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Modeling and Applications of Dielectric Substrates at Mm-Wave Frequencies and Realization of a New Class of Metamorphic Materials

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Modeling and Applications of Dielectric Substrates at MM-Wave Frequencies and Realization of a New Class of Metamorphic Materials

DISSERTATION

submitted in partial satisfaction of the requirements for the degree of

DOCTOR OF PHILOSOPHY

in Electrical Engineering and Computer Science

by

Anna Papió Toda

Dissertation Committee:
Professor Franco De Flaviis, Chair
Professor Filippo Capolino
Professor Michael Green

2015
DEDICATION

To my family and friends,
for their support and help
in making this thesis possible.
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ABSTRACT OF THE DISSERTATION

Modeling and Applications of Dielectric Substrates at MM-Wave Frequencies and Realization of a New Class of Metamorphic Materials

by

Anna Papió Toda

Doctor of Philosophy in Electrical Engineering and Computer Science

University of California, Irvine, 2015

Professor Franco De Flaviis, Chair

The millimeter-wave frequency band spectrum represents a great opportunity for ultrahigh-speed, short-range wireless communications and a wealth of new applications such as target positioning or tracking. However, a number of challenges remain for this spectrum to be a viable solution for high-volume consumer applications.

One challenge is the determination of the electromagnetic properties of dielectric substrate packaging materials and metamorphic structures. This is vital for the optimal and robust design of beamforming mm-wave systems. In this thesis we present “The Covered Transmission Line Method”, a new method to determine the complex permittivity of printed circuit and packaging materials at the mm-wave frequencies. This method is relevant in that it allows for testing of a variety of materials without changes on the setup and with minimal sample processing.
Some applications however, require more advanced materials, called metamorphic materials, which are able to change their response to electromagnetic waves. We present in here a unique electromagnetically metamorphic material that can undergo four distinct electromagnetic states (Perfect Electric Conductor, Perfect Magnetic Conductor, Perfect Amplification and Perfect Absorption). Fundamental mathematical and electromagnetic analysis has been used to obtain a full wave analytical model of the scattering properties of this novel composite material. For the first time, a truly metamorphic surface that can precisely be tuned to any electromagnetic state has been fabricated and tested. This is achieved by loading the basic elements (printed circular rings) of the surface with active devices (tunnel diodes) that can sweep their terminal resistance from a high negative to a high positive value.

The combination of small wavelength and large available bandwidth make mm-wave target positioning systems a viable option to achieve the desired accuracy. However, such systems require the use of beamforming mm-wave antenna systems to mitigate the high path losses (there are 21 dB more path losses at 60 GHz than at 5 GHz). In this work, we present a novel high-gain beam selection 60 GHz band Grid Array antenna that fulfills the requirements of beamforming, low-cost and small size for integration with mobile devices.

Regarding target positioning applications, we study the impact of using beamforming mm-wave antenna systems on the location precision of multiple targets in a realistic indoor environment. The positioning error is assessed when omnidirectional antennas are used at the receiving sensors and when they are substituted by beamforming antennas.
Chapter 1

INTRODUCTION

Wireless communication systems have experienced a rapid evolution since Guglielmo Marconi received the first transatlantic wireless message on December 12\textsuperscript{th} 1901 [1]. Initially, wireless communications were administrated and controlled by the military and a reduced number of companies for their private use. However, during the last decades of the 20\textsuperscript{th} century, and the first decades of the 21\textsuperscript{st}, better radiocommunications have made possible a wealth of new applications that have sparked consumer demand. Nowadays, wireless technologies have become one of the fastest growing segments of the telecommunications industry and are present in our daily activities under different forms.

The availability of several gigahertz-bandwidth unlicensed Industrial, Scientific, and Medical (ISM) bands in the 60 GHz spectrum represents a great opportunity for ultrahigh-speed, short-range wireless communications and a wealth of new uses, such as target positioning or tracking. However, a number of challenges remain for this spectrum to be a viable solution for high-volume consumer applications.

At millimeter-wave (mm-Wave) frequencies, antennas and systems are highly influenced by the electromagnetic characteristics of substrate and packaging materials. Accurate knowledge of the electrical properties of printed circuit materials and metamorphic structures is vital for the optimal and robust design of such systems. In this thesis, a variety of substrate materials
including metamorphic structures are proposed and considered for present and future mm-wave applications (Chapter 2).

In Chapter 3 we propose a new method to determine the complex permittivity of printed circuit and packaging materials at the mm-wave frequencies. The “covered transmission line method”, which allows for testing of a variety of materials without changes on the setup and minimal sample processing, is explained in detail and its accuracy and measurable permittivity ranges are also discussed. Several printed circuit material measurements are presented to show that packaging materials must be accurately characterized at high frequencies because their properties can appreciably differ from those at lower frequencies. Moreover, depending on the fabrication process, different thickness samples of the same material can have different permittivity values. The complex permittivity variation versus temperature is also analyzed. We show that at low frequencies, this variation is negligible, but at mm-wave frequencies, it can represent up to a 5% variation in the relative permittivity and have a severe impact on the loss tangent.

Chapter 4 focuses on the analysis and realization of a class of metamorphic materials that can have potential applications to mm-wave beamforming systems. Nowadays most metamaterials are characterized using numerical techniques and specialized software. In this thesis we use fundamental mathematical and electromagnetic analysis to obtain a full wave analytical model of this novel metamorphic material. The analytical model allows for quick exploration of the design parameters impact on the structure’s electromagnetic response and therefore rapid material design. For the first time, a tunable metamorphic material that can undergo 4 different metamorphic states (Perfect Electric Conductor (PEC), Perfect Magnetic Conductor (PMC), Perfect Amplification and Perfect Absorption) is fabricated and tested. Of special
interest is the use of active devices (tunnel diodes) that can sweep their terminal resistance from a high negative to a high positive value to bias the material and obtain a smooth transition between the 4 metamorphic states. This is demonstrated with a prototype at 3 GHz.

