seen that the phase velocity relative to c slows down with the square-root of ε_r and μ_r. Eq. 2.15 shows the index of refraction n, which is the ratio that the phase velocity slows down relative to free-space propagation.

$$v_p = \frac{1}{\sqrt{\varepsilon_r \varepsilon_0 \mu_r \mu_0}} \quad (2.13)$$

$$v_p = \frac{c}{\sqrt{\varepsilon_r \mu_r}} \quad (2.14)$$

$$n = \frac{c}{v} = \sqrt{\varepsilon_r \mu_r} \quad (2.15)$$

In a non-magnetic medium the index of refraction is equal to the square root of the relative permittivity, i.e., in a non-magnetic material $\mu_r = 1$, see Eq. 2.16. The FR4 and the Rogers printed circuit board materials are considered as non-magnetic materials. The phase velocity is then simplified to the expressions in Eqs. 2.17 and 2.18. The relation between frequency and wavelength in non-magnetic material is then dependant of the relative permittivity as seen in Eq. 2.19. It should be noted that the phase velocity is valid for propagation within the dielectric material. For printed circuit board antenna integration it must be noted that the radiating antenna element has air on one side and a dielectric on the other side, i.e., the propagation depends on the effective permittivity. The antennas electrical field in the reactive near-field region spreads out partially into the dielectric material and partially into the surrounding air. Together with the length, and width ratio this gives an effective dielectric constant that depends on the actual implementation.

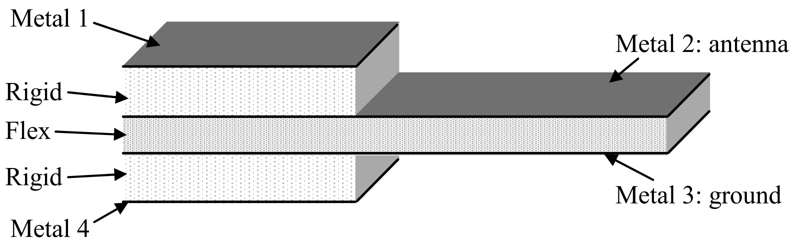
$$n = \sqrt{\varepsilon_r} \tag{2.16}$$

$$v_p = \frac{1}{\sqrt{\varepsilon_r \varepsilon_0}} \tag{2.17}$$

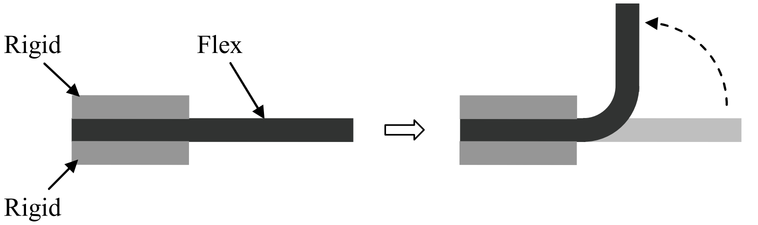
$$v_p = \frac{c}{\sqrt{\varepsilon_r}} \tag{2.18}$$

$$f = \frac{c}{\lambda \sqrt{\varepsilon_r}} \tag{2.19}$$

Conventional printed circuit boards consist of one or more dielectric materials. The dielectric material consists of non-magnetic material with a typical permittivity of 2-5. There are some printed circuit board materials that are soft, e.g., flexible polyimide-based material, so-called flex-foils. The polyimide material is a type of polymer consisting of imide monomers. Recent advances in process techniques make it possible to combine the soft flex-foil material with the traditional printed circuit board material, i.e., the flex-rigid concept as shown in Fig. 1. The main advantage with the flex-foil material is that it can be bent and shaped in many ways. The flexible properties are widely used to create foldable printed circuit boards [8]. In *Paper X* the possibilities of implementing circular monopole and dipole antennas on the flex-rigid structure are studied. For instance, it is shown that a dipole can be placed entirely on the flexible part while its balun is integrated in the rigid part.



(a) Substrate cross-section.



(b) Bendable property.

Fig. 1. Flex-rigid structure: (a) detailed cross-section, and (b) bendable property.

The antenna bandwidth can be defined with respect to several parameters. For UWB antennas the impedance bandwidth is one of the primary ones. There are several ways to define the impedance bandwidth, which means that a requirement is set up for the impedance mismatch, i.e., the reflection measured with a return loss or standing wave ratio (SWR) criteria. Since an absolute bandwidth value says very little about the performance a reference is commonly used, i.e., the bandwidth is set in relation to the center frequency. The center-frequency is then defined linearly or with respect to the geometrical average. Eq. 2.20 shows that the bandwidth ΔBW is a frequency span determined by the upper frequency band limit f_2 and the lower limit f_1. Eq. 2.21 shows the equation for the linear center frequency f_{C_lin}, i.e., the arithmetic center frequency. Eq. 2.22 shows the equation for the logarithmical center frequency $f_{C\_\log}$, i.e., the geometrical center frequency. The bandwidth ratio is then defined by Eq. 2.23, were f_{C_x} represents the linear or logarithmical center frequency. A common way to present relative bandwidth is to express it in percentage as in Eq. 2.24. Eqs. 2.25, 2.26 show complete expressions for the bandwidth ratio, dependant only of bandwidth limits, linearly and logarithmically, respectively [2], [6], [7].

$$\Delta BW = f_2 - f_1 \tag{2.20}$$

$$f_{C\_lin} = \frac{f_1 + f_2}{2} \tag{2.21}$$

$$f_{C\_\log} = \sqrt{f_1 f_2} \tag{2.22}$$

$$\Delta bw_{ratio} = \frac{\Delta BW}{f_{C\_x}} \tag{2.23}$$

$$\Delta bw_{ratio}\% = 100\frac{\Delta BW}{f_{C\_x}} \tag{2.24}$$

$$\Delta bw_{ratio\_lin}\% = 200\frac{f_2 - f_1}{f_1 + f_2} \tag{2.25}$$

$$\Delta bw_{ratio\_\log}\% = 100\frac{f_2 - f_1}{\sqrt{f_1 f_2}} \tag{2.26}$$

In the above section bandwidth definitions are described. Apart from impedance definitions the frequency limits f_1 and f_2 can be defined by radiation and polarization performance definitions [7]. Moreover, different antenna types have different advantages and weaknesses, and therefore different definitions are needed from case to case. In general interesting parameters are the direction of the main beam, lobe levels, beamwidth constraints as the half-power beamwidth, beam coverage and angle constraints, directivity, gain, efficiency, phase linearity, and effective area. The direction of the main lobe is the center of the main radiating lobe. Side, and back lobe ratios are definitions of the degree of suppression of secondary lobes compared to the main lobe. Directivity (D) is a definition of how focused the antenna radiation is, i.e., only the rate of focus because no loss is included. Gain (G) is similar to the directivity, but with the difference that it is the actual radiation strength with a reference, i.e., radiation in a certain direction and at a particular frequency with a reference radiation. For instance, the isotropic gain (dBi) has a reference to be an isotropically (uniformly in all three dimensions) radiating source [2], [6], [7]. As seen in Eq. 2.27 the relation between directivity and gain is also a definition for radiation efficiency (η) [7]. The radiation efficiency can also be seen as a relation in power flow, i.e., between the actual radiated power (P_{rad}) and by the antenna input accepted power ($P_{in\_accepted}$) [7], as seen in Eq. 2.28. A third way of describing the antenna radiation efficiency is to relate to radiation resistance (R_{rad}) and antenna radiation loss (R_L), i.e., the quote between radiation resistance and radiation resistance plus radiation loss as seen in Eq. 2.29. The radiation resistance requires a more detailed description for understanding of the radiation efficiency. A charged electron has because of the charge an own electric field. The field then dynamically creates a force upon the electron itself. As a result the motion of the electron is resisted by the force. Moreover, the related drag force is therefore behind the radiation resistance. Current flowing in opposite direction to the radiation resistance is converted to electromagnetic energy. Moreover, the oscillating electrons collide with atoms during their motion, causing loss in terms of heating or ohmic resistance [7].

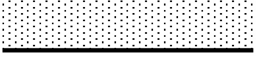
$$\eta = \frac{G}{D} \tag{2.27}$$

$$\eta = \frac{P_{rad}}{P_{in\_accepted}} \tag{2.28}$$

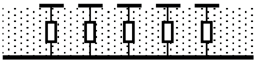
$$\eta = \frac{R_{rad}}{R_{rad} + R_L} \tag{2.29}$$

2.2.2. Parasitics, and Resistive Loading

The main advantage of this type of techniques is that the resistive loading damps reflections and therefore extends the antennas operational impedance bandwidth. The loading element can be a termination or a parasitic dielectric loading applied to the antenna. In this section the dielectric material is utilized for loading purposes. Naturally when an antenna is integrated onto a printed circuit board the dielectric has impact on the antenna as well but the main focus here is when the dielectric material is added for electrical loading purposes and not as physical carrier. The largest drawback of these techniques is the low efficiency due to the fact that signal is consumed as material or termination loss [2], [3]. However, there are some exceptions when resistive loading might be a good choice to damp undesired behavior of the antenna, or to achieve a low input reflection for an electrically small antenna or suppression of undesired back-lobe radiation [2]. For a transmitting antenna the lower radiated power can be compensated by increasing the antenna input power, assuming that the device is not a handheld device limited by battery lifetime requirements. For the receiver side, loss degrades the sensitivity of the antenna and therefore also lowers the signal to noise ratio (SNR) [2]. The technique of resistive loading is definitively a questionable solution in practice but some useful enhancement in term of lowered dispersion by proper loading is possible [3], [4]. The dispersion performance is improved due to lowered reflections from the antenna end or edges. In particular impulse UWB systems are sensitive towards dispersion, i.e., short timed impulses continuously occupies a wide frequency band during the actual transmission [2], [3]. For instance, frequency-independent antennas are known to suffer from dispersion [2], [3]. A dispersive behavior is clearly seen with a transient response, commonly with a monocycle impulse as the response signal [9]. It is shown in [10], [11] that the ringing effects in a bowtie antenna can be reduced with resistive loading. For impulse UWB communication the ringing is important due to that ringing increase the lowest possible cycle repetition time [2], [10], [11]. These types of antennas are interesting in ground penetrating radar systems [10], [11]. Restive loading can be achieved in many ways; resistors can be placed within the substrate with modern printed circuit board technology as proposed in [12], see Fig. 2. As an example it is shown that backward traveling waves can be minimized in an Archimedean spiral antenna.



(a) Unloaded microstrip lines.

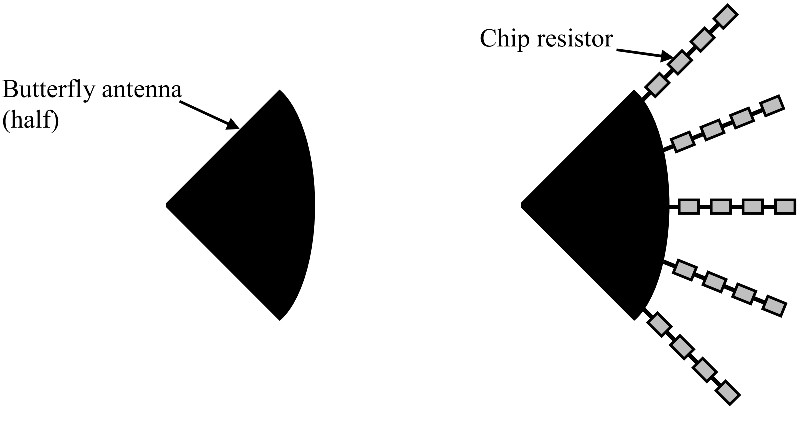


(b) In substrate loaded microstrip lines.

Fig. 2. Inside substrate loading: (a) microstrip lines without loading, and (b) microstrip lines with in-substrate loading.

Enhancement of the receiver sensitivity in impulse systems of both electrical and magnetic dipole antennas is possible with suitable loading [13]. Resistive coupling also affects the mutual coupling in antenna arrays [14]. It is shown in

[14] that the radiation efficiency of a dual dipole antenna array is directly dependant of the resistive loading on the dipole antenna arms. As shown by a number of authors above the reflection can be reduced with resistive loading in various antenna technologies. Furthermore, other effects as energy and radiation concentration for various antenna geometrics when loaded and not loaded resistively have also been studied. For instance it is shown in [15] that the effect of loading varies with the geometric ratio. It is also shown that bowtie and butterfly antennas with narrow-angled thin geometry exhibits an increased radiation focus when resistively loaded, while wide-angled broad geometries exhibit a decreased focus [15]. Figs. 3a and 3b show a radial stub (butterfly antenna half) without and with resistive loading, respectively.



(a) Without loading. (b) With loading.

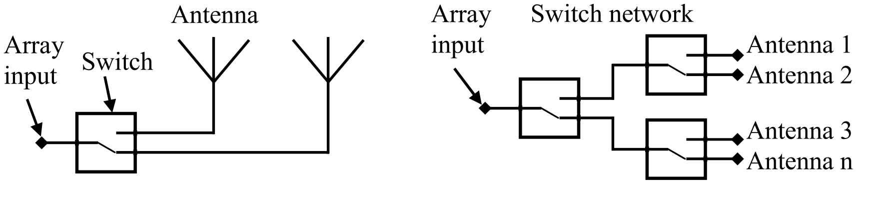
Fig. 3. Radial stub (Butterfly antenna half): (a) without loading, and (b) with loading.

Transverse electromagnetic (TEM) horn antennas are in general quite broad-banded. However, utilizing dielectric loading the fractional bandwidth of the antenna can be increased and potential ringing effects are minimized [16]. Introducing extra loss through loading should not be seen as the primary antenna design solution to reach high impedance bandwidth but should still be seen as a possible path if the enhancements justify the drawbacks. It should be mentioned that the best performing UWB antennas currently available in terms of low reflection on the market has some type of loading [2]. Such antennas can have a reflection of -30 dB or lower across decades of frequency bandwidth [2].

2.2.3. Multi- Band and Resonance Antenna Systems

An antenna array is formed by combining an arbitrary number of antenna elements. The difference between this category and the previous technique to extend the bandwidth using parasitic coupling is that in this category an arbitrary number of radiating antenna elements or antenna components are used in multi-resonant structures. The structure can be fed directly with electrical interconnects or with aperture feeding. Impedance match with electrical feeding systems is achieved using switches, power dividers or using multiplexing techniques [17], [18]. Furthermore, multiple antennas can be used to enhance performance, e.g., mitigation of narrowband interferers [19].

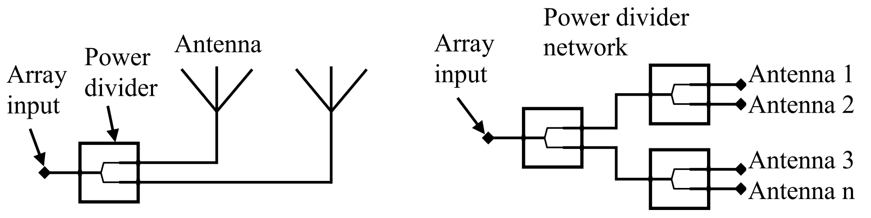
A system based on switching selects the operative radiating antenna element by switching the electrical path as shown in Fig. 4. The switching time and switch noise are two important factors. The switch settling time must be fast enough for the application to guarantee a matched impedance route during the transmission. Moreover, it is desired that the switching structure does not noticeably increase noise in any way [17].



(a) Two antenna array. (b) Switch network.

Fig. 4. Switched antenna system: (a) a two antenna array, and (b) a four output port switch network.

Using power dividers as shown in Fig. 5 are a well known solution to electrically combine several radiating antenna elements. Array structures [17] have been used for many purposes, e.g., to create a linear array with a focused radiation, and beam steering, etc. Each signal that is divided equally by two is lowered by 3 dB plus loss in the power divider compared to the original signal. Naturally when the signal is spread to many radiating antenna elements the signal received at each antenna is severely weakened. Moreover, compared to switched antenna systems power dividers do not need any settling time.



(a) Two antenna array.

(b) Switch network.

Fig. 5. Antenna array using power dividers: (a) two antenna array, and (b) power divider network.

A compromise is to use sub-arrays built with power dividers that then are combined with switches, as proposed in *Paper I*. The respective sub-arrays typically consist of one to four radiating antenna elements, an arbitrary number of sub-arrays can then be combined with a switching network [17]. Therefore as seen in Fig. 6 power dividers and switches can be used to combine an arbitrary number of narrowband antennas in parallel in order to create a wideband multi-band antenna system [17]. Furthermore, the number of switches and power dividers can be choosen freely, and the system does not need to be symmetrical.

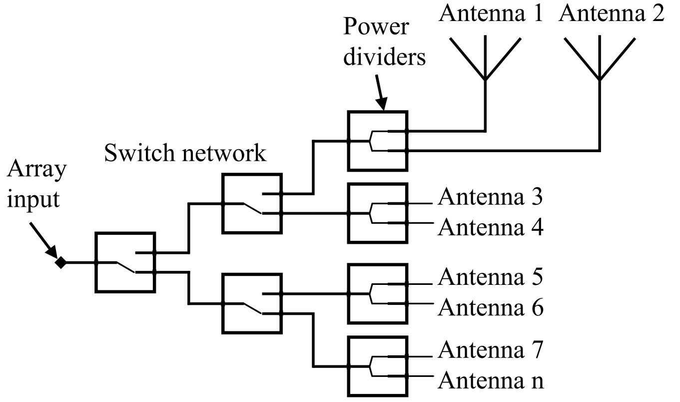
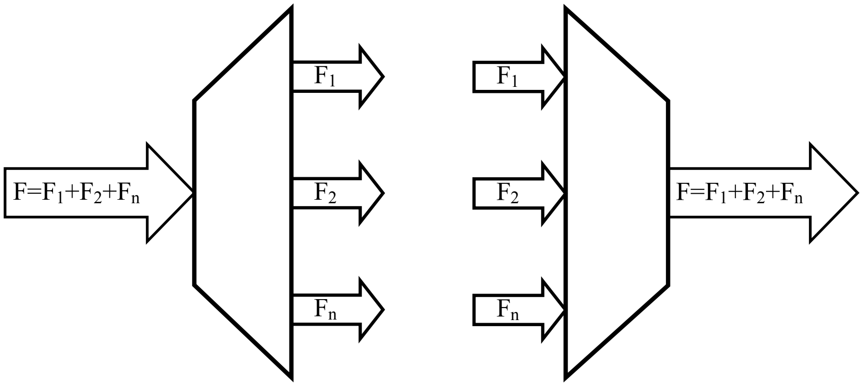


Fig. 6. An antenna array consisting of switches and power dividers.