Despite the various advantages offered, mm-wave communications suffer from high path losses (for instance, there are 21 dB more path losses at 60 GHz than at 5 GHz) that must be compensated to achieve reliable communication links. At mm-wave frequencies the wavelength is small enough to enable the integration of highly efficient antennas into the radio package, thus reducing size and cost. Multiple Antennas in Package (AiP) have been reported, however, if they use only one antenna element, the gain is not high enough to compensate the extra path loss. Microstrip patch array antennas that allow beam scanning have also been proposed for 60 GHz applications. These arrays require complicated feeding networks that introduce loss and they exhibit undesirable bandwidth limitations. Moreover, phase shifters are required to scan the beam thus increasing the overall system cost. Therefore, new beamforming antenna systems need to be studied.

In Chapter 5 we present different 60 GHz beamforming grid array antenna designs [2] fabricated on a printed circuit material electrically characterized using the covered transmission line method of Chapter 3. These antenna designs fulfill the requirements of beamforming, low-cost and small size for integration with mobile devices and are suitable for a wide range of applications including but not limited to short-range communications, positioning systems and beam-scanning systems. The antennas cover the entire 57-64 GHz band, have high efficiencies and gains of over 11 dB on the band. The importance of the substrate choice and characterization is evident from the comparison between simulation and measurement results.
High accuracy indoor wireless localization systems have attracted interest as a wealth of applications such as indoor autonomous navigation, object and body motion tracking for gaming, asset management, etc. are gaining momentum. Moreover, an increasing number of companies that track individuals inside the premises of shopping centers, retail stores, sport arenas and museums are emerging. For them, knowing the precise location of each individual is capital to understanding consumers’ behavior, increasing sales, providing better customer service or enhancing the visitor experience. These companies are using a diversity of technologies, such as WI-FI triangulation, cell-phone signals and proximity beacons to locate each individual, but their accuracy is in the range of meters, which is not sufficient in some cases. The combination of small wavelength and large available bandwidth make mm-Wave positioning systems a viable option to achieve the desired accuracy.

From a research point of view, radio frequency (RF) based indoor positioning systems have been the focus of numerous studies. Great attention has been devoted to the system architecture design in order to obtain higher positioning accuracies. However, no much attention has been paid to the receiving antennas configuration. The majority of tracking or positioning systems use omnidirectional receiving antennas for simplicity. Nevertheless, beamforming antenna systems can be used to filter out some of the multipath in rich scattering environments and reduce the positioning errors.

In the present work (Chapter 6), we study the impact of using beamforming mm-wave antenna systems on the location precision of multiple targets in a realistic indoor environment. The positioning system architecture is a 60 GHz carrier-based ultra-wide band (UWB). The 60 GHz band has been chosen because of the available bandwidth and because it allows for the design of very compact high-gain beamforming antennas that can easily be integrated in the
mobile devices. Moreover, future portable electronic devices are expected to integrate 60 GHz technology, making the proposed tracking system a viable option for commercialization.

The positioning error in an indoor environment is assessed when omnidirectional antennas are used at the receiving sensors and when they are substituted by beamforming antennas. Results show that beamforming systems can effectively mitigate the multipath effects and achieve positioning accuracies close to those achievable in free space scenarios. The performance of omnidirectional systems is severely degraded by multipath and multiple access interference.

Chapter 7 concludes this thesis with final remarks and future work recommendations.
Chapter 2

MM-WAVE SUBSTRATE MATERIALS
INCLUDING METAMORPHIC SURFACES

Semiconductor (CMOS) and SiGe process technologies have now made the design of low-cost highly integrated millimeter-wave (mm-wave) radios possible in silicon [3-5]. In combination with an optimum packaging approach, this represents a unique opportunity to develop the Gb/s radio that could address the increasing demand in terms of data throughput of the emerging broadband wireless communication systems [6].

Packaging of mm-wave components is particularly challenging because of the associated complexity in both the design and fabrication. The analysis of the mm-wave requirements and technology development shows that multichip packages and integration of a considerably higher complexity and multiplicity of electronic functions within the package are necessary. Moreover, the small wavelength allows for integration of the antennas into the radio package, which has a large influence on the package design and implementation. Dimensional tolerances and parasitics of packages and assemblies have increasingly pronounced effects at mm-wave than at lower frequencies. Therefore, higher frequencies often demand high-precision machining, accurate alignment, high-resolution photo-lithography and low-loss materials with well-characterized dielectric properties.
To date, several mm-wave antennas in package solutions have been reported [7-9]. However, most of these designs are realized on ceramic substrates, which lead to higher production costs when beamforming antenna arrays need to be placed on the package. In order to reduce production cost, new low-cost materials need to be considered, trading off between performance and cost.

In this chapter, we introduce different type of substrate materials that can be of use at the mm-wave frequencies. This comprises common substrate materials, as provided by a wealth of manufacturers, and metamorphic surfaces specifically designed for a particular application.

2.1. Common Substrate Materials

Some candidates for mm-wave applications include the organic substrates, such as fiberglass-epoxy composites (FR-4) and high-temperature fiberglass-epoxy composites (FR-5). However, these materials’ dielectric properties have not generally been characterized very accurately in the millimeter-wave bands.

Important properties of substrate materials include low electrical loss, high thermal conductance, anisotropy, low thermal expansion, high interfacial adhesion to metal surfaces or other thin films and surface roughness. Low electrical loss decreases heating and signal attenuation, high thermal conductivity rapidly removes heat from the circuit and low thermal expansion promotes circuit durability. Surface roughness limits the minimum pattern width and spacing and affects the loss and signal integrity performance of a high speed propagating signal. In [10, 11] the significance of anisotropy in substrate materials for microwave integrated-circuit and antenna applications is discussed.
Ceramic substrates include high-temperature co-fired ceramics (HTCC) and low-temperature co-fired ceramics (LTCC). The advantages of ceramic materials over polymers for substrates are durability, low thermal expansion coefficient and relatively high thermal conductivity. The permittivity of ceramics is strongly influenced by the microstructure and grain size and is relatively constant with frequency.

Organic substrates include fiberglass-epoxy composites (FR-4), high temperature fiberglass-epoxy composites (FR-5), polymide-glass and polymide-quarts among others [12]. The composites commonly consist of a mixture of plastics, glass, and/or ceramics, together with reinforcing materials. Typical reinforcing materials used are paper fabric, woven glass cloth, random fiberglass fibers and aramid fiber cloth. Organic materials may be anisotropic. Laminations and woven-glass cloth are usually the cause of dielectric anisotropy. Moreover, the fabric and fiber weaving have some variability due to manufacturing limitations, and this translates into variability in permittivity.