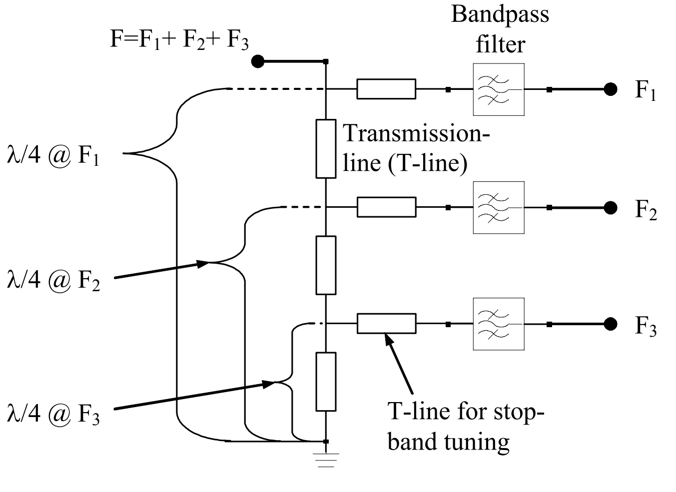
In a similar manner the dividers and switches can be replaced with a frequency multiplexing technique in order to build a broadband and multi-band antenna system [18]. The multiplexing technique has the advantages of both a switched antenna system and a power-divided system, i.e., no settling-time compared to a switched system and the signal is frequency-selected to the desired antenna element. Fig. 7 shows the principle of frequency multiplexing. Fig. 7a shows how the frequency band F is demultiplexed into the three parallel sub-bands F_1, F_2, F_3, while Fig. 7b shows the opposite way when doing a multiplexing operation.



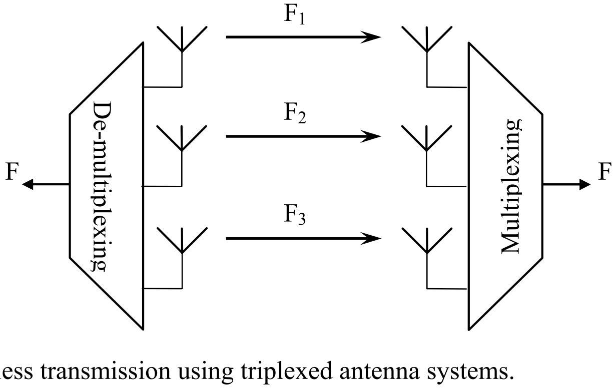
(a) Demultiplexing. (b) Multiplexing.

Fig. 7. Frequency multiplexing: (a) demultiplexing, and (b) multiplexing.

In *Papers V* and *VII* it is shown that multiplexing techniques can be used to obtain frequency-selective antennas. *Paper V* presents the concept of frequency-multiplexed inverted-F antennas, and *Paper VIII* presents a fully integrated printed circuit board triplex antenna system suitable for Mode 1 UWB (3.1-4.8 GHz). The proposed antenna system consists of three antennas, three bandpass filters and a frequency multiplexing network. It is shown that inverted-F antennas can be combined in parallel using the frequency multiplexing technique. The technique is demonstrated using edge-coupled filters in *Paper V*, and using combined broadside- and edge-coupled filters in *Paper VIII*. Fig. 8a shows a schematic of a triplexer. The network is realized with a microstrip technology. The triplexer consists of three series quarter-wavelength transmission lines, three bandpass filters, and three transmission lines for tuning of the filter impedance. The series transmission lines provide a high impedance at the respective frequency band. The filter tuning lines optimize the stop band impedance of each filter, to provide a high stop band impedance in the neighboring bands. The network is optimized together with the filters to achieve flat passbands and a symmetric performance between the sub-bands. Fig. 8b shows a wireless communication utilizing frequency-triplexed transmitter and receiver antennas. Moreover, either side can be exchanged for a single antenna operating in the same frequency band while having three signal sub-channels.



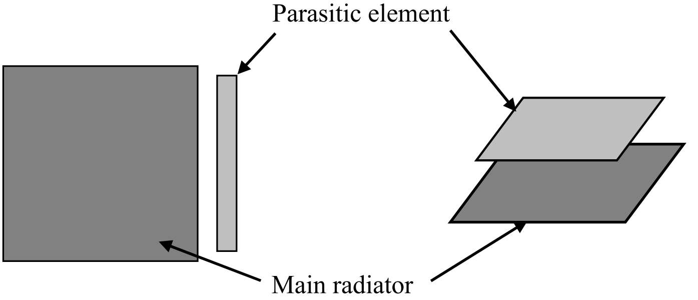
(a) A triplexer network.



(b) Wireless transmission using triplexed antenna systems.

Fig. 8. Frequency multiplexing network: (a) triplexer network, and (b) wireless transmission using triplexed antenna systems on both transmitter and receiver side.

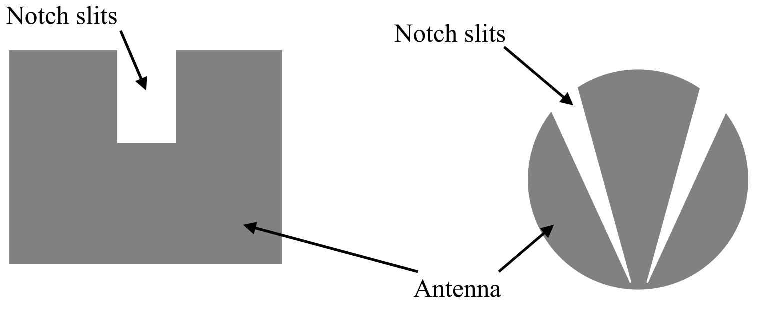
Another method is that one or several parasitic elements are added to enable a wider bandwidth, i.e., more possible resonances. This means that the antennas are equipped with one or several extra resonators such that each one is tuned to resonate in a nearby frequency, i.e., the parasitic elements are fed by the main antenna element due to electromagnetic coupling [20]. The passive radiators can be used both to increase the impedance bandwidth and to strengthen suppression of notches [20]. Furthermore, parasitic elements are widely used to control the beam pattern, e.g., a more focused pattern as with the Yagi antenna [7]. Fig. 9 shows two typical configurations for parasitic elements used as passive radiators. The passive radiators can be placed side by side with the driving antenna or stacked in a multi-layer construction. The electrically fed main radiators are coloured dark grey while the passive radiators have a lighter grey colour.



(a) Side by side radiators. (b) Stacked radiators.

Fig. 9. Parasitic elements: (a) side by side, and (b) multilayer stacked.

A typical multi-band antenna is composed of several antennas or sub-band components, joined together. However, sometimes the objective is more or less the opposite; one wideband antenna for multi-band operation is desired. If some bandwidth need to be blocked for preventing interferences, frequency notching techniques shown in Fig. 10 are used to create small notches that block undesired bands, i.e., to divide the wide frequency band into two or more narrow sub-bands [21]. Usually a half wavelength resonant structure is used to create one notch, i.e., to create a high VSWR at the desired frequency. For instance, circular and elliptical dipoles can be equipped with notching characteristics, with triangular or elliptical half-wavelength cuts [21]. Furthermore, in a similar manner narrow-band notching can be achieved in a slot antenna technology [22]. With additional frequency-compensation stubs attached to a square dipole antenna, bandwidth limitation, and control of the antenna is possible [23]. With U-shaped slots a planar monopole antenna can have multi-band abilities]. One wideband planar monopole antenna and three U-shaped slots results in a frequency band with two band limiting notches [24]. Fig. 10 shows two notched antennas [24]. The illustrations show the principle that slits can be cut out in resonating antenna elements in order to divide or broaden the bandwidth.



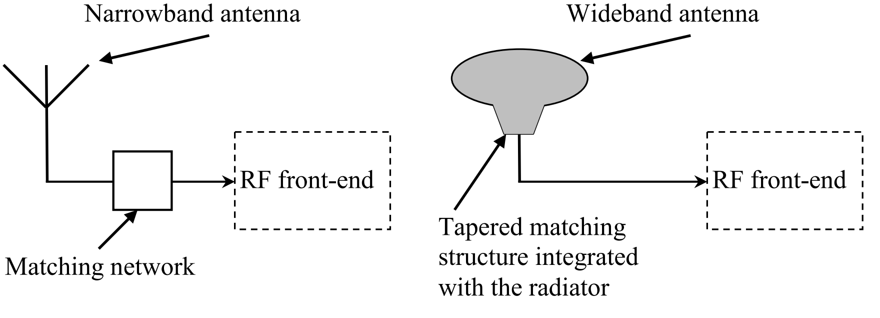
(a) Notched square patch antenna. (b) Notched circular antenna.

Fig. 10. Notched radiators: (a) square patch with a rectangular notch, and (b) circular monopole antenna with two triangular notches.

Multiple antenna solutions can also be used in a more traditional antenna-array style to ensure high gain throughout the whole frequency-band [25]. Several antenna elements can form a wideband multi-band antenna array with a common junction matched to 50 Ω or other desired input impedances [17], [18]. In general multi-band solutions create freedom to combine antennas for different services to one transceiver provided that a proper isolation between the antenna elements is preserved.

2.2.4. Wideband Impedance Matching Through Geometrical Control

It is possible to transform a well known, well defined, and narrowband antenna structure into a geometrical smooth wideband structure [2]. The principle is quite simple, although in the same time it is infinitively complex to formulate. The whole antenna structure, all the way from the feed-line to the radiating elements, shall be designed such that the impedance transformation is done smoothly from the RF front-end impedance to the free space impedance [2], [7]. The goal is not to create a wideband matching for a narrowband antenna but to create a wideband matched antenna [2]. The antenna becomes a virtual taper that aims to preserver smooth impedance transition throughout the entire frequency band. A clear physical example of the philosophy is to compare a classical straight thin-wire monopole or dipole antenna with a circular, elliptical, or other smooth geometry monopole or dipole antenna [2]. The straight wire implementation has a distinct physical length, i.e., a physical length that favors operation at a specific narrow frequency range. The distinct length makes the antenna relatively frequency dependant [7]. The antenna impedance bandwidth is therefore limited to a frequency-range where the electrical length has a minor deviation from the optimal wavelength condition. This kind of geometrical solutions varies from a classic square or circle to almost any geometrical shape. The common philosophy behind them is that they rely on area shapes, i.e., area geometries that allow the input impedance to be controlled over a wide frequency bandwidth. Circular, elliptical, triangular, square, bulb, and heart-like structures have been proposed to give some examples [26]-[33]. Fig. 11 shows a comparison between a typical narrowband antenna and a geometrically controlled wideband antenna. The narrowband antenna is matched at a small frequency range using lumped or distributed matching networks [7]. The wideband antenna has an integrated tapered-matching due to the geometrically controlled design.



(a) Narrowband antenna. (b) Wideband antenna.

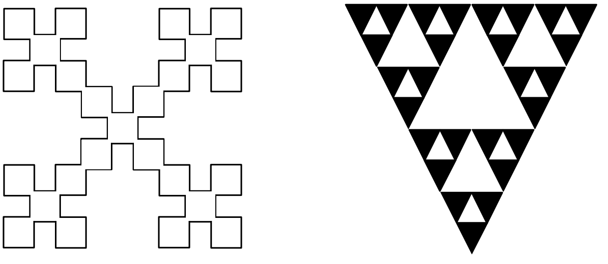
Fig. 11. Antenna systems: (a) a narrowband antenna implementation, (b) a wideband antenna implementation.

2.3. A Summary of UWB Antenna Technologies

The field of proposed UWB antenna solutions reaches from large bulky three dimensional solutions to small planar electrical monopoles [2], [34], [35], and [37]. Frequency-independent antenna structures, magnetic antennas, and slot solutions are some other interesting categories. A major filed for UWB short range communications is for indoor applications, where an omnidirectional radiation pattern is desired [34], [36]. However, in point to point links a high directivity antenna is more suitable. Horn and reflector antennas are typical antennas that have such properties [2]. Moreover, the technologies that are most relevant for this thesis are prioritized, i.e., planar structures suitable to be printed on printed circuit boards.

2.3.1. Frequency-independent Antennas

Frequency-independent antennas rely on a scaled geometry, such that the operational center changes with the frequency, i.e., a frequency-independant physical structure. The major advantage is that the frequency bandwidth is limited only by the degree of scaling. The largest size sets the lower frequency limit, while the minimum physical size defines the upper frequency limit. The major disadvantage is that since the operational center changes, the phase center changes as well [2], [7]. The moving phase center causes dispersion, i.e., non-linear phase behavior [2]. Some transceiver architectures are sensitive towards this kind of distortions. Typical frequency-independent antennas are the spiral antenna, fractal based antennas, and logarithmic periodic antennas [2], [7]. Back in the 1970s, B. Mandelbrot [7] defined the term “fractal” to be any set of geometrical objects that have self-similar shapes. A basic property is that a fractal must have a fractional dimension. Mathematically this means how effectively the object fills space. A fractal curve fills the space better than any classical Euclidean surface (a plane solid surface obeying Euclid’s postulates), and therefore this property reduces the antenna size relative to the Euclidean surface [7]. Figs 12a and 12b show two examples of fractal antennas, a Fractal loop antenna and a Sierpinski antenna, respectively. The Fractal loop antenna consists of squares that are repeated with a self-similar pattern, while the Sierpinski antenna consists of triangles that are repeated likewise. The fractal loop antenna is a three times iterated fractal antenna, i.e., the first iteration from one single square, the second from five squares joint, and the third iteration from five five-squares joint together. The Sierpinski antenna shown is a four times iterated antenna, i.e., one triangle, three triangles, three three-triangles, and nine three-triangles, respectively. The principle of operation is that regardless if the antenna is built up with squares or triangles geometries each iteration creates a unique antenna, and the next iteration is a self-scaled version of the previous iteration. Moreover, all the iterations have their own resonant frequencies [7], e.g., for the Sierpinski antenna this means that each iteration is a unique bow tie antenna.



(a) Fractal loop antenna. (b) Sierpinski antenna.

Fig. 12. Fractal antennas: (a) Fractal loop antenna, and (b) Sierpinski antenna.

Another frequency-independent antenna is the spiral antenna; the equi-angular spiral antenna is shown in Fig. 13. The most important property of the equi-angular spiral antenna compared to non-logarithmic spiral antennas is that the antenna better remains frequency-independent when it is truncated [7].

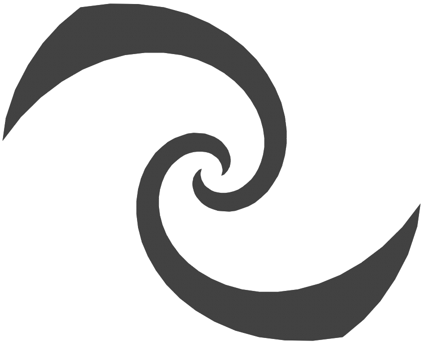
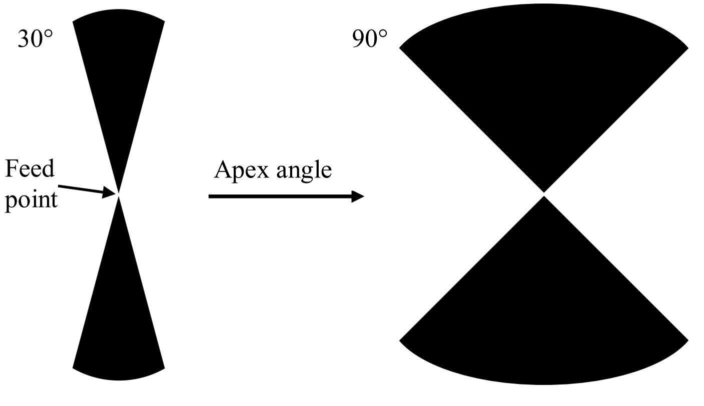


Fig. 13. Equi-angular spiral antenna.

2.3.2. Electrical Antennas (Small Element)

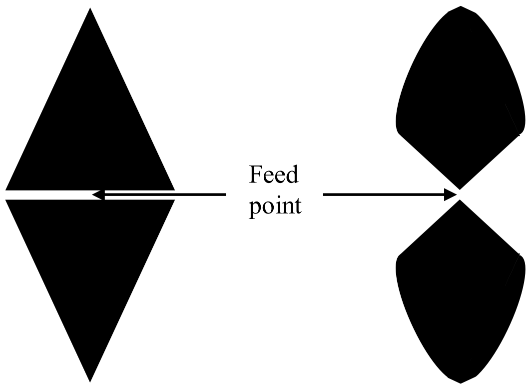
The group of electrical antennas contains a large geometrical variety of physical implementations. The radiation pattern is in general rather omnidirectional, and therefore the antennas are suitable for many types of wireless communications. Implementations are common in both three dimensional and two dimensional geometries. Due to the omnidirectional pattern and better phase behavior than for instance frequency-independent antennas this group contains some of the most important UWB antennas [7], [27]-[31]. Electrical antennas are voltage driven antennas that may have many shapes to achieve desired impedance matching, radiation pattern, and polarization. Conical, circular, bulbous, and squares are some common shapes used in broadband communications. The principle is that curved geometries create wideband tapered impedance-matching [2], [7]. The antenna realization may be a dipole as well as a monopole with ground-plane. A planar cone (bow tie) antenna has its volumetric equivalent [2]. The oldest known electrical antenna is Lodges conical antennas reported back in the 1890s [2]. Moreover, Lodge introduced the planar bow tie antenna at the same time [2]. Figs. 14a and 14b show a bow tie antenna, with an apex angle of 30º and 90º, respectively.



(a) 30º Apex angle. (b) 90º Apex angle.

Fig. 14. Bow tie antennas: (a) antenna with 30º apex angle, and (b) antenna with 90º apex angle.

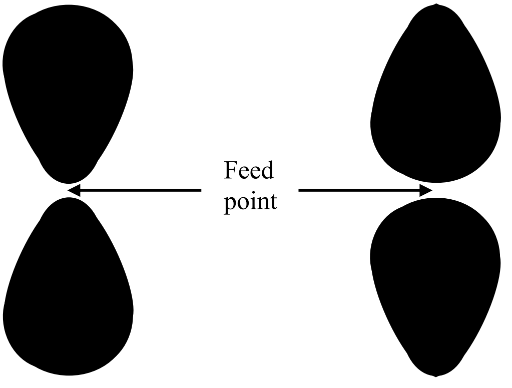
The inversed bow-tie more known as the diamond dipole antenna was introduced by Masters 1947 [2]. Other common implementations of the bow tie or diamond structures are the Bishop's hat, and the hexagonal dipole. Fig. 15a shows a diamond dipole antenna, and Fig. 15b shows a Bishop's hat antenna. The major reason for these modified diamond dipoles is that the traditional bow tie is known to difficult too match severely limiting the practical usefulness of the antenna [2], [31]. The Bishop's hat is to be seen as a tapered diamond dipole where the integrated taper to the antenna provides a wideband matching.



(a) Diamond dipole antenna. (b) Bishop's hat antenna.

Fig. 15. Diamond dipole antenna: (a) a diamond dipole antenna, and (b) a Bishop's hat antenna.

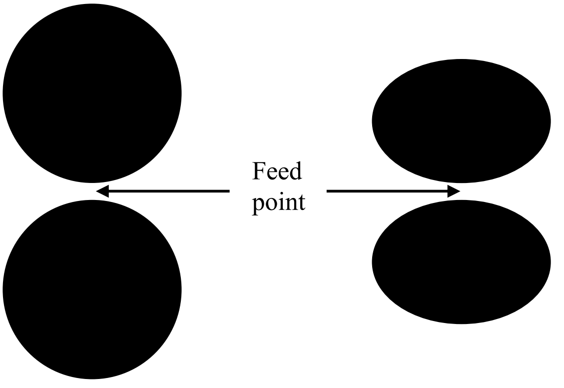
Another important antenna type in the art of electrical antennas is the bulbous antenna. The name originates from the simple fact that some of the first antennas within this category had bulbous silhouettes. Two typical bulbous antennas are shown in Fig. 16. Today several circular, elliptical, and bulbous shapes are classified as bulbous antennas [2], [26]-[31]. The center resonance frequency of bulbous antennas depends not only on the antenna length but also on the circumference [7]. This property gives the advantage of a more compact shape relative to a narrowband straight wire dipole antenna [2]. In fact as discussed in section 2.2.4 the operational frequency band is not dependant only of a distinct arm-length anymore. Instead the bandwidth is more defined by the surface revolution [2]. Naturally since the circumference distance from the feed-point to the far-side is longer than a straight line, a bulb shaped geometrical antenna has a lower operational frequency than a straight wire antenna with the same length. Lindenblad's television antenna is known as the original bulbous antenna. The antenna has a reported VSWR lower than 1.1 impedance bandwidth of 24 MHz at a center frequency of 164 MHz, designed to cover four 6 MHz television channels [2].



(a) Bulbous dipole antenna I. (b) Bulbous dipole antenna II.

Fig. 16. Two typical bulbous dipole antennas: (a) bulbous dipole antenna I, and (b) bulbous dipole antenna II.

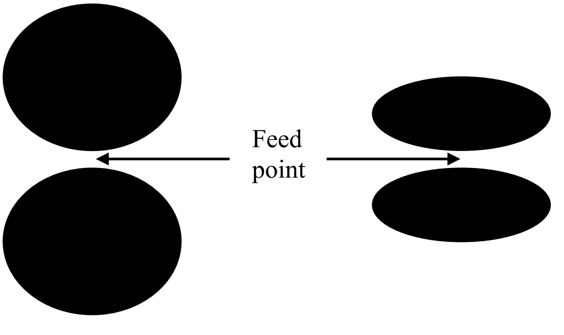
Another sub-category of bulbous antennas investigated for UWB systems are covers circular, and elliptical antennas. Circular, and elliptical shapes have been proposed both as monopoles and dipole antennas for UWB. The circular shape versus the elliptical shape is a trade-off between radiation pattern and the impedance matching. The circular shape is symmetrical and therefore has a relatively uniform radiation pattern, i.e., an omnidirectional pattern. For the elliptical antennas the radiation pattern becomes less uniform as the eccentricity increases [2], [7]. Fig. 17 shows a circular dipole antenna and an elliptical dipole antenna.



(a) Circular dipole antenna. (b) Elliptical dipole antenna.

Fig. 17. Dipole antennas: (a) circular dipole antenna, and (b) elliptical dipole antenna.

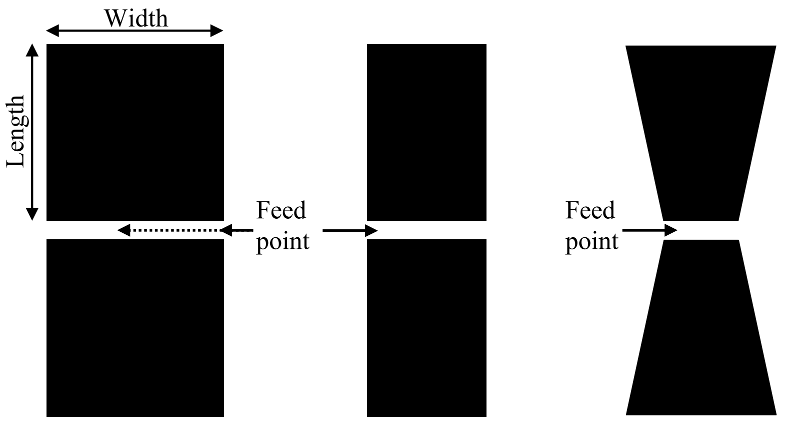
Ellipses with low eccentricity are referred to as fat ellipses while really oval shaped discs are known as skinny ellipses. Fig. 18a shows a fat ellipse, and Fig. 18b shows a skinny ellipse. Depending on parameter priorities different optimal ratios can be chosen. As mentioned earlier a more circular structure has a more uniform radiation.



(a) Fat ellipse. (b) Skinny ellipse.

Fig. 18. Elliptical dipole antennas: (a) fat ellipse, and (b) skinny ellipse.

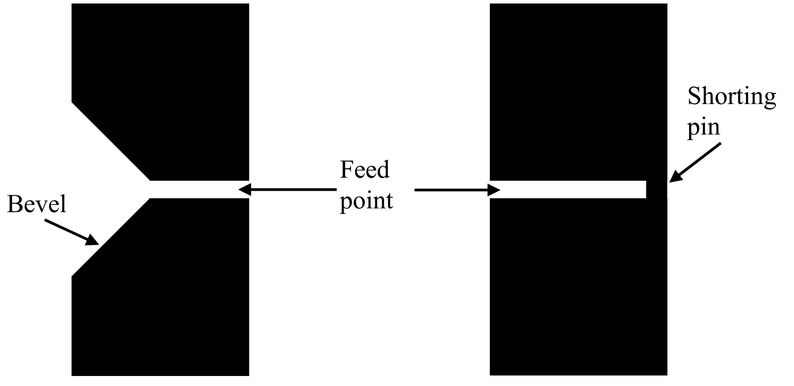
Rectangular or square (when symmetrical) antennas are another set of antennas that have been explored for UWB. Monopole and dipole designs have been presented [2], [6], [38]. In a typical monopole antenna implementation on PCB the ground plane is placed side by side with the radiator [6], and not to be mixed with the microstrip patch antenna. A planar square dipole antenna may have more than 40% relative bandwidth with respect to its arithmetic center frequency [6]. A square monopole and dipole can be seen as a thick version of a thin wire monopole and dipole, respectively. However, with the difference that the thin wire antennas rely on one resonance determined by its length and the square antennas have a bandwidth determined by the square geometry. Moreover, the square antenna has a lower edge frequency of bandwidth that is dependant of the antenna length and width. Figs. 19a, 19b, and 19c show a square, rectangular, and a trapezoidal antenna, respectively.



(a) Square. (b) Rectangular. (c) Trapezoidal.

Fig. 19. Rectangular, square antennas: (a) square, (b) rectangular, and (c) trapezoidal.

The square antenna has a relatively high relative bandwidth. However, efforts have still been done to extend the antenna performance. Several modifications have been suggested for the square antenna to improve the impedance bandwidth, e.g., a bevel cut, shorting pins, and slot hole in the antenna [6]. Figs. 20a and 20b show a square antenna with a bevel and antenna with a shorting pin, respectively. A bevel may be used in any of the square antenna corners to make a desired change to the input impedance [6]. A slot hole in the antenna will steer the current distribution through the radiating element, and therefore also change the input impedance [6], [7]. A square antenna that has 40% relative bandwidth without any modification can at least have 50% with a bevel or any other of the suggested modifications [6].



(a) With a bevel (b) With a shorting pin.

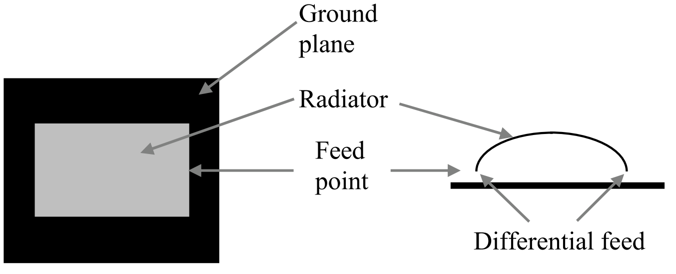
Fig. 20. Square antenna: (a) with bevel, and (b) with a shorting pin.

2.3.3. Magnetic and Slot Antennas (Small Element)

The magnetic antennas are current driven. The antennas rely on one or more current loops and have dominant magnetic near-fields. Magnetic fields couple less than electric fields therefore magnetic antennas are popular in embedded systems [7], [2]. Small element magnetic antennas can be seen as a physical implementation of magnetic Hertzian dipoles [2]. If an ideal electrical Hertzian dipole is aligned so that it generates an omnidirectional vertical polarization pattern the corresponding ideal magnetic Hertzian dipole generates a horizontal polarization pattern [2], [7]. Typical magnetic antennas are large current radiators, mono-loops, loops, and slot antennas [2], [7].

Fig. 21 shows a typical large current radiator antenna. The large current radiator antennas consist of a relatively large conducting sheet above a ground-plane. The conducting sheet is fed differentially Differential feeding is done on two opposite sides, usually the two short sides in a rectangular geometry. There are two documented issues with the large current radiating antennas. First as with all conducting sheets, current tends to concentrate on the edges, i.e., making it difficult to achieve a uniform current distribution in the sheet. However, it has been reported that cutting the sheet into parallel sections reduces this issue [2].

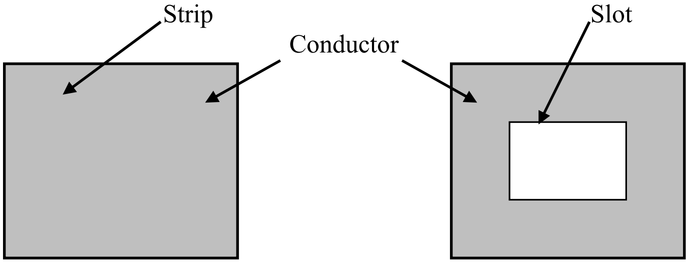
The second issue is that magnetic field is equally generated on each side of the metal, causing the problem that up to half the magnetic field energy is trapped by the ground plane. Naturally this impacts negatively on the antenna efficiency since the resulting narrowband resonances must be damped with some type of loss mechanism [2], [7].



(a) Top view. (b) Side view.

Fig. 21. A large current radiator: (a) top view, and (b) side view.

As shown in the previous sections printed strips are used in many ways to build frequency-independent, electrical, and magnetic antennas. The difference between a strip and a slot is shown in Fig. 22. More or less the physical opposite of a strip is a slot. A slot is a hole cut out in the metal to form an aperture that shapes the current flow.



(a) Strip. (b) Slot.

Fig. 22. Strips and slots: (a) a strip, and (b) a slot.

Fig. 23 shows a λ/2 slot antenna in an infinite ground plane. This type of slot antenna is virtually a dual of a dipole antenna. In the slot antenna magnetic fields replace the electric fields of a dipole antenna. The radiation from the short edges of the slot can be neglected due to the fact that the width is much smaller than the slot length and one wavelength [7]. Moreover, due to the same reasoning the current flow at the long-sides are parallel and in counter phase, i.e., the fields are opposite directed and cancel out each other. The net radiation of the slot antenna comes from the current that is spread out, away from the slot [2], [7]. The current a distance from the slot gives rise to vertical H field. For highest possible radiation efficiency the ground plane should be infinitively large. Reasonable implementation that still provides a high efficiency is a ground plane of one wavelength around the slot [7].

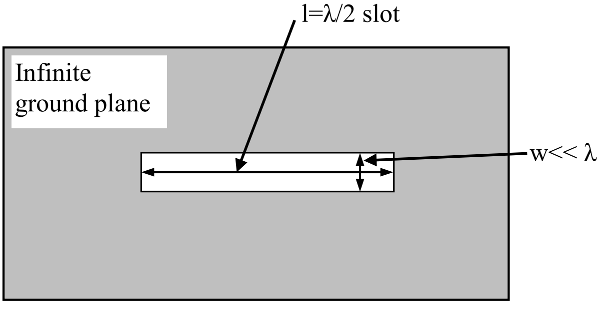


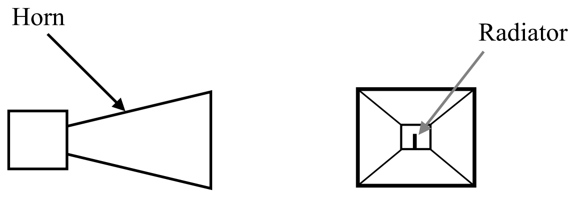
Fig. 23. A λ/2 slot in an infinite ground plane.

2.3.4. Horn and Reflector Antennas

The term omnidirectional is widely used in the context of UWB. This is mainly due to the large interest of high speed short range indoor communications. In indoor communication the devices may have any direction relative to each other, i.e., the communication should not be direction-dependant. However, there are other applications where it may be preferable to direct the radiation/ sensitivity, e.g., point to point links [39], [40]. Horn and reflector antennas have in general high directivity owing to their physical nature [7], [39]. The horn only receives signal from the direction which the horn points, and the reflector antennas only receive signals from the opposite side of the reflector.

Horn antennas are widely used in radiation measurements due to their low level of cross polarization, and linear gain [7]. Fig. 24 shows a horn antenna. Figs

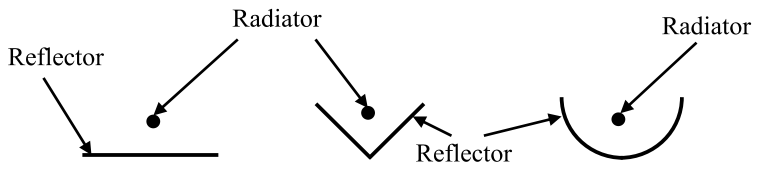
24a, 24b show the side and front view, respectively. The horn is a tapered transmission line that provides a stable gain over a wide frequency band [7].



(a) Side view. (b) Front view.

Fig. 24. A Horn antenna: (a) side view, and (b) front view.

A reflector antenna consists of a radiating element that is backed up with a reflector. For instance, the reflector can be a planar surface, a corner, or a parabolic structure. The choice of reflector determines the field of view, i.e., the degree of signal concentration [7]. Figs. 25a, 25b, and 25c show a planar, a corner, and a parabolic reflector, respectively.



(a) Planar. (b) Corner. (c) Parabolic.

Fig. 25. Reflector antennas: (a) planar, (b) corner, and (c) parabolic.

2.4. UWB antenna considerations

The various UWB regulations presented later in chapter 4 show that the performance requirements of a UWB antenna is high. For instance, utilizing the entire unlicensed band 3.1-10.6 GHz in USA the bandwidth will be 7.5 GHz at the CF of 6.85 GHz. That is more than 100% of the fractional bandwidth. The most fundamental requirement for the UWB antenna is impedance control over the allocated frequency band, i.e., VSWR within the operational bandwidth. The antenna is said to have a certain impedance bandwidth at a specified reflection constraint. For an optimal impulse transmission, an ideal UWB antenna should produce radiation patterns with constant magnitude and a phase shift that varies linearly with frequency. Naturally a high radiation efficiency, high gain relative to the directivity and omnidirectional pattern are desired. No antenna technique can at once fulfill all the requirements without compromises, especially if a low profile antenna structure is chosen. If there are no volume constraint few or none performance compromise has to be done. It is generally known that if size is no issue then the antenna designer's degree of freedom is high [2]-[7], [41]. Both Multi-band OFDM and DS-UWB proposals split the spectrum into several bands. One reason is that a frequency-divided spectrum can be used to allocate slots for users. The second reason is that systems with different maximum capacities can be created, i.e., using a smaller band to reduce spectral requirements, either to reduce the maximum number of users or reduce the capacity for each user. An antenna that covers the entire 3.1-10.6 GHz band can be used by any system, but is not always the optimal choice. The various choices in spectrum make several multi-antenna solutions interesting. A narrower band gives the possibility to either optimize other performance parameters or to reduce the physical size. The receiver sensitivity is dependant of the antenna gain and the SNR. For really high data rate, a high SNR is required by any UWB technology. For indoor communications a wide omnidirectional pattern is desired, along with a flat gain response, independent of the frequency [42]-[46]. The Multi-band OFDM UWB has, because of the OFDM technique, a high robustness against variations in gain, phase, and group delay that the antenna may cause [4], [5], [44]-[46]. Systems using OFDM therefore is less sensitive towards phase-dispersive antenna technologies, i.e., the phase nonlinearity within each carrier is the limitation and not the system frequency bandwidth. Meanwhile, impulse based UWB systems are sensitive towards ringing phenomena in the time domain, i.e., phase nonlinearities in the frequency domain. Therefore frequency-independent antennas with a varying phase center are not suitable for impulse communication which simultaneously uses a large frequency-band. However, such antennas may still be used in very wide-banded systems if simultaneously only a smaller sub-part of the whole frequency-band is used during each transmission, i.e., a multi-band system, preferably utilizing the OFDM technique. The phase linearity is important only within the sub-band and not the entire frequency band [4], [42].

3. Types of Antennas Used in This Work

The variety of antenna technologies is as large as the variety of wireless technologies. For UWB applications the fractional bandwidth limits the choices of antennas, and four different antenna technologies are discussed here.

3.1. Patch Antenna

Fig. 26a shows the right oriented coordinate system used for the antenna illustrations shown in Figs 26b and 26c. Fig. 26b illustrates a half wavelength patch antenna and Fig. 26c a quarter wavelength patch antenna. The dimension definitions and the feed point are illustrated. The length L determines the resonance frequency with the chosen material. The phase velocity v_p is dependant of the material. The width w affects the impedance at the edge of the antenna, which affects the position of the feed point to achieve input matching. The half wavelength patch antenna is symmetric along the length axis, with zero input impedance at the center and two equal edge impedances. The edge impedance can vary up to a couple of hundred ohms. It is dependant of the material, substrate height and the patch width. For the quarter wavelength patch antenna one of the edges are connected to ground. The quarter wavelength patch antenna works similarly to the half wavelength patch antenna, but the resonance length is equal to one-quarter wavelength instead of a half wavelength [7], [47]-[52].