There exist many types of organic substrates with different thermal expansion coefficient, thermal conductance, metal adhesion and roughness that can adapt to 60 GHz requirements. Moreover, proper characterization of permittivity variations along with a robust package design may lead to some of these organic substrates to be a viable solution for the low-end mm-wave devices market. In Chapter 3, a wealth of common PCB substrate materials are electromagnetically characterized at the mm-Wave band, so the reader can assess if they adapt to the requirements of his/her specific design.
2.2. Electromagnetically metamorphic materials

Electromagnetic metamorphism is defined as the transformation of a scattering object of fixed shape and dimensions from one electromagnetic state to another. Early investigations of electromagnetic scattering from objects characterized by an equivalent variable surface impedance $Z_s$ revealed a set of electromagnetic states, with unique scattering properties at a given frequency, which were related to the reflection coefficient:

$$\Gamma = \frac{Z_s - Z_o}{Z_s + Z_o}$$  \hspace{1cm} (1)

where $Z_o = 120\pi \Omega$ is the free space impedance.

A particularly significant observation is the behavior of the backscattered field as the surface impedance $Z_s = R_s + jX_s$ varies. In figure 1, the magnitude and phase of the reflection coefficient as a function of $R_s/Z_o$ for different normalized reactances $X_s/Z_o$ is presented. It is useful to note the different metamorphic states that the surface undergoes when $X_s=0$:

- As $R_s \to 0$, $\Gamma = -1$, the object is a Perfect Electric Conductor (PEC),
- As $R_s \to +Z_o$, $\Gamma = 0$, the object is a Perfect Absorber (Pabs),
- As $R_s \to \infty$, $\Gamma = 1$, the object is a Perfect Passive Magnetic Conductor (PPMC),
- As $R_s \to -\infty$, $\Gamma = 1$, the object is a Perfect Active Magnetic Conductor (PAMC),
- As $R_s \to -Z_o$, $\Gamma = \pm \infty$, the object is a Perfect Amplifier.
Fig. 2.1. Reflection coefficient vs. Rs/Zo for different normalized reactance values Xs/Zo: a) Magnitude, b) Phase.
These states are universal for all scatterers and interfaces with local radii of curvature of the order of or larger than the incident field wavelength.

With the application of the Whittaker-Shannon sampling theorem, the continuous surface impedance can be replaced by a patterned surface with a discrete array of elements that locally have the same variable electromagnetic properties.

The use of patterned surfaces to alter the electromagnetic boundary condition of a metallic PEC surface has been widely studied and has found a widespread of applications. In antenna design, metamorphic surfaces, also known as frequency selective surfaces (FSS) [13], artificial magnetic conductors (AMC)[14] or high impedance surfaces (HIS) [15], are used to miniaturize the antennas [16, 21], eliminate surface waves [17], and steer and enhance the radiation in a given direction [18, 23]. They have also found applications as absorbers or cloaking devices [19, 20], to name a few.

Given the aforementioned applications, patterned surfaces present great potential for 60 GHz systems. Moreover, more recently, tunable devices or circuits have been incorporated in such surfaces to include active control of electromagnetic waves. In [22] a varactor-tunable high-impedance surface is presented. In [23] the radiation pattern of a leaky wave antenna was steered by loading the surface with varactor diodes and changing their biasing point. In [24] the resonance frequency of the antenna was effectively tuned by loading the HIS with diodes that change between their on and off states. In [25] an implementation of a multi-layer patterned structure that could undergo three metamorphic states by electronic control of a series of switches was presented. Also, resistive loading of certain patterned structures has been studied to control the electromagnetic response of a surface versus frequency [26].
If the load resistance of a kind of patterned surfaces is swept, they smoothly change from one metamorphic state to another at a given frequency. Despite the importance of this property, to the authors’ best knowledge, the only efforts towards this objective are given in [27] where a discrete realization of $Z_s$ was achieved in terms of an array of impedance loaded half wire loop antennas. In Chapter 4, we present a realization of a metamorphic surface, consisting of variable impedance loaded planar loops over a ground plane that could potentially have a wealth of applications at mm-waves.
Chapter 3

CHARACTERIZATION OF DIELECTRIC SUBSTRATE MATERIALS

At millimeter-wave (mm-wave) frequencies, antennas and systems are highly influenced by the electromagnetic characteristics of printed circuit and packaging materials. Accurate knowledge of the complex permittivity of the package material is vital for optimal and robust design of such systems. In this chapter, the covered transmission line method is used to determine the complex permittivity at 60 GHz for a wealth of common FR-4 and FR-5 type packaging materials. From the measurements, the need for accurate characterization of the packaging materials at high frequencies becomes evident, since their properties may differ appreciably from those at lower frequencies. Moreover, depending on the fabrication process, different thickness samples of the same material may have varying permittivities. The complex permittivity variation versus temperature is also analyzed. We show that at low frequencies this variation is negligible, but at mm-wave frequencies it can represent up to a 5% variation in the relative permittivity and have a severe impact on the loss tangent. The covered transmission line method is explained in detail and its accuracy and measurable permittivity range are also discussed.
3.1. Dielectric Measurement Methods for the Permittivity of Thin Substrates

Dielectric properties of a certain substrate sample depend on frequency, homogeneity, anisotropy, temperature and surface roughness. No single technique can accurately characterize all materials over all frequencies and temperatures. Each frequency band and loss regime usually requires a different method.

The measurement of thin materials (<1 mm) presents a special challenge in that uncertainty in thickness of the specimen translates into uncertainty in the permittivity. Measurement methods on thin films that depend less on sample thickness and more on transverse dimensions yield more accurate results for the relative permittivity ($\varepsilon_r$).

Fabry-Perot resonators and open resonators \[28-29\] have been used for the measurement of very low-loss, thin, planar samples up to terahertz frequencies. However, preparation of the test samples is critical as measurement results highly vary with the thickness of the specimen and some skill is required to successfully operate the resonators. Another approach is the free-space setup \[29\] where the sample is placed between two antennas and reflection and/or transmission measurements are used to extract the material properties. The limitation of the free space set up technique, as pointed out in \[30\], is that the samples introduce losses larger than the measurement accuracy. This can limit the capability of measuring thin samples of low loss materials. The component of the permittivity perpendicular to the substrate plane ($\varepsilon_{zz}$) is measured using these techniques.