Eq. 3.1 shows the relation between the patch length L, effective permittivity ε_{eff} and wavelength λ_0. Calculation of ε_{eff} is shown in Eq. 3.2. In the equations ε_r is the relative permittivity, λ_0 is the wavelength at the resonance frequency, h is the substrate height, and w is the patch width. The $\Delta\ell$ term in Eq. 3.1 is the fringe field length. The fringe field between the patch edge and the ground plane slightly reduces the needed length for a specific resonance frequency, determined by Eq. 3.3. Eq. 3.4 is an empirical formula to calculate the width for near optimum radiation [7], [47]-[49].

$$L = \frac{\lambda_0}{2\sqrt{\varepsilon_{eff}}} - 2\Delta\ell \tag{3.1}$$

$$\varepsilon_{eff} = \frac{\varepsilon_r + 1}{2} + \frac{\varepsilon_r - 1}{2}\left[1 + \frac{12h}{w}\right]^{-1/2} \quad \text{for } w/h > 1 \tag{3.2}$$

$$\Delta\ell = 0.412h\frac{(\varepsilon_{eff} + 0.3)(w/h + 0.264)}{(\varepsilon_{eff} - 0.258)(w/h + 0.8)} \tag{3.3}$$

$$w = \frac{\lambda_0}{2}\frac{1}{\sqrt{\frac{\varepsilon_r + 1}{2}}} \tag{3.4}$$

To achieve the best possible performance the patch antenna must be matched by calculating the matched feed point, see Figs. 26 and 27. Eq. 3.5 shows the input admittance, indicated by the transmission line model shown in Fig. 26d. In Eqs. 3.5-3.8, ε_{eff}, ε_r, λ_0, h, and w are specified same as in Eqs. 3.1-3.4. Furthermore, z is the distance from the edge to the feed point, Y_{in} is the input admittance, Y_0 is the characteristic admittance, Z_0 is the characteristic impedance, η_0 is the wave impedance in free space, k_0 is the free space wave number, and β is the wave number with respect to ε_{eff}. Eq. 3.5 together with Eqs. 3.6-3.10 is used to calculate the matched feed point. G is the radiation conductance as shown in Fig. 26d, and given by Eq. 3.6 for small heights. The characteristic admittance and impedance of the patch are given by Eqs. 3.7 and 3.8. The slot reactance jB shown in Fig. 26d is given by Eq. 3.9. The wave numbers k_0 and β are given by Eqs. 3.10 and 3.11 [7], [47]-[49].

$$Y_{in}(z) = \frac{2G}{\cos^2(\beta z) + \frac{G^2 + B^2}{Y_0^2}\sin^2(\beta z) - \frac{B}{Y_0}\sin(2\beta z)} \tag{3.5}$$

$$G = \frac{1 - \frac{(k_0 h)^2}{24}}{120\lambda_0} \tag{3.6}$$

$$Y_0 = \frac{1}{Z_0} \tag{3.7}$$

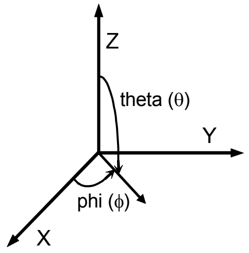
$$Z_0 = \frac{\eta_0}{\sqrt{\varepsilon_{eff}}\left(\frac{w}{h} + 1.393 + 0.667\ln(\frac{w}{h} + 1.444)\right)} \tag{3.8}$$

$$jB \approx jk_0 \Delta\ell \frac{\sqrt{\varepsilon_{eff}}}{Z_0} \tag{3.9}$$

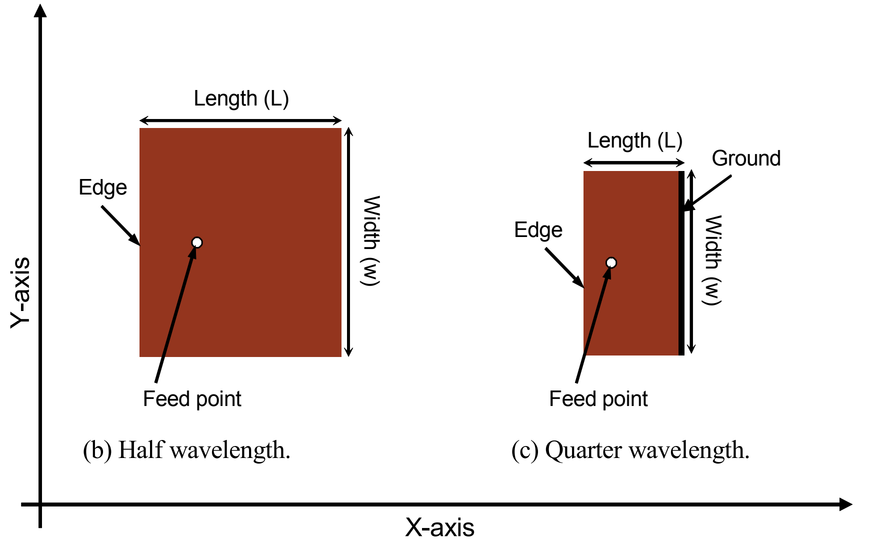
$$k_0 = \frac{2\pi}{\lambda_0} \tag{3.10}$$

$$\beta = \frac{2\pi}{\lambda} = 2\pi\sqrt{\frac{\varepsilon_{eff}}{\lambda_0}} \tag{3.11}$$

$$\eta_0 = \frac{\sqrt{\mu_0}}{\sqrt{\varepsilon_0}} = 377\,\Omega \tag{3.12}$$

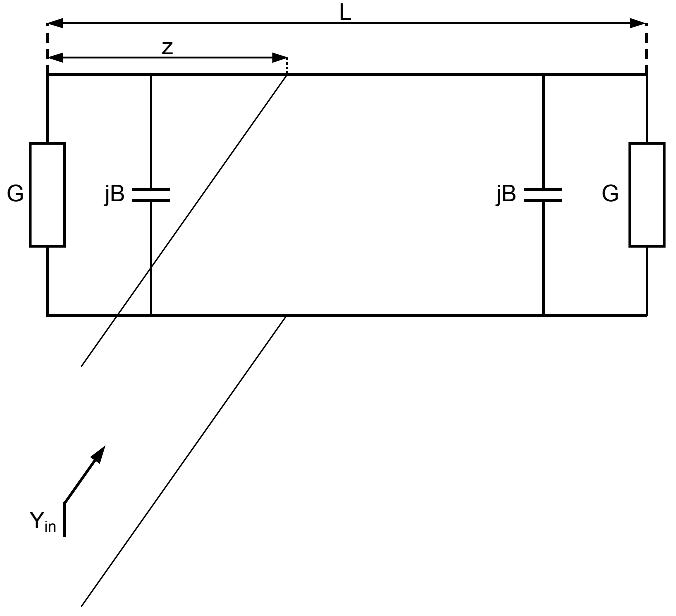


(a) Coordinate system.



(b) Half wavelength.

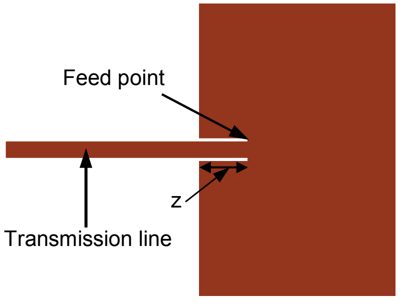
(c) Quarter wavelength.



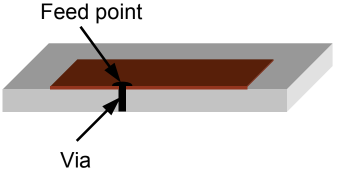
(d) Transmission line model for a half wavelength microstrip patch antenna.

Fig. 26. Illustration of the patch antenna: (a) coordinate system, (b) half wavelength patch antenna, (c) quarter wavelength patch antenna, and (d) transmission line model for a half wavelength microstrip patch antenna.

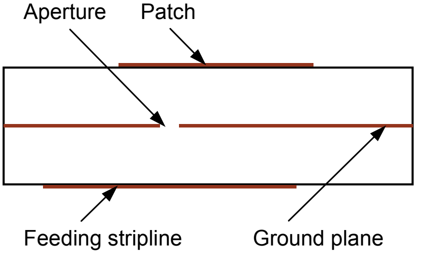
Matching of the patch antenna can be done in several ways. If the antenna is fed from the same layer as the patch itself matching can be made by the tap-in feeding technique as shown in Fig. 27a. That means cutting a slit such that the transmission line ends at the match point of the antenna. Utilizing a multi-layer PCB the feeding can be made either with a probe or an aperture, see Figs 27b and 27c. Probe feeding is done with a micro-coax cable or a via. Aperture coupled feeding is done with a stripline on a layer beneath the aperture slot [7], [47]-[51].



(a) Tap-in feeding.



(b) Probe feeding with a via.



(c) Aperture coupled feeding.

Fig. 27. Patch antenna feeding techniques: (a) tap-in feeding, (b) probe feeding with a via, and (c) aperture coupled feeding.

Fig. 28. shows a PCB prototype. The patch antenna has a CF of 6.5 GHz and is integrated in a PCB.

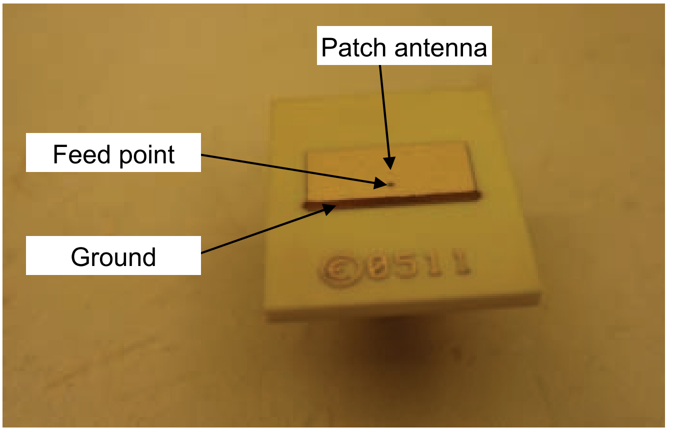


Fig. 28. A quarter wavelength patch antenna prototype.

3.2. Spiral Antenna

As shown in Fig. 29, the circumference of the radiation zone determines the radiation frequency. The circumference should be > $2\lambda_0$, where λ_0 is the wavelength. When the circumference surpasses this size the antenna will radiate a tilted beam [53], [54]. The tilted beam consists of one component for each wavelength the circumference reaches, i.e., for circumference 2-3λ_0 the beam consists of two components and so on. The relation between them determines the radiation performance in a certain direction at a specific frequency [50], [51]. If the current flowing backwards is smaller than that flowing forward, there is a net current flow through the spiral and radiation occurs. The net current-flow and the standing wave ratio (SWR) are related to each other, if the substrate loss is low [7], [53], [54].

Eq. 3.13 shows how the radius r dependants on the winding angle ϕ_{wnd} and the spiral constant a_s. In the equations C_{st} is the inner circumference, C_{end} is the outer circumference, I is current (real part I_r and imaginary part jI_i), and Z_{in} is the input impedance (real part R_{in} and imaginary part jX_{in}). The spiral constant a_s determines the turn distance Δs. Eq. 3.14 shows the definition of C_{st} and Eq. 3.15 shows the definition of C_{end}. Eqs. 3.16 and 3.17 show the requirements for a tilted beam. Eqs 3.18 and 3.19 show that the current flow and the input impedance are complex for an spiral antenna [7], [53], [54].

$$r = a_s \phi_{wnd} \tag{3.13}$$

$$C_{st} = 2\pi a_s \phi_{st} \tag{3.14}$$

$$C_{end} = 2\pi a_s \phi_{end} \tag{3.15}$$

$$C_{st} < \lambda_0 \tag{3.16}$$

$$C_{end} > 2\lambda_0 \tag{3.17}$$

$$I = I_r + jI_i \tag{3.18}$$

$$Z_{in} = R_{in} + jX_{in} \tag{3.19}$$

Feeding to the spiral can be done either from the center or from the outside. In the prototypes shown in this work the feeding is done by a via to the center of the spiral, see Fig. 29. The input impedance depends on the line width together with the distance to the ground plane, because the characteristic impedance of the spiral arm is dependant of the line width as in the case of a microstrip line. The real part of the input impedance can be controlled by the line width, while the imaginary part is more difficult to control. If implemented in a narrow band system the spiral antenna can be matched using a classical RF matching technique for optimal performance in that frequency region. However if a wideband operation is required the issue must be solved with other techniques [7], [53], [54].

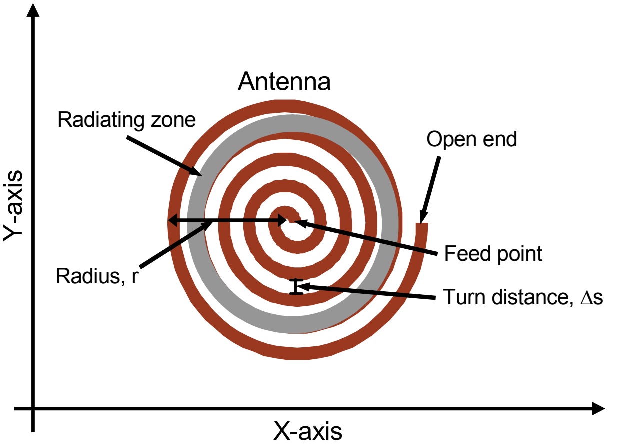


Fig. 29. Illustration of a monofilar spiral antenna.

Fig. 30 shows a photo of a monofilar spiral antenna prototype, where Δs is the turn distance and r is the radius of the antenna as shown in Fig. 29. The monofilar spiral antenna prototype has a radius of 50 mm and a turn distance of 6.0 mm.

Fig. 30. Photo of a monofilar spiral antenna prototype.

3.3. Inverted-F Antenna

The inverted-F antenna originates from the inverted- L antenna [7]. Fig. 31 shows the principle of inverted-F antenna, and the inverted-L antenna is shown as a reference. These antennas, which offer low profile and a fractional bandwidth enough for many narrow-band services, are widely used [7]. Fig. 31a shows the inverted-L antenna. The antenna consists of one horizontal, and a vertical section, l_A and l_B, respectively. The total (inverted-L antenna) radiator length ($l_{inv.-L}$) is given by the simple relation Eq. 3.20. The vertical section is similar to a short monopole antenna, which is capacitively loaded by the horizontal section. Because of the horizontal section loading effect the resonant frequency is lowered than that from a corresponding monopole antenna [7]. The total antenna length is shorter or much shorter than λ/4, see Eq. 3.20.

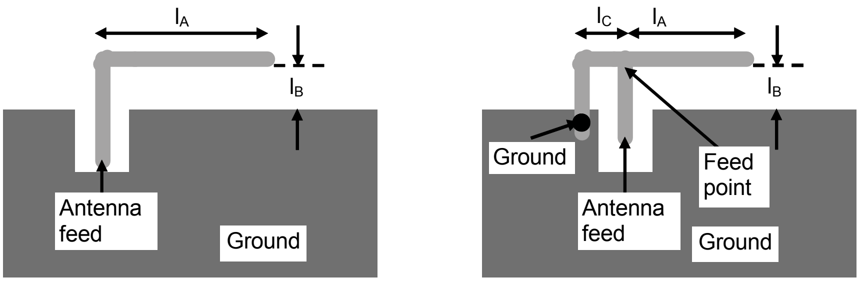
$$l_{\text{inv.-L}} = l_A + l_B << \lambda/4 \quad (3.20)$$

With this type of antenna configuration the vertical section produces an omnidirectional pattern, while the contribution from the horizontal section can be neglected due to image current cancellation. Image current cancellation occurs because the current in the horizontal section and the ground plane has opposite directions. The electromagnetic fields produced by the two currents then cancel out each other [7].

In the inverted-F antenna in Fig. 8b a horizontal section l_C is added, the section l_C is connected to the ground-plane with an additional vertical section l_B. Eq. 3.21 shows the inverted-F antenna length definition $l_{inv.-F}$. Eq. 3.22 shows a design rule regarding the feed point.

$$l_{\text{inv.-F}} = l_A + l_B + l_C \approx \lambda/4 \quad (3.21)$$

$$l_C << \lambda \quad (3.22)$$



(a) inverted-L antenna. (b) inverted-F antenna.

Fig. 31. Inverted antennas: (a) inverted-L antenna, and (b) inverted-F antenna.

When integrated on a printed circuit board via is used to connect the antenna to the ground-plane. Feeding is done using a microstrip-line of 50-Ω characteristic-impedance. The length of the antenna is one quarter-wavelength at the center-frequency [55], [56]. Since the open-end has high impedance and the short-end has zero impedance, a desired feed point impedance (typically 50 Ω) can therefore be found somewhere on the line [55]-[57].

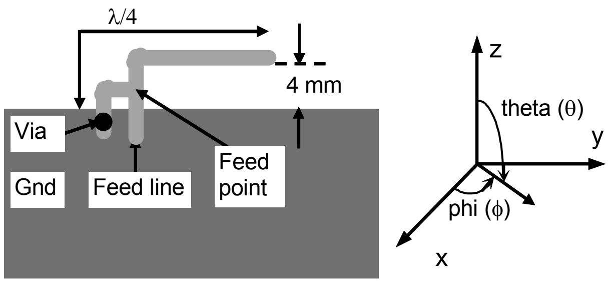


Fig. 32. A typical modified inverted-F antenna integrated on a printed circuit board.

3.4. Monopole and Dipole Antennas

Circular disc and other wideband implementations of monopole and dipole antennas have been studied using the flex-rigid concept in Paper VIII. The theory of the straight wire finite length monopole and dipole antennas is the foundation of all wideband monopole and dipole implementations. Fig. 33 shows a straight wire dipole antenna, consisting of two symmetrical metal conductors. The antenna arms have a finite length and are ended with an open-end. The current reaches zero at the open end so the current distribution along the antenna arms is not constant. However, if the antenna is center-fed the phase distribution along its length will be constant. Furthermore, due to the physical length of the antenna, radiation contribution from different segments of the antenna will reach an observer with different phase delays. The phase delay variation causes the radiated wave-fronts from the antenna-segments to be added constructively or destructively, at each observation point [7]. The dipole straight wire dipole antenna is linearly polarized with the electrical field vector in the wire axis. The radiation resistance of an ideal half-wave dipole antenna is 60-73 Ω, depending of the wire length-width ratio [7].

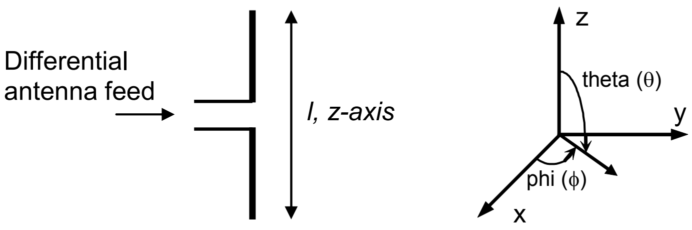


Fig. 33. Straight wire dipole antenna.