Microstrip rings \[31\] or cavity resonators \[29\] have also been used to determine the permittivity based on the resonance frequency and quality factors.
Measurement methods based on planar transmission lines have been used up to 100 GHz [28, 32, 33]. Measurements of simple transmission lines, such as the microstrip line method, can be used to extract the properties of the substrate based on the propagation constant measurement. The covered transmission line method [28, 34-38] can also be used to extract the properties of any material covering the transmission line. This last method has the advantage of allowing the measurement of a number of different substrates with a fixed test setup. However, it requires a priori knowledge of the host transmission line material’s electrical properties. The components of the permittivity parallel to the substrate plane ($\varepsilon_{xx}$ and $\varepsilon_{yy}$) can be measured using these techniques.

With the exception of crystalline substrates such as sapphire and quartz, the bulk of materials used as substrates for microwave integrated-circuit applications exhibit varying degrees of anisotropy, with $\varepsilon_{xx}=\varepsilon_{yy}\neq\varepsilon_{zz}$ [10]. To measure anisotropy different techniques should be combined to obtain the relative permittivity components in the parallel ($\varepsilon_{xx}$ and $\varepsilon_{yy}$) and perpendicular ($\varepsilon_{zz}$) planes to the substrate. However, given the most common integrated-circuit transmission line propagation characteristics, determination of $\varepsilon_{xx}$ and $\varepsilon_{yy}$ is in general sufficient for accurate modeling of the circuit characteristics.

### 3.2. Covered Transmission Line Method

This technique represents a modification of the two-layer stripline method [32] and has been extensively used at low frequencies (up to 10 GHz) [35, 37, 38], and up to the mm-Wave band [28]. The permittivity extraction procedure is based on the measurement of the propagation constant of a covered transmission line. In [28] an implementation using a coplanar waveguide (CPW) as a transmission line is presented. In our study we use a 50 $\Omega$ microstrip
line etched on a “standard” substrate, whose complex permittivity \((\varepsilon_r_1, \tan\delta_1)\) is determined beforehand using some other accurate measurement technique (for instance using the open resonator method). The standard substrate is permanently connected to a network analyzer using two V-connectors placed at the ends of the microstrip line. The unknown test substrates (with complex permittivities defined by \(\varepsilon_r_2\) and \(\tan\delta_2\)) are etched free of copper on both sides and cut down to appropriate dimensions to fit the setup. The test piece should be greater than two wavelengths wide but can be of any suitable length and is placed covering the central part of the microstrip line.

In figure 3.1, a schematic representation of the test fixture is presented. It is important to notice the air gap that forms between the test sample and the standard substrate since this may determine the final accuracy of the results. In an effort to minimize this gap a series of holes may be drilled at a certain distance from the microstrip line to apply vacuum and bring the test piece closer to the standard substrate. The addition of these holes impacts the microstrip line propagation constant, but does not affect the measured permittivity given the permittivity extraction procedure employed (see section 3.2.1).

The complex permittivity is obtained by measurement of the S-parameters with and without the test substrate and working backwards in an electromagnetic (EM) simulation tool. In the following a detailed description of the RF test fixture, the measurement setup and complex permittivity de-embedding process are given.
The covered transmission line technique can be used to measure the complex permittivity of a number of different substrates of different dielectric constants and thicknesses with the same setup. The method requires the accurate determination of the dielectric constant of the standard substrate but is almost insensitive to discontinuities in connectors (the same cables as connected for one measurement are kept throughout all measurements) and microstrip-coated microstrip line junctions (this effect is also captured in the EM simulation). Moreover, the
error introduced in the measurement because of the substrate dimension tolerances is negligible. The variation of the dielectric constant over frequency can easily be measured by changing the test frequency and using the standard substrate permittivity corresponding to that frequency in the simulations.

**3.2.1. Complex Permittivity Measurement Procedure**

The complex permittivity of the test sample can be found following the method depicted in figure 3.2. The S-parameters of the covered and uncovered transmission line are measured. Then, the increase in transmission losses $\Delta \alpha$ and phase difference $\Delta \theta$ between both measurements is calculated using:

\[
\Delta \theta = \text{phase}(S_{12})_{cov} - \text{phase}(S_{12})_{uncov}
\]

\[
\Delta \alpha \text{ (dB)} = \text{mag}(S_{12})_{cov} - \text{mag}(S_{12})_{uncov}
\]  

(1)

$\Delta \theta$ and $\Delta \alpha$ account for the electrical length change due to the difference of the effective dielectric constant in the covered microstrip section compared to the bare microstrip line. In section 3.2.4, a detailed derivation of (1) is given.

The setup is then replicated using an EM simulation tool and the S-parameters are extracted for the microstrip line scenario and the covered microstrip line scenario. For the covered case, approximate values of the complex permittivity and air gap dimensions are considered. Then, the simulated $\Delta \theta$ and $\Delta \alpha$ values are calculated and compared with the measured ones and an iterative curve fitting process, with varying test substrate permittivity and air gap dimensions, is run until convergence is achieved.
In figure 3.3 an example of the curve fitting process to obtain the permittivity of a h=762 μm thick R4350 substrate material from Rogers Corporation is detailed. With the initial guesses on air gap dimensions and complex permittivity of the test substrate, the measured and Iteration I simulated Δθ curves have different slopes. Careful analysis and optimization have shown that for a given permittivity value range \( \varepsilon_r \pm 0.5 \) the curve’s slope is mainly affected by the air gap, which indicates that the initial guess was not accurate. Air gap dimensions are then re-adjusted until both Δθ curves have the same trend (Iteration II). To finally match the Δθ curves the relative permittivity is varied until convergence is achieved (Iteration III). After this process, the Δα measured and simulated curves may still be offset, since this curve is most affected by the loss tangent (\( \tan\delta \)). Further optimization with varying loss tangent values leads to convergence of both parameters and therefore to the correct complex permittivity values (Iteration IV). Notice that modifications of either the estimated relative permittivity or the loss
tangent will impact the transmission parameter S21 both in magnitude and phase. Therefore, in Iteration IV, simultaneous adjustments on the estimated loss tangent and relative permittivity may be done to achieve final convergence of both $\Delta \theta$ and $\Delta \alpha$ curves simultaneously.

![Complex permittivity curve-fitting procedure.](image-url)

Fig. 3.3. Complex permittivity curve-fitting procedure.
3.2.2. RF Test Fixture Design

In order to be able to make successful measurements, careful attention has to be paid to the selection of the standard substrate and the microstrip line dimensions. The first requirement is that the connector to microstrip transition introduces minimal reflection. This requires the width of the microstrip line to be comparable to the diameter of the V-connectors center pin. The second requirement is that the standard substrate is fully electrically characterized at 60 GHz and finally, that it is also low loss at this frequency.

After some research, our choice is in favor of Rogers RT/d 5880 high frequency material, from Rogers Corp. This substrate has excellent electrical stability over frequency, the lowest loss tangent of their product portfolio and measured permittivity data is provided up to 50 GHz. Figure 3.4 illustrates the permittivity and loss tangent versus frequency of the RT/d5880 as provided by the manufacturer and obtained using the resonant cavity permittivity measurement technique.

Fig. 3.4. Complex permittivity of RT/d 5880 versus frequency.
Considering that the center-connector diameter of the V-connectors being used is 250 \( \mu \)m, the substrate thickness has been chosen as \( H_1 = 127 \mu m \). This choice leads to a 50 \( \Omega \) microstrip line design at 60 GHz with a width (w) of 376 \( \mu m \), which facilitates the transition. The microstrip thickness (T) is 17 \( \mu m \). Table 3.1 summarizes the RF test fixture design parameters. The substrate permittivity is given at 60 GHz.

<table>
<thead>
<tr>
<th>Material</th>
<th>RT/d 5880</th>
</tr>
</thead>
<tbody>
<tr>
<td>( \varepsilon_r )</td>
<td>2.195</td>
</tr>
<tr>
<td>( \tan \delta )</td>
<td>0.001</td>
</tr>
<tr>
<td>( H_1 )</td>
<td>127 ( \mu m )</td>
</tr>
<tr>
<td>w</td>
<td>376 ( \mu m )</td>
</tr>
<tr>
<td>T</td>
<td>17 ( \mu m )</td>
</tr>
</tbody>
</table>

Table 3.1. RF test fixture design parameters.

The designed prototype consists of 3 different pieces that are mounted together. The first piece contains the microstrip line, air extraction holes and vacuum chamber. It consists of a 127 \( \mu m \) duroid RT/d 5880 substrate over a 6mm brass piece that acts as a ground plane and provides mechanical support. Copper on the top side of the duroid is etched to build the 50 \( \Omega \) microstrip line. On the back side of the brass the vacuum chamber is milled and small holes are drilled at a distance from the microstrip line and along it. A hole from one of the sides of the brass onto the vacuum cavity is drilled to allow connection of a vacuum pump. A 12.7 mm brass piece is added underneath to seal the vacuum chamber. A series of mounting holes and pins for connection and alignment with other parts have been included.

The remaining pieces allocate the V-connectors that allow interconnection between 1.85 mm coaxial cables and the microstrip line. Machining dimensions for the mounting holes required for installation of the microstrip V female sparkplug connector V102F are critical. Therefore,
a series of long mounting holes have been drilled to allow precise mating with the first piece and ensure good contact between the glass bead and the microstrip line.

The complete RF test fixture, as it is after assembly, is shown in figure 3.5.

![Designed RF test fixture.](image)

**3.2.3. Measurement Setup**

Dielectric permittivity measurements are made in the frequency domain using the E8361C PNA Network Analyzer from Agilent Technologies. The RF test fixture is connected to the PNA via two 1.85 mm coaxial connectors and to a vacuum pump as shown in Figure 3.6. Accurate calibration of the system, up to the 1.85 mm coaxial cable tips, has been performed using a SOLT method.

The materials to be tested, with thicknesses ranging from 100 μm to 787 μm, have been cut into coupons of about 4 x 5 cm² and etched free of copper.
3.2.4. Permittivity Measurement Range and Accuracy

Given a permittivity measurement method, it is of interest to know the range of measurable permittivities and the factors that determine the measurement accuracy. In this section, we analyze the covered transmission line method permittivity extraction procedure mathematically. This allows us to determine the range of measurable permittivities, taking into account the test setup parameters. Also, this gives us insight into how the EM model of the setup impacts the de-embedded permittivity accuracy.

The covered transmission line setup can be represented as the cascade of 3 different $T$-parameters matrixes: $T_1$ corresponding to the Port 1 to coated transmission line region (Sec.
1), \(T_c\) corresponding to the coated transmission line region (Sec. C) and, \(T_3\) corresponding to the coated transmission line to port 2 section (Sec. 3) as depicted in figure 3.7.

**Fig. 3.7.** Cascade of elements configuring the a) Covered transmission line setup; b) Empty transmission line setup.

Then, the total \(T\)-matrix can be written as:

\[
[T_T] = [T_1][T_c][T_3]
\] (2)

Without loss of generality, if we consider sections 1 and 3 to be equal, with \(T_1 = T_3 = T\), then:

\[
[T_T] = [T][T_c][T] = \begin{bmatrix} T_{T11} & T_{T12} \\ T_{T21} & T_{T22} \end{bmatrix}
\] (3)

Where each term can be expressed as a function of the individual matrix components:

\[
T_{T11} = T_{11}^2 T_{c,11} + T_{c,12} T_{21} T_{11} + T_{c,21} T_{11} T_{12} + T_{c,22} T_{21} T_{12}
\]

\[
T_{T12} = T_{12}^2 T_{c,21} + T_{c,12} T_{22} T_{11} + T_{c,11} T_{11} T_{12} + T_{c,22} T_{22} T_{12}
\]

\[
T_{T21} = T_{21}^2 T_{c,12} + T_{c,11} T_{21} T_{11} + T_{c,21} T_{11} T_{22} + T_{c,22} T_{21} T_{22}
\]

\[
T_{T22} = T_{22}^2 T_{c,22} + T_{c,11} T_{21} T_{12} + T_{c,21} T_{22} T_{12} + T_{c,12} T_{21} T_{22}
\] (4)