The antenna with a length l is excited with a sinusoidal current distribution. Eq. 3.23 shows the antenna pattern factor $F(\theta)$. The pattern is dependant of the relation between the physical length l and the wave number k. As seen in Eq. 3.24 the wave number is directly related of the wavelength. Therefore the longer the antenna is compared to the wavelength, the more complex radiation pattern the antenna gets [7].

$$F(\theta) = \frac{\cos\left(\frac{1}{2}kl\cos(\theta)\right) - \cos\left(\frac{1}{2}kl\right)}{\sin\theta} \qquad 3.23$$

$$k = \frac{2\pi}{\lambda} \qquad 3.24$$

Half-wave dipole, l=λ/2:

$$F(\theta) = \frac{\cos\left(\frac{\pi}{2}\cos(\theta)\right)}{\sin\theta} \qquad 3.25$$

Full-wave dipole, l=λ:

$$F(\theta) = \frac{\cos(\pi\cos(\theta) - 1)}{\sin\theta} \qquad 3.26$$

l equal to one and a half wavelength, l=3λ/2:

$$F(\theta) = \frac{\cos\left(\frac{3\pi}{2}\cos(\theta)\right)}{\sin\theta} \qquad 3.27$$

The electrical length of a straight wire dipole and monopole depends on the properties of the medium within the reactive near field of the radiating antennas. When integrated on a printed circuit board the wavelength then depends on the dialectical materials permittivity and the surrounding air. Moreover when placed on a printed circuit board with a permittivity > 1 the phase velocity is lowered as discussed in Section 2.2.1. As shown in Eq. 3.28 the radiating antenna then experiences an effective permittivity that depends on the mentioned surrounding circumstances. Furthermore, the effective permittivity does not only depend on the material properties but also on the length, and width of the wire. Eq. 3.29 shows that the phase velocity now depends on the effective permittivity, and Eq. 3.30 shows that the relation between frequency and wavelength now also depends on the effective permittivity.

$$e_{eff} = \frac{e_r + 1}{2} + \frac{e_r - 1}{2}\left(\frac{1}{\sqrt{1+\frac{12h}{w}}} + 0.04\left(1-\frac{w}{h}\right)^2\right) \qquad (3.28)$$

$$v = \frac{c}{\sqrt{e_{eff}}} \qquad (3.29)$$

$$f = \frac{c}{\lambda\sqrt{e_{eff}}} \qquad (3.30)$$

For the corresponding straight wire monopole antenna shown in Fig. 34 the radiation pattern is halved, i.e., applying the limit $-\pi/2<\theta<\pi/2$ to Eqs. 3.23-3.27. The monopole radiation-pattern above the ground plane is preserved due to the effect of mirroring of the ground plane. Moreover, since the radiation is concentrated in one half-sphere the gain is under ideal circumstances increased with 3 dB. These statements are only valid with an ideal infinitively large ground plane [7]. The radiation resistance of an ideal quarter-wave monopole antenna is 30-36.5 Ω, depending on the wire dimension [7].

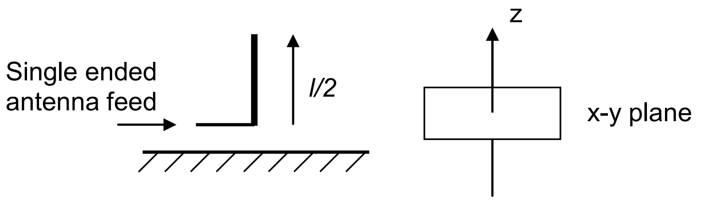
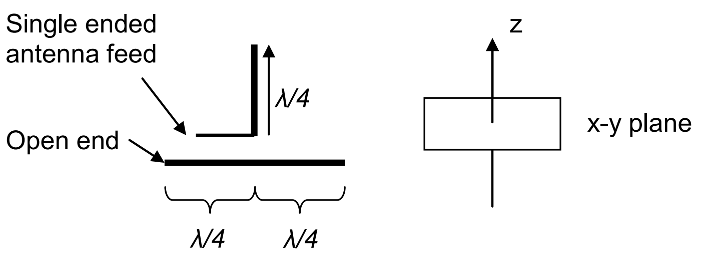
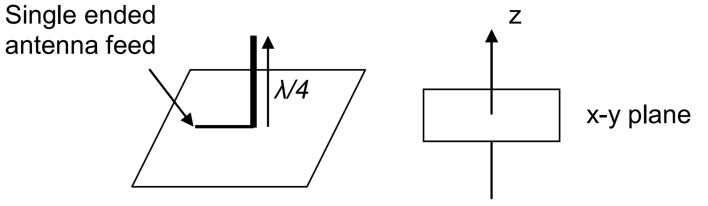


Fig. 34. Monopole antenna with infinite ground plane.

Naturally an infinitely large ground plane is not possible in a real antenna implementation. Fig. 35a shows a monopole antenna with a λ/4 counterpoise as the ground plane. The monopole antenna with several λ/4 open stubs is also known as the Marconi antenna after the inventor. The principle is to maintain a symmetrical antenna ground. As long as the ground is symmetrical and the radiating antenna element is perpendicular to the ground the vertical polarization is maintained. Moreover, the antenna end should have a long distance to the ground in order to maximize the radiation resistance and efficiency. Fig. 35b shows a generic monopole antenna with a finite ground plane. The polarization remains linear as long as the finite ground plane is symmetrical around the antenna. If the ground plane is unsymmetrical then the polarization is still linear but is tilted towards the largest part of the ground plane [7].



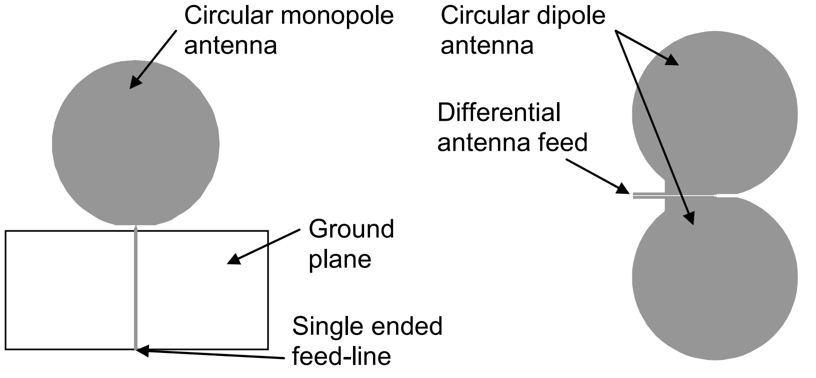
(a) Monopole antenna with a λ/4 counterpoise.



(b) Monopole antenna with a non-ideal ground-plane.

Fig. 35. Monopole antennas with non-ideal ground planes: (a) with λ/4 counterpoise, and (b) with general non-ideal ground-plane.

The circular dipole antenna is a wideband implementation of a finite length dipole antenna. The antenna is a geometrical antenna. That means that the antenna not only relies on the arm length but also on the surface revolution [2], [59]. Fig. 36a shows a circular monopole antenna with a finite ground plane. Fig. 36b shows a circular dipole antenna. The circular structure provides an omnidirectional radiation pattern [2].



(a) Circular monopole antenna. (b) Circular dipole antenna.

Fig. 36. Circular antennas: (a) monopole antenna, and (b) dipole antenna.

4. Ultra-wideband Radio

The UWB technology promises a high performance in order to meet demanding high data rate applications. At the moment short-range and high-speed data communications are one of the areas that draw the most attention but other applications as vehicular radar, imaging systems, and sensor networks are also possible applications. The worldwide interest for UWB is seen not only in the amount of research work, and scientific publications that have been produced in recent years, but also on the regulatory side. A survey of today's UWB regulations around the world is displayed in the end of this chapter.

4.1. Overview

Ever since the federal communications commission (FCC) in USA approved the frequency band 3.1-10.6 GHz for indoor communications, the interest for wideband and wireless communications has virtually exploded. The combination of its high data rate capacity and low power consumption provides a solid platform for the future market-share growth. Without loosing the advantage of wireless operability the power consumption should be kept to the minimum. UWB radio communications are an interesting option to reach these new combined demands of high capacity and low power consumption. Equivalent isotropically-radiated power (EIRP) for indoor UWB communications was regulated to –41.3 dBm/MHz by the US government back in 2002, which makes for example the receiver sensitivity an important issue. The emission level is measured in a 1-MHz bandwidth [4]. Currently the development of physical layers, stacks, chips, devices, and applications draws a lot of research attentions because of the enormous potential of UWB communication systems. The adjustments of regulations of other frequency bands than the 3.1-10.6 GHz band is constantly being discussed for various UWB applications. The 3.1-10.6 GHz

UWB band has been released for indoor and handheld communications and high frequency imaging systems. Fig. 37 shows the FCC spectral mask for indoor UWB communications. There exist regulations for other frequency-ranges as well, but currently the possibility of wireless personal area network (WPAN) services within 3.1-10.6 GHz draws the most attention. For WPAN services there are currently two dominating technologies, one based on the orthogonal frequency division multiplexing (OFDM) technique, and the other based on the direct sequence spread spectrum (DSSS) technique [4], [42]-[46].

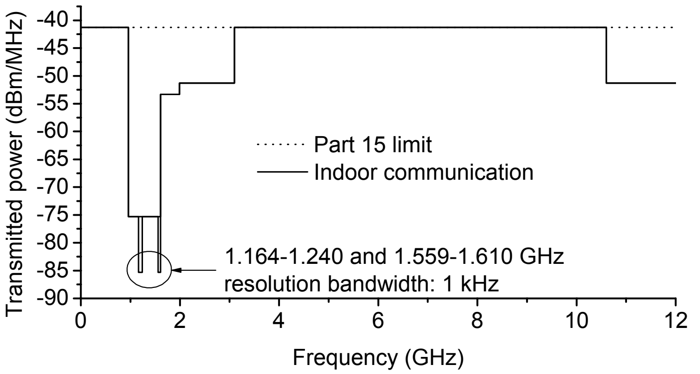


Fig. 37. FCC spectral mask for indoor communications.

4.2. History

Pulse radio for communication systems that occupy a large frequency bandwidth is not a new invention. Already back at a time when Guliano Marconi researched in radio communications, large frequency bandwidths were used. The first spark-gap transmitter he built utilized a huge amount of bandwidth because, without modern filtering possibilities [4], the transmitted power was spread throughout all possible frequencies. The first work in the time-domain with short impulses was done in the early 1960s. The first purposed attempt to make use of a wide spectrum came in the late 1960s when the first UWB radar system was introduced. The motivation to use a UWB technology in radar systems was its high sensitivity to scatters and its low power consumption [4], [5].

The take-off of today's UWB development came in 1989 when the Department of Defense (DoD) in USA specified the term "ultra-wideband" for devices occupying a bandwidth of at least 1.5 GHz, or a fractional bandwidth exceeding 25% at –20 dB. Later on, FCC adopted similar definitions and started the rule-making of UWB in 1998. The definition was later adjusted to at least 500 MHz bandwidth or a fractional bandwidth exceeding 20% at –10 dB, i.e., the 500 MHz rule is applied at the 2.5 GHz CF and above [1]. Such large bandwidths are far more than what is conventionally used even in modern communication systems. Fig. 38 illustrates typical bandwidths used by conventional, spread-spectrum, and UWB communications, respectively. The first version of the resulting report and order (R&O) regarding UWB devices was released in April 2002. The R&O defined three categories of UWB applications, imaging systems, communication and measurement systems, and vehicular radar systems [2], [4], [5].

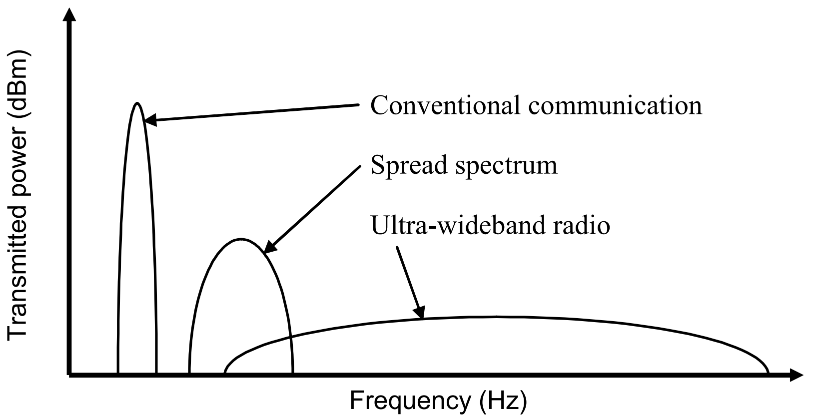


Fig. 38. Bandwidth utilizations.

USA was first to allow commercial UWB communications, but efforts are being made to develop complete regulations in Europe and some countries in Asia as well. International as well as regional interests are currently struggling towards an industry-wide standardization of UWB radios for indoor communications, covering many things from toys to multimedia data transmissions.[4], [45].

4.3. Theory

Eqs. 4.1-4.3 show Shannon's channel capacity theorem. It is the channel capacity with respect to additive, white Gaussian noise. Were S is the signal power, and N is the noise power in watts, respectively. SNR is the signal to noise ratio, i.e., signal power divided by noise power. B is the channel bandwidth in Hz, and C is the channel capacity in bit/s. It is seen that C grows linearly with B, but only logarithmically with SNR. With this conclusion it is easy to argue that a wide channel bandwidth is the main route to a high channel capacity, i.e., to reach a high data rate.

$$SNR = \frac{S}{N} \tag{4.1}$$

$$C = B\log_2\left(1 + \frac{S}{N}\right) \tag{4.2}$$

$$C = B\log_2(1 + SNR) \tag{4.3}$$

4.4. Technology and Applications

UWB is based on a technology known as spread spectrum. In general spread spectrum means that the information transferred is spread over a frequency spectrum, i.e., pieces of information is transferred at different frequencies according to a designed scheme. In its original concept, spread spectrum conveyed data using a series of short radio pulses that were transmitted at different frequencies within a specific range of spectrum. This technique was already used in World War II by the US military to ensure that its communications avoid being monitored by enemies. Since the signals at any specific frequency were incomplete, the enemy was not able to intercept the entire message. Spread-spectrum was widely used by the military for years in secure communications. In 1985, FCC opened up three bands for spread-spectrum use in commercial communication devices, establishing a 1 watt power limit as a way of preventing interference, soon the technology was used in a variety of communication devices. For example, cordless telephones and wireless local-area-network products using spread-spectrum are now common. UWB adds several layers of sophistication to the spread-spectrum concept. The UWB-pulses amplitude, polarity, timing and other characteristics forming a data stream can be carefully controlled using packets and error-correction methods. A UWB data transmission may involve billions of pulses spread over several gigahertz, enabling high performance.

Many categorizations can be made when it comes to UWB application possibilities. One common denominator is the scalability and large variety in potential usage. Roughly it can be seen as two areas, one covering wireless communications, and the other covering all radar applications. In more detail the communication area can be divided into WPAN services and sensor networking. WPAN is typical for short range high speed communications and sensor networking can use the UWB physical layer together with the MAC layer.

4.4.1. Wireless Personal Area Network (WPAN)

UWB WPAN covers high speed data communications within a short range 10-20 m. These include home networks, ad-hoc connections between hand-held devices and other consumer electronics. The vision is to provide demanding services as real-time high-quality streaming video and general cable replacements, no matter if it is audio, video or files to be exchanged. With the maximum EIRP of -41.3 dBm/MHz communications can be done simultaneously with other existing WPAN technologies. Outside the 3.1-10.6 GHz band EIRP is restricted even further to assure complete coexistence with other existing standards and technologies [4], [42], [58], [60]. There are currently two dominating technologies, one based on the Multi-band OFDM technique and the other based on the DS-UWB technique [4], [45].

4.4.2. Imaging Systems

A radar system has the possibility of becoming an imaging system when the pulses are much shorter than the target, because information about the target can be extracted in two or three dimensions. Intelligent data computation is then utilized on the dataset to build an image. The aim of low resolution radar is to detect stand or moving objects while high resolution radar and imaging systems can detect movement within or through the objects. Military, search and rescue applications are typical usages of imaging systems. Owing to the controversy concerning personal integrity this field will not be entirely open for public usage. However, so far FCC has issued the following usage areas. Medical imaging, similar to X-ray and computed axial tomography (CAT) scans. Other applications include through-wall imaging for detecting people or objects in law-enforcement or rescue applications, and construction applications, for through-wall imaging and ground-penetration. Reflected UWB pulses consist not only of time shift and amplitude variation but also of changes in the signal shape. The level of details depends on the pulse-length, which allows a large scalability for various applications from indoor object imaging, through wall imaging, and underground vision to medical diagnostics. Finding objects as power lines and water pipes within walls, floors, and roofs is one of civil applications. Otherwise the military, police, and other authorities are large potential users. The recent rising of terror threat increases the need of surveillance, along with other usage of surveillance systems [4], [44].

4.4.3. Sensor Networks

Sensor networks are built up by a number of nodes spread over a certain area. The purpose is to gather data from the area, either directly by a point-to-point network or by several multi-hop nodes with the ad-hoc technique. Traditionally a sensor network gathers data in a small data quantity but UWB has the possibility to gather a large amount of data. A physical layer with low-power consumption, suitable for sensor networks, and run by a battery, is being developed [61]. For ultra low-power and low data-rate Zigbee and other low data-rate standards are probably the best choice [62], but a UWB sensor network can transmit a large amount data at a data rate up to 480 Mbps [4], [44].

So far UWB has become known as one of possible next generation solutions for high-speed and short-range wireless communications. The IEEE 802.15.4x Task Group works on low rate alternative of physical layers for wireless personal area networks (LR-WPAN). The IEEE 802.15.4a Task Group is mostly known for standardization of the ultra low-power technologies like Zigbee [62]. The group has also contributed to make use of UWB in ultra low-power systems, by utilizing lower data-rate than conventional WPAN UWB-systems. UWB impulse radio operating in the unlicensed UWB spectrum and a chirp spread spectrum operating in the unlicensed 2.4 GHz spectrum [44], [62].

4.4.4. Vehicular Radar Systems

The short UWB pulses have the possibility to enhance the resolution of conventional proximity and motion sensors [44], [63]-[65]. Electronics has made its way into vehicles for some time and collision avoidance is so far the dominating radar implementation. Besides collision avoidance it can also be used for cruise control, improving airbag deployment and adapting vehicle settings depending on the shape of the road [44]. Moreover the usage of UWB for radar purposes does not imply limitations to use it for transferring data to and from the vehicle, e.g., to download information when arriving at a supermarket, gas station or any other place equipped with UWB transmitters [44]. Recently the UWB radars operating within the 22-29 GHz frequency band have received a lot of research attention [64], [65].

4.5. High Speed Short Range Communication using UWB

4.5.1. Multi-Band Orthogonal Frequency Division Multiplexing

Specifications given are valid for the multi-band OFDM physical layer currently supported by Wimedia alliance [45]. The multi-band OFDM UWB system provides a WPAN with data payload communication capabilities of 53.3, 80, 110, 160, 200, 320, 400, and 480 Mbps, where the support for transmitting and receiving at data rates of 53.3, 110, and 200 Mbps are mandatory [4], [45], [58].