If we now make use of the T- to S- parameters transformation, we obtain:
\[ T_{r,22} = \frac{1}{S_{r,21}} = \frac{-\det (S_c) S_{11} S_{22} - S_{22} + S_{c,11} 1 - S_{22}}{S_{c,21} S_{21} S_{21} S_{21}} + \frac{-S_{r,22} S_{11} 1}{S_{c,21} S_{21} S_{21}} + \frac{1}{S_{c,21} S_{21} S_{21}} \]

(5)

or:

\[ S_{r,21} = \frac{S_{c,21} S_{21}^2}{1 - S_{11} S_{c,22} - S_{22} S_{c,11} + \det (S_c) S_{11} S_{22}} \]

(6)

Taking into account that each section is reciprocal, we have:

\[ S_{r,21} = \frac{S_{c,12} S_{12}^2}{1 - 2S_{11} S_{c,11} + S_{c,11} S_{11}^2 - S_{c,12} S_{11}^2} \]

(7)

If we now consider the measurement equipment dynamic range \((D_r)\), the range of measurable permittivities can be found. This is done by observing that \(D_r \geq |S_{r,21}|\) for a successful measurement. For instance, if in our simulation tool we take as a reference a dielectric substrate sample of 38 mm in length to model \([T_c]\) and we sweep the permittivity, we can obtain the plot of figure 3.8, where each solid line defines a different insertion loss \(S_{r,21}\).

Notice that we have considered that \(S = \begin{bmatrix} S_{11} & S_{12} \\ S_{21} & S_{22} \end{bmatrix} = \begin{bmatrix} 0 & 1 \\ 1 & 0 \end{bmatrix}\), which corresponds to perfectly matched lossless sections 1 and 3. The measurement equipment dynamic range \(D_r \geq |S_{r,21}|\) defines a border up to which certain material permittivities will be measurable. For example, with a dynamic range \(D_r = 30 \text{ dB}\), we will be able to measure a permittivity of \(\varepsilon_r = 12\) and \(\tan \delta = 0.02\), but we will not be able to measure a material with \(\varepsilon_r = 12\) and \(\tan \delta = 0.04\). A closed form expression to determine the maximum measurable \(\tan \delta\) given a dynamic range...
and permittivity can be obtained from doing a curve fitting of the data in figure 3.8. Equation (8) gives the relation between the three parameters:

$$\tan\delta = p_1\varepsilon_r^2 + p_2\varepsilon_r + p_3$$  \hspace{1cm} (8)

with:

$$p_1 = 1.0869e^{-6}D_r^2 - 7.6234e^{-5}D_r + 0.0019871$$

$$p_2 = -4.1284e^{-5}D_r^2 + 0.0026445D_r - 0.063968$$

$$p_3 = 4.0819e^{-4}D_r^2 - 0.023331D_r + 0.52644$$

where $D_r$ is expressed in dB. Equation (8) is valid for $-90 \text{ dB} \leq D_r \leq -30 \text{ dB}$ and $1 \leq \varepsilon_r \leq 20$. In figure 3.8, the regression lines (dashed lines) are included for comparison.

Fig. 3.8. Complex permittivity measurement range vs. measurement equipment dynamic range.
If we now consider sections 1 and 3 not to be ideal, new curves can be obtained using (7) for each $S_{11}$-$S_{12}$ pair values, which will diminish our measurement capability. Figure 3.9 shows how the permittivity measurement range is decreased for equipment dynamic ranges of 30 dB and 60 dB as the mismatch of sections 1 and 3 increases from $S_{11} = 0$ to $S_{11} = 0.9$.

Fig. 3.9. Complex permittivity measurement range vs. setup mismatch losses.

Figures 3.8 and 3.9 prove the versatility of the method to measure a wide range of permittivities if careful attention is paid to the selection of the measurement equipment and the design of the setup.

If we now analyze the transmission line of Figure 3.7 b) using the procedure depicted in equations (2) to (7), we obtain:

$$S_{TM,21} = \frac{S_{M,12}S_{12}^2}{1 - 2S_{11}S_{M,11} + S_{M,11}^2S_{11}^2 - S_{M,12}^2S_{11}^2}$$  \hspace{1cm} (9)

And by dividing (7) by (9), we obtain:
\[
\frac{S_{T_{\text{21}}}}{S_{T_{\text{M21}}}} = \frac{S_{c,12} \left\{1-2S_{11}S_{M,11}+S_{M,11}^2S_{11}^2-S_{M,12}^2S_{11}^2\right\}}{S_{M,12} \left\{1-2S_{11}S_{c,11}+S_{c,11}^2S_{11}^2-S_{c,12}^2S_{11}^2\right\}} = \Delta \alpha e^{-j\Delta \theta} \quad (10)
\]

which is equation (1).

Since the permittivity de-embedding procedure is based on the simulation of the setup in an EM tool to match the simulated \(\Delta \alpha\) and \(\Delta \theta\) to the measured values, it is obvious from (10) that the accuracy of the de-embedded permittivity is dependent on the degree up to which the simulation scenario replicates the behavior of the setup. That is, how well we can model \([T]\), \([T_c]\) and \([T_M]\). It can be noted here that the accuracy with which we model \([T_M]\) and \([T_c]\) will drive the error provided that \(S_{11} \approx 0\). This implies that a high quality low loss connector needs to be used and the microstrip line has to be precisely designed to be 50 \(\Omega\). Under these circumstances, the impact of the model for \([T]\) in the error is small.

To obtain a good model for \([T_M]\), the standard substrate material permittivity has to be precisely known, since this will strongly impact the accuracy of the de-embedded permittivity. Apart from the standard substrate permittivity, the model of Sec. C needs to carefully replicate the air gap between the microstrip line and the cover, to minimize the errors.

### 3.3. Complex Permittivity Measurement Results

Figure 3.10, presents a graph summarizing the tested materials. Some organic and ceramic substrates ranging from low to high loss and low to high cost from two different vendors have been considered, so one can evaluate the performance-cost tradeoff.
Fig. 3.10. Summary of tested materials.

The measured complex permittivity of each material is presented in table 3.2. Reference permittivity values given by the vendors are also specified for comparison purposes. When available, different thicknesses of the same substrate have been considered.

<table>
<thead>
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<th>T (µm)</th>
<th>Catalog Data</th>
<th>Measured Data @ 60 GHz</th>
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<tr>
<td></td>
<td></td>
<td>f</td>
<td>εr</td>
</tr>
<tr>
<td>FR370H</td>
<td>127</td>
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<td>Ultralam</td>
<td>102</td>
<td>10</td>
<td>2.9</td>
</tr>
<tr>
<td>R5880</td>
<td>254</td>
<td>50</td>
<td>2.195</td>
</tr>
<tr>
<td>R5880</td>
<td>508</td>
<td>50</td>
<td>2.195</td>
</tr>
<tr>
<td>PTFE</td>
<td>12700</td>
<td>60</td>
<td>2.05</td>
</tr>
</tbody>
</table>

Table 3.2. Complex permittivity measurements of various substrates.
From the table we can observe that materials rated to be used at high frequencies (the vendor permittivity is given at 50 GHz) appear to have very close relative permittivity values at 60 GHz (less than 10% variation in $\varepsilon_r$) but slightly higher loss tangent. The loss tangent variation in some cases is due to the fact that the measured $\Delta \alpha$ curve exhibits slight undulations (see figure 3.3), which are not replicable in simulations. Therefore, we believe that we have achieved good convergence when the simulated values approach the average of the measured one. This will impact the de-embedded permittivity considerably if for different thickness samples the measured curve exhibits different degrees of non-linearity, which happens in general for thicker samples.

For FR-4 type materials (FR370HR and FR408) it can be noticed that the relative measured permittivity decreases as the thickness of the sample increases. This effect can be explained by the fabrication process of the thicker samples, which are made as a stack-up of thinner samples. Special glues, with permittivity close to that of the air, are used between the layers. This implies that the considered stack-up is no longer homogeneous but a mix of material plus glue. As the sample thickness increases, the glue proportion increases and given its lower permittivity, the effective sample permittivity appears to be lower to the electromagnetic wave. At the same time, the loss tangent increases with thickness.

In the RF and microwave bands, the relative permittivity is a monotonically decreasing function of increasing frequency. For high-loss materials, $\varepsilon_r$ decreases faster as frequency increases (dispersion) than it does in low-loss materials. However, it can be observed that for the FR370HR and FR408 materials, the measured permittivities are higher than those provided by the manufacturer at lower frequencies. Differences can be attributed to two different factors: the resin content and the thickness of the measured sample. The permittivity
values provided by the manufacturer were obtained with the Bereskin Stripline [39] method, which uses samples with thicknesses in between 0.5 and 1.5 mm, though the actual measured sample thickness is not specified. As seen from measurements, by increasing the sample thickness, the permittivity lowered, approaching that provided by the Isola Group. Regarding the resin content, Isola’s samples had 50% resin content, but the specific resin content of our tested materials was not specified.

The R5880 material has also been tested to ensure that the standard substrate material permittivity value used in the EM tool is accurate. Since the measured permittivity of the test sample coincides with that of the standard substrate model of the EM tool, no further adjustments need to be made on the EM tool model.

Finally, a sample of PTFE has been measured to provide a comparison with other measurement methods presented in literature. Permittivity results provided here agree very well with those reported in [28] and [40], meanwhile the loss tangent cannot be precisely estimated. For extremely low loss materials the insertion loss is below the equipment measurement uncertainty (see section 3.5), and the $\Delta\alpha$ curve is dominated by noise.

### 3.4. Temperature Behavior

An additional concern that could compromise the performance of mm-W systems is the operating temperature. Due to thermal radiation from integrated circuit components, power supplies, etc., the packages operation temperature is usually higher than the 23-25º C values at which the dielectrics’ complex permittivity is given by the manufacturers. Therefore, the
effect of these higher temperatures on the package dielectric permittivity should be taken into consideration.

For this phase of the investigation, the setup depicted in figure 3.11 was used. It consists of an oven with controllable temperature, where the prototype previously described is placed. A computer program controls the oven, the VNA and a thermocoupler. Thermo-stable coaxial cables have been used to ensure reliable measurements across the temperature range.

![Complex permittivity over temperature measurement setup](image)

Fig. 3.11. Complex permittivity over temperature measurement setup.

Each sample is exposed to a temperature profile going from 0º C to 80º C in 10º C steps, with 25 minutes soak time at each stage. A thermocouple is used to ensure that the prototype and sample have reached the desired temperature at which point the S-parameters are measured.

As a first test, we need to assess what is the permittivity variation versus temperature of the substrate used in the setup (R5880). In figure 3.12, the measured increase in transmission losses $\Delta \alpha /\text{cm}$ and phase difference $\Delta \theta$ for a R5880 sample, with a thickness of 254 µm are
presented. It can be observed that there is a variation of about 10 deg in the phase difference from 0º C to 60º C. This implies that the material relative permittivity is slightly affected by the temperature change. The loss tangent is not appreciably affected by temperature.

Fig. 3.12. Measured transmission loss increase ($\Delta \alpha$/cm) and phase difference ($\Delta \theta$) for a 254 µm thick R5880 sample.
In order to obtain accurate values of the permittivity for the test materials at temperatures different than 20º C, the simulation permittivity of the standard material has to be properly adjusted in the EM model for each temperature of interest. The measured complex permittivity values at 0º C and 60º C for the substrate material R5880 can be found in table 3.3. The relative permittivity variation is 2% at 0º C and 4.8% at 60º C. The loss tangent is unaffected. Taking into account these adjusted permittivity values vs. temperature in our EM tool, the complex permittivity of a series of packaging materials has been measured at different temperatures. The results can be found in Table 3.3. Variations of up to 5% in the relative permittivity with respect to that at 20º C have been measured for all tested materials. The loss tangent variation versus temperature highly depends on the material, with FR4 type materials having the greatest variations.