An OFDM channel transfers with a total of 122 subcarriers for modulated and pilot signals. Convolutional coding is used for forward error correction with a coding rate of 1/3, 11/32, 1/2, 5/8, and 3/4, respectively. Coded data is interleaved between three bands with time-frequency coding (TFC). Three bands together form one band group. There is room for four full band groups within the 3.1-10.6 GHz band. The band group 3.1-4.8 GHz is the originally specified band-group. This band group is referred to as the mode 1 band-group. Four three-band groups and one two-band group are specified, providing a total of 14 frequency-bands. The defined TFC division provides up to eighteen logical channels or independent piconets. Fig. 39 shows the mode 1 band group in which three sub-bands are centered at 3.432, 3.960, and 4.488 GHz, respectively [4]. The channel distance is 528 MHz between each CFs. The rest of the unlicensed UWB spectrum should be divided in the same way [4], [45], [58].

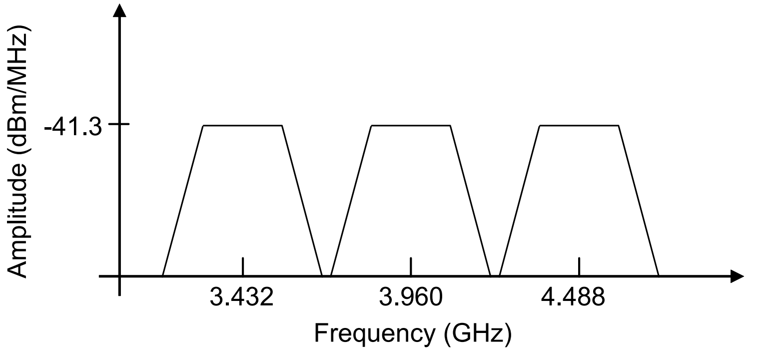


Fig. 39. Mode 1 base-band group.

Fig. 40 shows the detailed transmitter power spectral density (PSD) mask within one band [45]. The values shown are normalized relative to the maximum output power, i.e., 0 dB corresponds to –41.3 dBm/MHz. The frequency is normalized to the CF. This mask is applied for all frequency bands. It is mandatory to stay within the mask for all pulses [4], [45], [58].

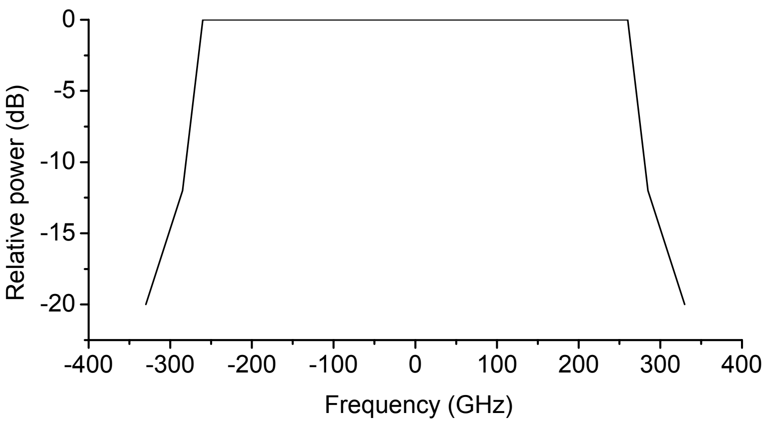


Fig. 40. Transmit power spectral density mask.

4.5.2. Direct Sequence Spread Spectrum

Specifications given are valid for the UWB systems that use the DSSS technology. The proposed DS-UWB solution provides a WPAN with data payload communication capabilities of 28, 55, 110, 220, 500, 660, 1000 and 1320 Mbps [4], [45], [60].

Fig. 41 shows the DS-UWB reference mask [45]. The proposed UWB system employs direct sequence spreading of binary phase shift keying (BPSK) and quaternary bi-orthogonal keying (4BOK) UWB pulses. Convolutional coding is used as forward error correction with a coding rate of 1/2 or 3/4. The proposed UWB system also supports operation in two different bands. The lower frequency band occupies the spectrum from 3.1 to 4.8 GHz. The second higher frequency band occupies the spectrum from 6.2 to 9.7 GHz. The suggested DS-UWB supports use of both BPSK and 4-BOK modulations, while a complete support for BPSK is mandatory, and a 4-BOK support is intended as an optional performance improvement. However, the transmit capability of 4-BOK is still mandatory. The reason is that implementing 4-BOK on the transmitter-side does not noticeably increase the transmitter complexity. Within each band there is support for up to six piconet channels to have unique operating frequencies and acquisition codes. Support for piconet-channels 1-4 which are in the low band is mandatory. Support for piconet-channels 5-12 are optional. Channels 5 and 6 are optional in the low band, and channels 7-12 are optional channels in the high band. Various data rates are supported with variable spreading sequence code lengths. The sequence length varies from 1 to 24 pulses [45], [60].

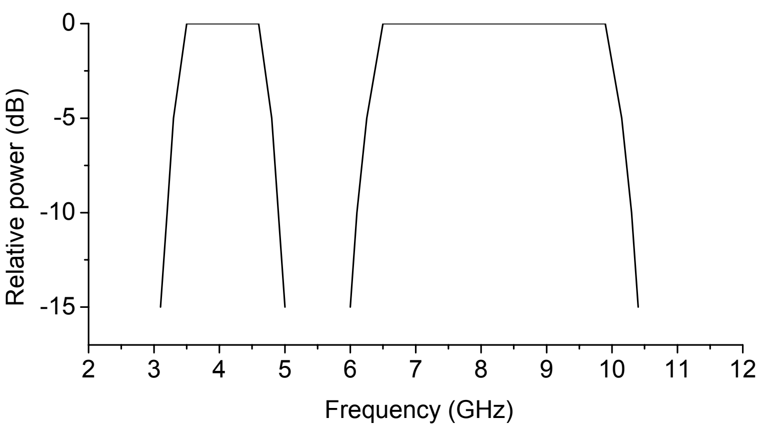


Fig. 41. DS-UWB reference masks.

4.6. Regulation

In order to have an as wide market as possible, business interests consistently try to push the regulatory instances to allow UWB. Everything started with FCC in the U. S. and worldwide adoptions are being made, often with some local limitations or extra requirements. However, the regulation work is going on around the world. For instance, in Europe ETSI is the organization setting up the legal framework for radio communications. Europe regulation efforts are a bit more complicated since there are many different wills from the member states in the European Union. All countries are separate states that have to adopt the proposed regulations within their territory. In Asia the UWB regulation, and development are progressing fast in several countries. For instance, the technology progressive Japan has launched frequency bands to be used for UWB services at the frequency bands 3.4-4.8 GHz and 7.25-10.25 GHz [66]-[79].

4.6.1. UWB in the US

When it comes to UWB in USA the most important event was when FCC unleashed the 3.1-10.6 GHz frequency-band in 2002. The frequency band was unleashed for the purpose of unlicensed reuse of licensed spectrum, i.e., legalizing unlicensed use of the 3.1-10.6 GHz frequency band at a maximum average transmit power of -41.3 dBm/MHz. A few years later, more detailed specifications regarding transmit power for various application-areas have been made. The most important areas and their frequency restrictions, and specifications are listed in the following sub-sections. The specifications given in the sub-sections below are not the complete FCC specifications for UWB transmitters. What are described are the FCC regulations for frequency spectrum usage, i.e., emission levels for the respective services [66]-[68].

- **Indoor communication:** Indoor communication is for high speed short range data transfers made with UWB transmitters. Fig. 42 shows the mean value emission mask for indoor communication. The resolution bandwidth is 1 MHz. Moreover, within the frequency ranges 1.164-1.240 GHz and 1.559-1.610 GHz the emission must also comply with a resolution bandwidth of 1 kHz. The maximum peak level of the emissions is limited to 0 dBm EIRP and average EIRP -41.3 dBm/MHz. The FCC definition for peak-level of the emission is the emission contained within a 50 MHz bandwidth centered on the frequency at which the highest radiated emission occurs [66]-[68].

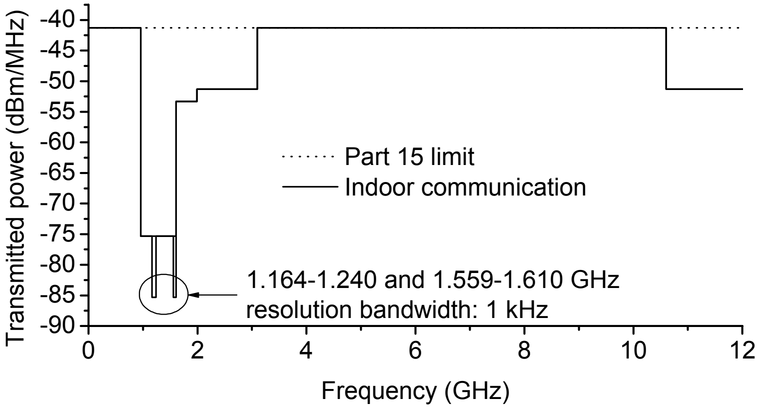


Fig. 42. FCC emission mask, indoor communication.

- **Outdoor, handheld communications:** High speed short range data communications for outdoors, typically handheld devices. Fig. 43 shows the mean value emission mask for outdoor communication. The resolution bandwidth is 1 MHz. Moreover, within the frequency ranges 1.164-1.240 GHz and 1.559-1.610 GHz the emission must also comply with a resolution bandwidth of 1 kHz. Similar restriction as indoor communication. However, the out-of-band rejection requirement between 1.61-3.10 GHz and above 10.60 GHz is 10 dB higher for out-door communication. The maximum peak level of the emissions is limited to 0 dBm EIRP and average EIRP -41.3 dBm/MHz [66]-[68].

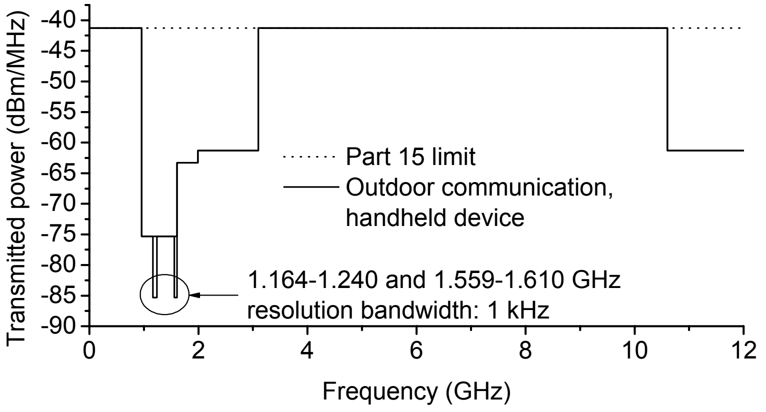


Fig. 43. FCC emission mask, outdoor, handheld communication.

- **Vehicular radar:** Used for collision avoidance and road safety applications. Fig. 44 shows the mean value emission mask for vehicular radar. The resolution bandwidth is 1 MHz. Moreover, within the frequency ranges 1.164-1.240 GHz and 1.559-1.610 GHz the emission must also comply with a resolution bandwidth of 1 kHz. The maximum peak level of the emissions is limited to 0 dBm EIRP and average EIRP -41.3 dBm/MHz [66]-[68].

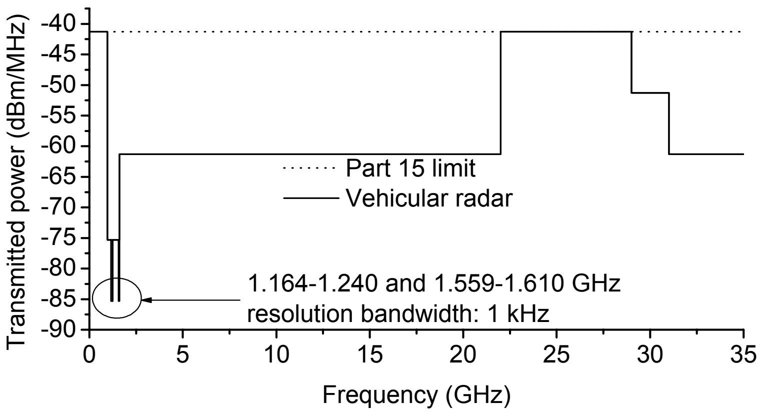


Fig. 44. FCC emission mask, vehicular radar.

- **Ground penetrating radar and wall imaging systems:** Fig. 45 shows the mean value emission mask for ground penetrating radar and wall imaging systems. The resolution bandwidth is 1 MHz. Moreover, within the frequency ranges 1.164-1.240 GHz and 1.559-1.610 GHz the emission must also comply with a resolution bandwidth of 1 kHz. The maximum peak level of the emissions is limited to 0 dBm EIRP and average EIRP -41.3 dBm/MHz [66]-[68].

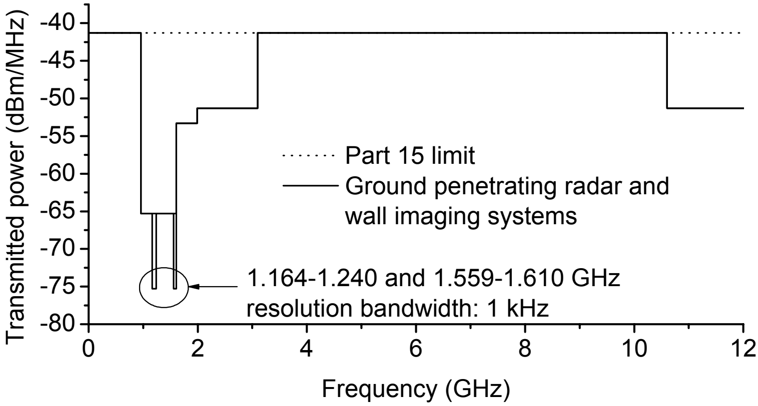


Fig. 45. FCC emission mask, ground penetrating radar, and wall imaging systems.

- **Through wall imaging systems:** Fig. 46 shows the mean value emission mask for through wall imaging systems. Operation under the provisions of this section is limited to through-wall imaging systems operated by law enforcement, emergency rescue or firefighting organizations that are under the authority of a local or state government. Moreover, within the frequency ranges 1.164-1.240 GHz and 1.559-1.610 GHz the emission must also comply with a resolution bandwidth of 1 kHz. The maximum peak level of the emissions is limited to 0 dBm EIRP and average EIRP -51.3 dBm/MHz [66]-[68].

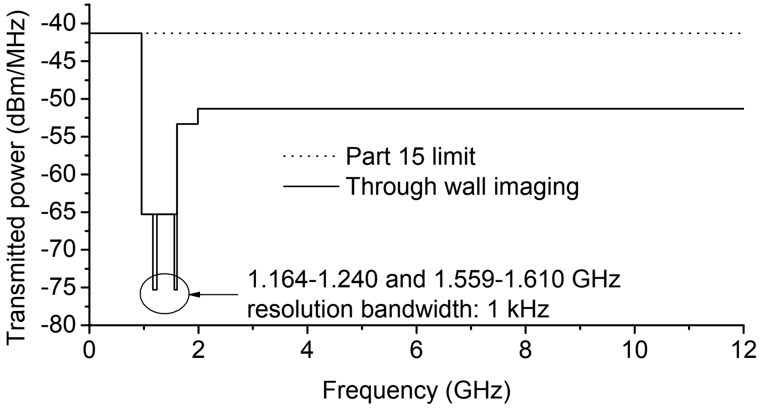


Fig. 46. FCC emission mask, through wall imaging systems.

- **Medical imaging systems:** Fig. 47 shows the mean value emission mask for medical imaging systems. Medical imaging, similar to X-ray and CAT scans. The resolution bandwidth is 1 MHz. Moreover, within the frequency ranges 1.164-1.240 GHz and 1.559-1.610 GHz the emission must also comply with a resolution bandwidth of 1 kHz. The maximum peak level of the emissions is limited to 0 dBm EIRP and average EIRP -41.3 dBm/MHz [66]-[68].

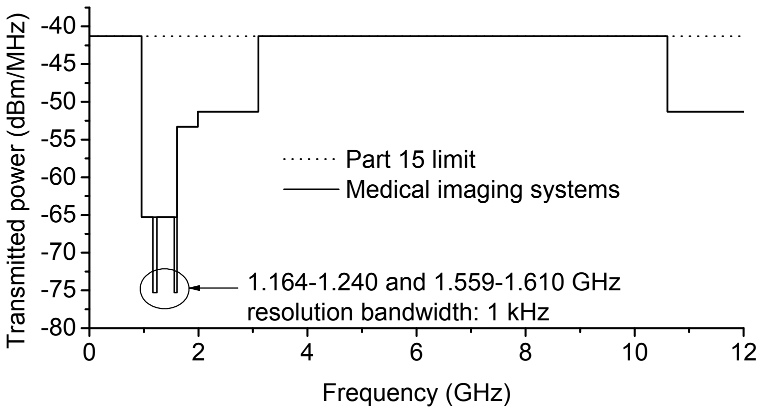


Fig. 47. FCC emission mask, medical imaging.

- **Surveillance systems:** Fig. 48 shows the mean value emission mask for surveillance systems. Systems for surveillance and guard purposes. This is for fixed surveillance systems operated by law enforcement, fire or emergency rescue organizations or by manufacturers licenses. Moreover, within the frequency ranges 1.164-1.240 GHz and 1.559-1.610 GHz the emission must also comply with a resolution bandwidth of 1 kHz. The maximum peak level of the emissions is limited to 0 dBm EIRP and average -41.3 dBm/MHz [68].

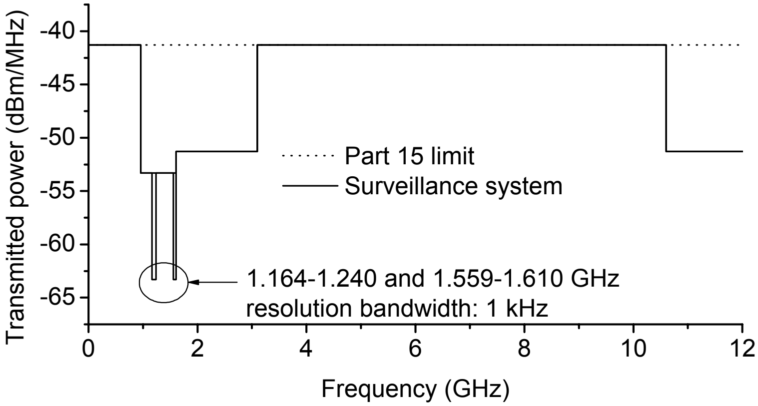


Fig. 48. FCC emission mask, surveillance systems.

- **Law enforcement:** Fig. 49 shows the mean value emission mask for law enforcement applications. The UWB equipment using this specification requires a special operator's license. The intended users are: "Operation of this device is restricted to law enforcement, emergency rescue and firefighter personnel. Operation by any other party is a violation of 47 U.S.C. 301 and could subject the operator to serious legal penalties." [68] Moreover, within the frequency ranges 1.164-1.240 GHz and 1.559-1.610 GHz the emission must also comply with a resolution bandwidth of 1 kHz [68]. Maximum average EIRP is -41.3 dBm/MHz [68].

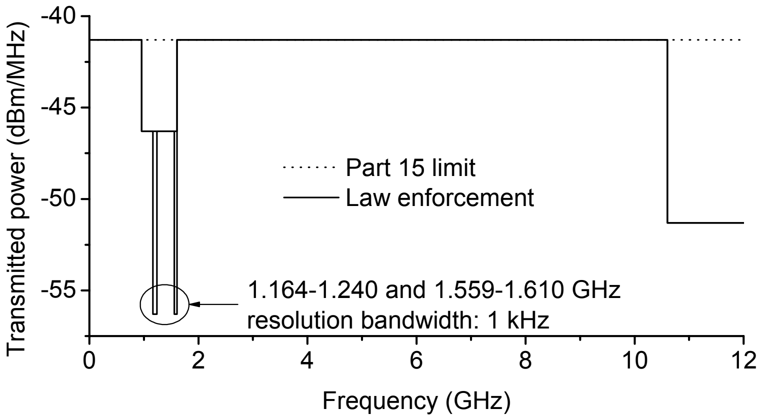
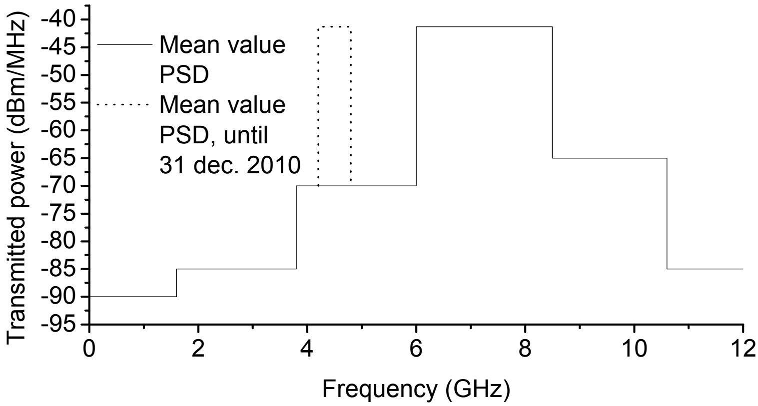


Fig. 49. FCC emission mask, law enforcement.

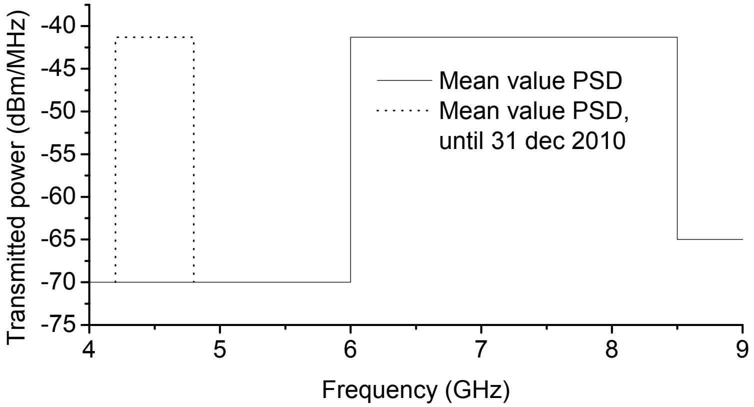
4.6.2. European UWB

In Europe there are a number of key players officially acknowledged by the European Commission (EC). The European Radiocommunications Office (ERO) was formally opened 1991 and is located in Copenhagen, Denmark. ERO is the permanent office supporting the Electronic Communications Committee (ECC) of the European Conference of Postal and Telecommunications Administrations (CEPT). CEPT is a cooperation organisation between the European post and telecom regulatory bodies. It was founded in 1959 as a cooperation forum between European postal and telecom administrations, but has, since 1992, been working exclusively with regulatory issues. ECC is the Committee that brings together the radio- and telecommunications regulatory authorities of the 48 CEPT member countries, i.e., with a total of 48 members the CEPT reaches outside the EU. The ECC publishes regulatory decisions, recommendations within the area. Another key player to the ECC within EU is the European Telecommunications Standards Institute (ETSI) that is officially responsible for standardization of Information and Communication Technologies (ICT) within Europe. ETSI has published to specification recommendation documents regarding UWB. However, EC gave mandate to CEPT, and ECC to harmonise UWB regulations in Europe. EC also publishes regulation documents, mostly to reach out to the EU member states in a more regulatory manor in order to force them in the same direction. Currently ETSI, ECC, and EC have all recommended and approved the use of UWB devices in the range 6.0-8.5 GHz, subject to the technical limits enforced by the FCC in the US without requiring the use of interference mitigation techniques. Furthermore, UWB devices placed on the market before 31st December 2010 are permitted by EC to operate in EU in the frequency band 4.2 - 4.8 GHz with the maximum mean EIRP spectral density of –41.3 dBm/MHz and the maximum peak EIRP of 0 dBm measured in 50 MHz. EC also stipulates the use of the maximum mean EIRP density of –41.3 dBm/MHz in the 3.4 - 4.8 GHz band provided that a low duty cycle restriction is applied in which the sum of all transmitted signals is less than 5 % of the time each second and less than 0.5 % of the time each hour, and provided that each transmitted signal does not exceed 5 milliseconds [69]-[74].

Fig. 50 shows the European UWB mean radiation emission mask. Fig. 48a shows the complete frequency band, and Fig. 48b shows in more detail frequency band regulations. The European UWB specifications are not yet as well specified as the FCC UWB specifications when it comes to applications. However, UWB transmitters are so far prohibited to be used aboard an aircraft or a ship, and at a fixed outdoor location. Furthermore, the UWB devices should comply with all regulations at temperatures between 15-35 ºC, and humidity between 20-75 %. Some technical restrictions regarding the measurement procedures are also stipulated. For instance, the resolution bandwidth should be 1 MHz and the video bandwidth should not be less than the resolution bandwidth. The detector mode should be root mean square (rms). Average time per point on spectrum analyzer scan should be 1 ms or less [69]-[74].



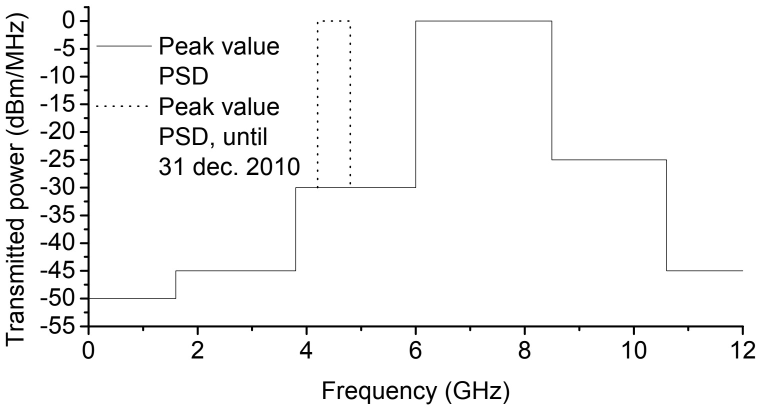
(a) European mean value emission mask, complete frequency band.



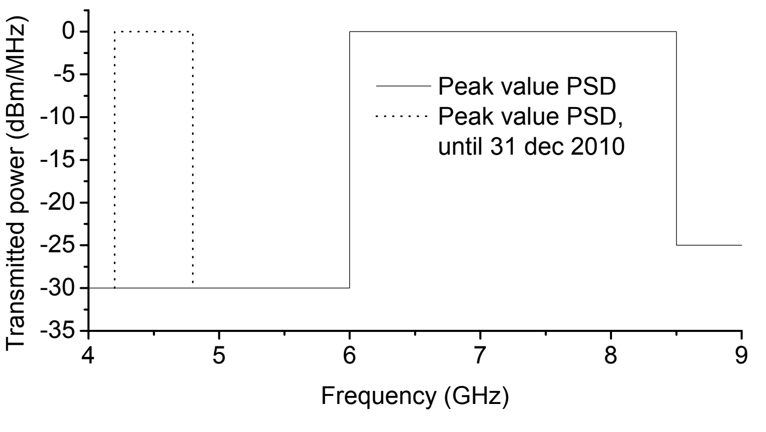
(b) European mean value emission mask, main frequency band only.

Fig. 50. European mean value radiation emission spectrum masks: (a) complete frequency band, and (b) main frequency-band only.

Besides the mean value restrictions there are emission limits regarding the maximum peak level versus frequency. Fig. 51 shows the peak value of the maximum emission mask. Besides the mean value specifications there are technical conditions for the peak value specifications. The resolution bandwidth should be at least 1 MHz and not greater than 50 MHz. The video bandwidth should not be lower than the resolution bandwidth. The signal detection should be made with peak detection using the maximum-hold, and the measurement sequence should continue until that the maximum-hold trace is stable, i.e., until the maximum value has been captured at all measured frequency points [69]-[74].



(a) European peak emission mask, complete frequency spectrum.



(b) European peak emission mask, main frequency spectrums only.

Fig. 51. European peak value radiation emission spectrum masks: (a) complete frequency spectrum, and (b) main frequency-spectrum only.

UWB devices placed on the market before 31st December 2010 are permitted to operate in the frequency band 4.2 - 4.8 GHz with the maximum mean EIRP spectral density of –41.3 dBm/MHz and the maximum peak EIRP of 0 dBm measured in 50 MHz [69]-[74].

4.6.3. UWB in Japan

The national institute of information and communications technology (NICT) in Japan permits UWB radio transmission equipment of radio stations using ultra wideband radio systems at two frequency bands. The allocated frequency bands are from 3.4 to 4.8 GHz and from 7.25 to 10.25 GHz. For the lower frequency band 3.4-4.8 GHz detect and avoid mitigation techniques to avoid interference with existing narrow band services are required. For devices without any interference mitigation techniques, the average power shall instead be -70 dBm/MHz and the peak power shall be -64 dBm/MHz. However, UWB devices are allowed to be used without any interference mitigation techniques in the 4.2-4.8 GHz band until the end of year 2008 [reg14]. Japan was first to suggest mitigation techniques as a solution to get acceptance for UWB transmissions [79].

Fig. 52 shows the mean value Japan UWB mean radiation emission mask. The average transmission power is limited to -41.3 dBm/MHz or lower on both bands. The peak value is set to 0dBm/50MHz or lower. For the devices without any interference avoidance technology that uses the 3.4-4.8 GHz band the average transmission power is limited even further to -70 dBm/MHz or lower. Furthermore, in this case the peak value is restricted to -30 dBm/50MHz or lower. Communication and modulation methods are not specified. Transmission speed of 50 Mbps or higher is required for the radio equipment, although lower speeds are permitted in order to avoid interference of noise [79].

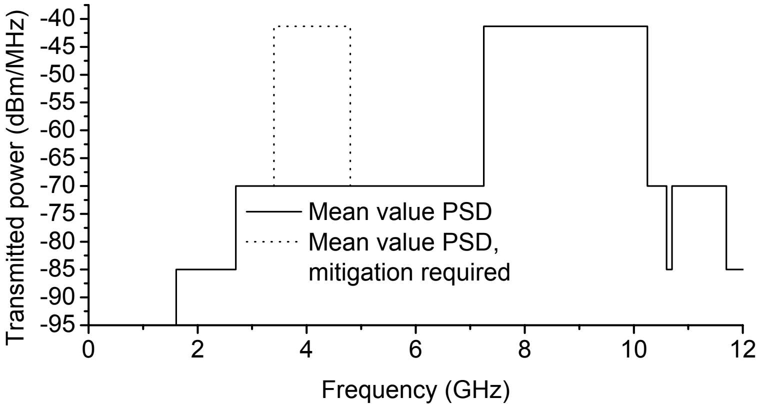


Fig. 52. Mean value radiation emission spectrum mask in Japan.

5. Summary of Papers

Paper I

Wideband Patch Antenna Array for Multi-band UWB

This paper describes a method to use patch antennas in a switched array configuration suitable for multi-band UWB. The paper proposes the concept of electrical parallelism to achieve requirements for a certain multi-band pulse. Furthermore, arrays are scaled throughout the unlicensed frequency band 3.1-10.6 GHz released by FCC with a number of patches used for each pulse. Different sub-arrays are then selected using switches.

The author's contributions: The idea of using patch antennas in parallel to achieve a bandwidth wide enough for one multi-band UWB pulse. Designing of sub-arrays consisting of one to four patches to minimize the total number of patches used for each pulse. Then the sub-arrays are combined with a switch-network to cover the entire 3.1-10.6 GHz frequency band. Wrote the manuscript.

Paper II

An integrated Spiral Antenna System for UWB

This paper reports simulated results of low profile monofilar spiral antennas for UWB, integrated in a PCB. It is found that monofilar spiral antennas can be utilized for UWB. Physical parameters are studied and their impact on antenna gain, CF, and relative bandwidth are presented. It is also shown that monofilar spiral antennas in parallel can be used to extend the antenna performance. Antenna gain is improved by introducing an additional air core between the antenna and the ground plane.

The author's contributions: Simulation study of how physical parameters impact on the antenna performance. It shows also which physical size is needed for a certain voltage standing wave ration (VSWR). The idea that antenna gain can be improved with an additional air core is shown. It is shown that both a single antenna and a dual antenna array cover the entire frequency band 3.1-10.6 GHz. Wrote the manuscript.

Paper III

Monofilar Spiral Antennas for UWB with and without Air Core

This paper reports an implementation of low profile monofilar spiral antennas for UWB, integrated in a PCB. It is verified that monofilar spiral antennas can be utilized for UWB. Physical parameters are studied and their impact on antenna gain, CF, and relative bandwidth was presented. It is proved by measurements that the antenna gain can be improved by introducing an additional air core between the antenna and the ground plane.

The author's contributions: Experimental study of how physical parameters impact on the antenna performance, in order to reveal which physical size is needed for a certain VSWR performance. The antenna gain with various physical antenna specifications is measured in a low reflection measurement chamber. The idea that antenna gain can be improved with an additional air core is also proved. It is shown that a single spiral antenna covers the entire frequency band 3.1-10.6 GHz. Wrote the manuscript.

Paper IV

Air Core Patch Antennas Suitable for Multi-band UWB

This paper presents a substrate solution that increases the gain of patch antennas. It is studied how an additional air core affects antenna gain and relative bandwidth. The study includes a number of different substrate heights and patch widths at two different frequencies.

The author's contributions: The idea that the patch antenna gain can be improved with an additional air core for UWB. The idea is further studied at two different frequencies with various antenna configurations. Patch antennas with and without an additional air core are studied and compared. Different heights of the additional air core are used as well. Wrote the manuscript.

Paper V

Frequency-multiplexed Inverted-F Antennas for Multi-band UWB

This paper describes a method to use inverted-F antennas in a multiplexed array configuration suitable for multi-band UWB. The paper proposes the concept of multiplexed parallelism to achieve requirements for a certain multi-band system. In this paper a radio front-end system consisting of three antennas and a frequency multiplexing network is built. Both the antennas and the frequency multiplexing network designed to be integrated in a printed circuit board.

The author's contributions: Design of the system and its sub-components. Calculation and simulation of suitable filters, and design of the inverted-F antennas. All simulations and measurements presented in the final paper. System simulation on theoretical level, simulation and design of coupled line filters with EM-simulations. All simulations, design of antenna prototypes, and all measurements. Wrote the manuscript.

Paper VI

All-Microstrip Design of Three Multiplexed Antennas and LNA for UWB Systems

A fully printed circuit board integrated triplexed antenna LNA system design suitable for multi-band OFDM UWB operating between 3.1-4.8 GHz. The triplexer utilizes a microstrip network and three edge-coupled filters.

The author's contributions: Design of the multiplexer and the antennas. Calculation and simulation of the frequency multiplexing network. All antenna simulations and measurements. All design and simulation of coupled-line filters, both schematic and EM simulations. Partially performed system set-up and simulations. Partially wrote the manuscript.

Paper VII

Frequency Triplexer for Ultra-wideband Systems Utilizing Combined Broadside- and Edge-coupled Filters

A fully integrated triplexer in a printed circuit board suitable for multi-band OFDM UWB operating between 3.1-4.8 GHz. The triplexer utilizes a microstrip network and three combined broadside- and edge-coupled filters. Separate filters were designed, simulated, manufactured, and measured for scientific comparisons.

The author's contributions: Design of the system and its sub-components. Calculation, and simulation of suitable filters and frequency multiplexing network. All simulations, except a few optimizations. All simulations and measurements presented in the final paper. System simulation on theoretical level, simulation and design of separate coupled lines, and combined broadside- and edge-coupled filters with EM-simulations. Wrote the manuscript.

Paper VIII

A Frequency-Triplexed Inverted-F Antenna System for Ultra-wide Multi-band Systems 3.1-4.8 GHz

This paper describes a method to use inverted-F antennas in a multiplexed array configuration suitable for multi-band UWB. The paper proposes the concept of multiplexed parallelism to achieve requirements for a certain multi-band system. In this paper a radio front-end system consisting of three antennas and a frequency multiplexing network is built. Both the antennas and the frequency multiplexing network designed to be integrated in a printed circuit board.

The author's contributions: Design of the system and its sub-components. Calculation and simulation of suitable filters, design of the inverted-F antennas. All simulations and measurements presented in the final paper. System simulation on theoretical level, simulation and design of coupled line filters with EM-simulations. All simulations, design of antenna prototypes, and all measurements. Wrote the manuscript.

Paper IX

Microstrip Bias Networks for Ultra-Wideband Systems

In this paper three different distributed RF-choke geometries are studied. Their performance and robustness against reflections from the DC-supply are studied. Their sensitivity for reflections when the DC-port is terminated with, low, high, and matched impedances is discussed. Both simulated and measured results are presented and discussed.

The author's contributions: Partially performed the simulations. Responsible for the manufacture of printed circuit board prototypes. Moreover, responsible for measurements and data extraction. Partially wrote the manuscript.

Paper X

A Frequency-triplexed RF Front-end for Ultra-wideband Systems 3.1-4.8 GHz

This paper describes a triplexed RF front-end system integrated in a printed circuit board suitable for multi-band OFDM UWB operating between 3.1-4.8 GHz. The front-end utilizes a microstrip network and three combined broadside- and edge-coupled filters.

The author's contributions: Design of the multiplexer, calculation and simulation of the frequency multiplexing network. All design and simulation of separate broadside- and edge-coupled filters with EM-simulations. Layout of the multi-layer PCB. Partially performed system set-up and simulations. Performed most of the measurements. Partially wrote the manuscript.

Paper XI

Mono- and Di-pole Antennas for UWB Utilizing Flex-rigid Technology

This paper demonstrates implementation of mono- and di-pole antennas on flex-rigid substrate. The flex-rigid concept is an interesting approach to build usable high performance wide band antennas suitable for UWB.

The author's contributions: Design and simulation of the antennas. All simulations, design of antenna prototypes. Wrote the manuscript.

References

[1] P. Lorrain, D. R. Corson, “Electromagnetism Principles and Applications – 2nd edition,” W. H. Freeman and Co., ISBN: 0-7167-2096-5, New York, 1997.

[2] H. Schantz, "The Art and Science of Ultrawideband Antennas," Artech House Inc., ISBN: 1-58053-888-6, 2005.

[3] D. Ghosh, A. De, M. C. Taylor, T. K. Sarkar, M. C. Wicks, E. L. Mokole, "Transmission and Reception by Ultra-Wideband (UWB) Antennas," *Antennas and Propagation Magazine, IEEE*, vol. 48, no. 5, pp. 67-99, Oct. 2006.

[4] L. Yang and G. B. Giannakis, “Ultra-Wideband Communications, an Idea Whose Time has Come”, *IEEE signal processing magazine*, pp. 26-54, November 2004.

[5] H. G. Schantz, “A Brief History of UWB Antennas,” *IEEE Aerospace and Electronic Systems Magazine*, vol. 1, pp. 22-26, Apr. 2004.

[6] Z. N. Chen and M. Y. W. Chia, “Broadband Planar Antennas, Design and Applications,” ISBN: 0-470-87174-1, John Wiley & Sons, Inc., 2006.

[7] V. F. Fusco, Foundations of Antenna Theory and Techniques, Edinburgh Gate, Harlow, Essex, England, Pearson Education Limited, 2005.

[8] B. Kim, S. Nikolaou, G. E. Ponchak, Y.-S. Kim, J. Papapolymerou, M. M. Tentzeris, "A curvature CPW-fed ultra-wideband monopole antenna on liquid crystal polymer substrate using flexible characteristic," *IEEE Antennas and Propagation Society Int. Symp. 2006,* pp. 1667-1670, 9-14 July 2006.

[9] Y.–C. Guo, X.–W. Shi, "UWB Antenna with Improved Efficiency for Pulse Radiation," *TENCON 2006. 2006 IEEE Region 10 Conference*, pp. 1-4, Nov. 2006.

[10] A. A. Lestari, A. B. Suksmono, E. Bharata, A. G. Yarovoy, L. P. Ligthart, "Small UWB antenna with improved efficiency for pulse radiation," *Antenna Technology: Small Antennas and Novel Metamaterials, 2005. IWAT 2005. IEEE International Workshop on*, pp. 295-298, 7-9 Mar. 2005.

[11] A. A. Lestari, A. G. Yarovoy, L. P. Ligthart, E. T. Rahardjo, "A UWB Antenna for Impulse Radio," *Vehicular Technology Conference, 2006. VTC 2006-Spring. IEEE 63rd*, vol.6, pp. 2630-2634, 2006.

[12] L. Schreider, X. Begaud, M. Soiron, B. Perpere, "Archimedean microstrip spiral antenna loaded by chip resistors inside substrate," *Antennas and Propagation Society International Symposium, 2004. IEEE*, vol. 1, pp. 1066-1069, 20-25 Jun. 2004.

[13] P. V. Kholod, T. N. Ogurtsova, G. P. Pochanin, "Comparison Characteristics of Small-Size UWB Receiving Antennas," *Ultrawideband and Ultrashort Impulse Signals, The Third International Conference* , pp. 182-184, Sep. 2006.

[14] P. L. Tokarsky, "Coupling Effects in Resistive UWB Antenna Arrays," *Ultrawideband and Ultrashort Impulse Signals, The Third International Conference*, pp. 188-190, Sep. 2006.

[15] F. Sagnard, D.-L. Ton, J. Badet, "Analysis of the Transient Waveforms Radiated by a Combination of Thin-Wire Antennas Part II: Simulation Results," *Antennas and Wireless Propagation Letters* , vol.5, no.1, pp.213-216, Dec. 2006.

[16] A. S. Turk, D. A. Sahinkaya, M. Sezgin; H. Nazli, "Investigation of Convenient Antenna Designs for Ultra-Wide Band GPR Systems," *Advanced Ground Penetrating Radar, 2007 4th International Workshop on*, pp. 192-196, 27-29 Jun. 2007.

[17] M. Karlsson and S. Gong, “Wideband patch antenna array for multi-band UWB,” *Proc. IEEE 11th Symp. on Communications and Vehicular Tech.*, Ghent, Belgium, Nov. 2004.

[18] M. Karlsson, and S. Gong, "A Frequency-Triplexed Inverted-F Antenna System for Ultra-wide Multi-band Systems 3.1-4.8 GHz," *Accepted for publication in ISAST Transactions on Electronics and Signal Processing*, 2007.

[19] J. Ibrahim, R. M. Buehrer, "NBI Mitigation for UWB Systems Using Multiple Antenna Selection Diversity," *Vehicular Technology, IEEE Transactions on*, vol. 56, no. 4, pp. 2363-2374, July 2007.

[20] C. Kihun, K. Hyungrak, Y. Young Joong, "Multi-resonance UWB antenna with improved band notch characteristics," *Antennas and Propagation Society International Symposium, 2005 IEEE* , vol. 3A, pp. 516-519, 3-8 July 2005.

[21] H. G. Schantz, G. Wolenec, E. M. Myszka, "Frequency notched UWB antennas," *Proc. Ultra Wideband Systems and Technologies, 2003 IEEE*, pp. 214-218, 16-19 Nov. 2003.

[22] I.-J. Yoon, K. Hyungrak, C. Kihun, Y. Young Joong, K. Young-Hwan, "Ultra wideband tapered slot antenna with band-stop characteristic," *Antennas and Propagation Society International Symposium, 2004. IEEE* , vol.2, pp. 1780-1783 20-25 June 2004.

[23] C.-D. Zhao, "Analysis on the properties of a coupled planar dipole UWB antenna," *Antennas and Wireless Propagation Letters*, vol.3, pp. 317-320, 2004.

[24] W. S. Lee, D. Z. Kim, K. J. Kim; K. S. Son, W. G. Lim, J. W. Yu, "Multiple frequency notched planar monopole antenna for multi-band wireless systems," *Proc. IEEE 35th European Microwave Conf.*, Paris, France, Oct. 2005, pp 535-537.

[25] V. Brankovic, A Jordan, D. Simic, J. Weber, J. Bal, "High Gain Printed Ultra Wideband Antenna Concept," *Mobile and Wireless Communications Summit, 2007, 16th IST*, pp. 1-5, 1-5 July 2007.

[26] Y. Taeyoung, W. A. Davis, "Planar half-disk antenna structures for ultra-wideband communications," *Antennas and Propagation Society International Symposium, 2004. IEEE*, vol. 3, pp. 2508-2511, 20-25 Jun. 2004.

[27] Z. N. Chen, M. J. Ammann, X. Qing; X. H. Wu, T. S. P. See, A. Cai, "Planar antennas," *Microwave Magazine, IEEE*, vol. 7, no. 6, pp. 63-73, Dec. 2006.

[28] A. Kuramoto, "Flat type UWB antenna by printed circuit board and performance," *Antennas and Propagation Society International Symposium, 2005 IEEE* , vol. 1A, no., pp. 561-564, 3-8 Jul. 2005.

[29] D-H Kwon; Y. Kim; N. P. Chubinsky, "A printed dipole UWB antenna with GPS frequency notch function," *Antennas and Propagation Society International Symposium, 2005 IEEE*, vol. 3A, pp. 520-523, 3-8 Jul. 2005.

[30] D.-H. Kwon, Y. Kim, "Suppression of Cable Leakage Current for Edge-Fed Printed Dipole UWB Antennas Using Leakage-Blocking Slots," *Antennas and Wireless Propagation Letters*, vol. 5, no. 1, pp. 183-186, Dec. 2006.

[31] F. Tchoffo Talom; B. Uguen; L. Rudant; J. Keignart; J-F. Pintos; P. Chambelin, "Evaluation and Characterization of an UWB Antenna in Time and Frequency Domains," *Ultra-Wideband, The 2006 IEEE 2006 International Conference on*, pp.669-673, Sep. 2006.

[32] K. Gomi; M. Ishida; S. Hiura, "UWB module with antenna using organic substrates," *Wireless Technology, 2005. The European Conference on*, pp. 399-402, 3-4 Oct. 2005.

[33] J. Powell; A. Chandrakasan, "Differential and single ended elliptical antennas for 3.1-10.6 GHz ultra wideband communication," *Antennas and Propagation Society International Symposium, 2004. IEEE*, vol. 3, pp. 2935-2938, 20-25 Jun. 2004.

[34] M. A. Peyrot-Solis; G. M. Galvan-Tejada; H. Jardon-Aguilar, "State of the art in ultra-wideband antennas," *Electrical and Electronics Engineering, 2005, 2nd International Conference on*, pp. 101-105, 7-9 Sep. 2005.

[35] T. Taniguchi; T. Kobayashi, "An omnidirectional and low-VSWR antenna for the FCC-approved UWB frequency band," *Antennas and Propagation Society International Symposium, 2003. IEEE*, vol. 3, pp. 460-463, 22-27 Jun. 2003.

[36] Malik, W.Q.; Edwards, D.J.; Stevens, C.J., "Angular-spectral antenna effects in ultra-wideband communications links," Communications, IEE Proceedings- , vol.153, no.1, pp. 99-106, 2 Feb. 2006.

[37] G.-Y. Chen; J.-S. Sun; S.-Y. Huang; Y. D. Chen; C.-H. Lin, "Characteristics of UWB antenna and wave propagation," *Intelligent Signal Processing and Communication Systems, ISPACS 2005, Proceedings of 2005 International Symposium on*, pp. 713-716, 13-16 Dec. 2005.

[38] G. Brzezina; L. Roy, L. MacEachern, "Planar antennas in LTCC technology with transceiver integration capability for ultra-wideband applications," *Microwave Theory and Techniques, IEEE Transactions on*, vol. 54, no. 6, pp. 2830-2839, Jun. 2006.

[39] J. Han; C. Nguyen, "Investigation of time-domain response of microstrip quasi-horn antennas for UWB applications," *Electronics Letters*, vol. 43, no. 1, pp. 9-10, Jan. 4 2007.

[40] M. Klemm, I. Z. Kovacs; G. F. Pedersen; G. Triister, "Comparison of directlional and omnl-directlonal UWB antennas for Wireless Body Area Network applications," *Applied Electromagnetics and Communications, ICECom 2005, 18th International Conference on*, pp. 1-4, 12-14 Oct. 2005.

[41] W. L. Stutzman and G. A. Thiele, “Antenna Theory and Design,” 2nd ed., John Wiley & Sons, Inc., 1998.

[42] R. A. Scholtz, R. Weaver, E. Homier, J. Lee, P. Hilmes, A. Taha, and R. Wilson, “UWB Radio Deployment Challenges,” *The 11th IEEE International Symp. on Personal, Indoor and Mobile Radio Communications, 2000*, vol. 1, pp. 620-625, Sep. 2000.

[43] K. S. Shanmugan, “Estimating the Power Spectral Density of Ultra Wideband Signals,” *IEEE International Conf. on Personal Wireless Communications, 2002*, vol. 1, pp. 124-128, Dec. 2002.

[44] S. H. Krishnamurthy, A. Konanur, G. Lazzi, and B. L. Hughes, “Ultrawideband Vector Antennas for Sensing and Possitioning,” *Thirty-Eighth Asilomar Conf. on Signals, Systems, and Computers, 2004*, vol. 1, pp. 1237-1241, Nov. 2004.

[45] Geer, D., "UWB standardization effort ends in controversy," *Computer*, vol.39, no.7, pp. 13-16, July 2006.

[46] Balakrishnan, J.; Batra, A.; Dabak, A., "A multi-band OFDM system for UWB communication," *Ultra Wideband Systems and Technologies, 2003 IEEE Conference on*, pp. 354-358, 16-19 Nov. 2003.

[47] H. Pues and A. van de Capelle, “Accurate Transmission-line Model for the Rectangular Microstrip Antenna,” *IEEE Proceedings*, vol. 131, no. 6, pp 334-339, 1984.

[48] A. G. Derneryd, “Linearly Polarized Microstrip Antennas,” *IEEE Trans. on Antennas and Propagatio*, pp 846-851, 1976.

[49] J. Huang, “The Finite Ground Plane Effect on Microstrip Antenna Radiation Patterns,” *IEEE Trans. on Antennas and Propagation*, vol. AP-31, no. 4, pp 649-653, Jul. 1983.

[50] J. P. Kim and W. S. Park, “Analysis and Network Modeling of an Aperture-Coupled Microstrip Patch Antenna,” *IEEE Trans. on Antennas and Propagation*, vol. 49, no. 6, pp 849-854, Jun. 2001.

[51] K. R. Carver and J. W. Mink, “Microstrip Antenna Technology,” *The Analysis and Design of Microstrip Antennas and Arrays*, pp. 3-25, IEEE Press, New York, 1995.

[52] E. Levine, G. Malamud, S. Shtrikman, and D. Treves “A Study of Microstrip Array Antennas with the Feed Network,” *The Analysis and Design of Microstrip Antennas and Arrays*, pp. 3-25, IEEE Press, New York, 1995.

[53] H. Nakano, Y. Okabe, H. Mimaki, and J. Yamauchi, “A Monofilar Spiral Antenna Excited Through a Helical Wire,” *IEEE Trans. on Antennas and Propagation*, vol. 51, no. 3, Mar. 2003.

[54] H. Nakano, J. Eto, Y. Okabe, and J. Yamauchi, “Tilted- and Axial-Beam Formation by a Single-Arm Rectangular Spiral Antenna with Compact Dielectric Substrate and Conducting Plane,” *IEEE Trans. on Antennas and Propagation*, vol. 50, no. 1, Jan. 2002.

[55] C. Soras, M. Karaboikis, G Tsachtsiris, and V. Makios, “Analysis and Design of an Inverted-F Antenna Printed on a PCMCIA Card for the 2.4 GHz ISM Band,” *IEEE Antennas and Propagation magazine*, vol. 44, no. 1, Feb. 2002.

[56] K. Tsunoda and T. Taga, “Analysis of Planar inverted F antenna using spatial network method,” *Antennas and Propagation Society International Symp*. 1990, vol. 2, pp. 664-667, May 1990.

[57] M.-C. Huynh and W. Stutzman, “Ground plane effects on planar inverted-F antenna (PIFA) performance,” *Microwaves, Antennas and Propagation, IEE Proceedings*, vol. 150, no. 4, pp. 209-213, Aug. 2003.

[58] A. Batra, J. Balakrishnan, G. R. Aiello, J. R. Foerster, and A. Dabak, “Design of a Multiband OFDM System for Realistic UWB Channel Enviorments,” *IEEE Trans. on Microwave Theory and Tech*, vol. 52, no. 9, pp. 2123-2138, Sep. 2004.

[59] X. H. Wu; A. A. Kishk, "An Ultra-wide Band Omnidirectional Monopole Antenna," Antenna Technology: Small and Smart Antennas Metamaterials and Applications, *2007. IWAT '07. International Workshop on*, pp. 345-348, 21-23 Mar. 2007.

[60] Y. Ishiyama and T. Ohtsuki, “Performance Evaluation of UWB-IR and DS-UWB with MMSE-Frequency Domain Equalization (FDE),” *IEEE Global Telecommunications Conf., 2004*, vol. 5, pp. 3093-3097, Dec. 2004.

[61] I. Oppermann, L. Stoica, A. Rabbachin, Z. Shelby, and J. Haapola, “UWB wireless sensor networks: UWEN - a practical example,” *IEEE Communications Magazine*, vol. 42, no. 12 pp. S27-S32, Dec. 2004.

[62] N. Baker, “ZigBee and Bluetooth Strengths and Weaknesses for Industrial Applications,” *Computing & Control Engineering Journal*, vol. 16, pp. 20-25, May. 2005.

[63] X. Li; Z. Wang; Y. Fei; D. Guo, "The Interval Modulation System For The Ultra-Wideband Vehicular Radar," *ITS Telecommunications Proceedings, 2006 6th International Conference on* , pp.282-285, Jun. 2006.

[64] A. Oncu; B.B.M. W. Badalawa; T. Wang, M. Fujishima, "22-29GHz Ultra-Wideband CMOS Pulse Generator for Collision Avoidance Short Range Vehicular Radar Sensors," *Design Automation Conference, 2007. ASP-DAC '07. Asia and South Pacific*, pp. 94-95, Jan. 2007.

[65] T. Fukuda, N. Negoro, S. Ujita, S. Nagai, M. Nishijima, H. Sakai, T. Tanaka, D. Ueda, "A 26GHz Short-Range UWB Vehicular-Radar Using 2.5Gcps Spread Spectrum Modulation," *Microwave Symposium, 2007. IEEE/MTT-S International*, pp.1311-1314, 3-8 Jun. 2007

[66] FCC 02-48, “Revision of Part 15 of the Commission’s Rule Regarding Ultra-Wideband Transmission Systems”, First Report and Order, ET 98-153; adopted: February 14, 2002; released: April 22, 2002.

[67] FCC 04-285, “Revision of Part 15 of the Commission’s Rule Regarding Ultra-Wideband Transmission Systems”, Second Report and Order and Second Memorandum Opinion and Order, ET 98-153; adopted: December 15, 2004; released: December 16, 2004.

[68] FCC 47 CFR, “Part 15-5-4-07,” FCC, May 4, 2007.

[69] ECC Report 64, “The Protection Requirements of Radio-Communications Systems Below 10.6 GHz from Generic UWB Applications,” Helsinki, February 2005.

[70] ECC/DEC/(06)04, “ECC Decision of 24 March 2006 amended 6 July 2007 at Constanta on the harmonised conditions for devices using Ultra-Wideband (UWB) technology in bands below 10.6 GHz”, released: March 2006 and amended 6 July 207.

[71] ECC/DEC/(06)08, “ECC Decision of 01 December 2006 on the conditions for use of Ground-and Wall Probing Radar (GPR/WPR) imaging systems”, December 2006.

[72] ECC/DEC/(06)12, “ECC Decision of 01 December 2006 on the harmonised conditions for devices using Ultra-Wideband (UWB) technology with Low Duty Cycle (LDC) in the frequency band 3.4 – 4.8 GHz”, December 2006.

[73] ECC Report 94, “The technical requirements for UWB LDC devices to ensure the protection of FWA systems”, by Nicosia, December 2006.

[74] Commission Decision 2007/131/EC on allowing the use of the radio spectrum for equipment using ultra-wideband technology in a harmonised manner in the Community, released: February, 2007.

[75] Recommendation ITU-R SM.1754,” Measurement techniques of ultra-wideband transmissions”.

[76] Recommendation ITU-R SM. 1755,” Characteristics of ultra-wideband technology”.

[77] Recommendation ITU-R SM. 1756," Framework for the introduction of devices using ultra-wideband technology".

[78] Recommendation ITU-R SM. 1757," Impact of devices using ultra-wideband technology on systems operating within radiocommunication services".

[79] AWF3/14, "Technical Conditions on UWB Radio Systems in Japan", *MIC*, Japan.