<table>
<thead>
<tr>
<th>Material</th>
<th>0º C</th>
<th>20º C</th>
<th>60º C</th>
</tr>
</thead>
<tbody>
<tr>
<td>R5880</td>
<td>2.15</td>
<td>2.195</td>
<td>2.3</td>
</tr>
<tr>
<td>(254 µm)</td>
<td>tanδ 0.0012</td>
<td>0.001</td>
<td>0.001</td>
</tr>
<tr>
<td>FR370HR</td>
<td>4.15</td>
<td>4.22</td>
<td>4.38</td>
</tr>
<tr>
<td>(254 µm)</td>
<td>tanδ 0.023</td>
<td>0.028</td>
<td>0.03</td>
</tr>
<tr>
<td>FR408</td>
<td>3.83</td>
<td>3.89</td>
<td>4.054</td>
</tr>
<tr>
<td>(254 µm)</td>
<td>tanδ 0.017</td>
<td>0.019</td>
<td>0.02</td>
</tr>
<tr>
<td>R4350</td>
<td>3.21</td>
<td>3.26</td>
<td>3.36</td>
</tr>
<tr>
<td>(254 µm)</td>
<td>tanδ 0.0065</td>
<td>0.0065</td>
<td>0.0065</td>
</tr>
</tbody>
</table>

Table 3.3. Complex permittivity measurements over temperature.

It is relevant here to compare the permittivity variation of packaging materials at mm-wave frequencies to that of lower frequencies. In figure 3.13 we can observe the measured increase in transmission losses \( \Delta \alpha/\text{cm} \) and phase difference \( \Delta \theta \) for the FR370HR sample, with a thickness of 254 µm from 59 to 61 GHz. The same plot graphs but for the 9-11 GHz range is presented in figure 3.14. Contrary to the results shown in figure 3.13, no significant variation versus temperature is observed. This same trend is observed for all the materials tested.
In view of these results, on-package antenna and circuit designers should be well aware of the permittivity changes vs. temperature for their mm-wave designs. Even if at low frequencies the temperature is not a factor of concern at the mm-wave band it acquires importance. Taking into account the permittivity variation vs. temperature will ensure proper design performance.

Fig. 3.13. Measured transmission loss increase (Δα/cm) and phase difference (Δθ) in the 60 GHz band for a 254 µm thick FR370-HR sample.
Fig. 3.14. Measured transmission loss increase (Δα/cm) and phase difference (Δθ) in the 10 GHz band for a 254 µm thick FR370-HR sample.

3.5. Accuracy Assessment

As has been shown in section 3.2.4., the covered transmission line method accuracy is in general bounded by the accuracy in which the standard substrate permittivity can be
determined. The permittivity provided by the vendor was obtained using a cylindrical cavity resonator designed and used by Damaskos Inc. [41]. However, since we have been able to measure different samples of the substrate material and use our EM tool to find and adjust its permittivity, we are predominantly restricted by the air gap modeling accuracy as discussed in the next paragraphs.

The de-embedded permittivity using the covered transmission line will suffer from uncertainties in the measurement of the length and thickness of the samples under test, the air gap model, the metal surface roughness, the microstrip line matching and the VNA measurement uncertainties. In order to assess the impact of each source of uncertainty, a series of simulations have been conducted. First, we determined what is the change in phase difference and loss increase due to variations of 0.001 in the loss tangent and 0.01 in the relative permittivity. For this purpose the complex permittivity values for 2 of the tested materials, the Duroid 5880 and the FR370HR, which represent a low permittivity and low loss material and a high permittivity and high loss material respectively, were used. Increments of +/- 0.001 in the loss tangent have marginal impact on the phase difference. However, the impact on the loss increase is about 0.11 dB for the FR370 HR and 0.05 dB for the Duroid. The variation in losses is marginal for relative permittivity increments of 0.01. However, the phase difference is about 2.5 deg. for the material with high losses and reduces to 1.2 deg. for materials with low loss.

Based on these results, we then determined the impact in the measured permittivity of a 25.4 μm accuracy on the measurement of the samples length and thickness. With respect to length measurement uncertainties, a 25.4 μm variation does not appreciably change the insertion losses, therefore not influencing the loss tangent accuracy. The phase difference instead, is
only appreciably affected for low permittivity materials. The detected variation is about 0.5 deg., which would correspond to an uncertainty in the relative permittivity of +/- 0.005 according to the results presented in the previous paragraph. The impact of a 25.4 µm variation on the material thickness is more significant. For the high permittivity and high loss material, the variation on the loss increase is about 0.18 dB, which would correspond to an uncertainty in loss tangent of +/- 0.0015. The maximum variation on the phase difference is 10 deg., which corresponds to an uncertainty of +/- 0.04 in the relative permittivity. For the low permittivity and low loss material, a thickness variation of 25.4 µm has no appreciable impact on the loss tangent. The maximum variation on the phase difference is 2 deg. in this case, which corresponds to an uncertainty of +/- 0.015.

Regarding the air gap model, the following test was done. For marginal air gap modifications of 2.5 µm, the slopes varied considerably (see figure 3.15). We considered results converged when slopes had a difference below 0.5 deg. over the 2 GHz bandwidth. This translates into an uncertainty of +/- 0.03 in the measured relative permittivity and +/- 0.0015 in the loss tangent.

Given that the dimensions of the samples under test can be measured with accuracies greater than the 25.4 µm assumed for this analysis, the measurement tolerance will be dominated by the air gap accuracy. This leads to approximately +/- 0.03 uncertainties in the relative permittivity and +/- 0.0015 in the loss tangent estimation. These tolerances are close to those reported by Rogers Corporation at 10 GHz of +/-0.02 on the relative permittivity.
Another possible source of uncertainty in the permittivity measurement is the surface roughness. To assess its impact on the measured permittivity, a surface roughness of 2 μm using the Groisse model was considered in our simulation model. In Table 3.4, the measured permittivity with and without considering surface roughness is given. No appreciable differences were found. Results are coherent with the fact that the permittivity is obtained by
subtraction of two measurements (covered vs. uncovered microstrip line), where the same microstrip line is used. Therefore, the extra phase delay and losses due to surface roughness cancel out between the two measurements.

<table>
<thead>
<tr>
<th>Material</th>
<th>No Surface Roughness</th>
<th>2 µm Surface Roughness</th>
</tr>
</thead>
<tbody>
<tr>
<td></td>
<td>εr</td>
<td>tanδ</td>
</tr>
<tr>
<td>R5880 (254 µm)</td>
<td>2.195</td>
<td>0.001</td>
</tr>
<tr>
<td>FR370HR (254 µm)</td>
<td>4.22</td>
<td>0.028</td>
</tr>
</tbody>
</table>

Table 3.4. Complex permittivity measurement results vs. surface roughness.

The differences between measurement and simulation of sections 1 and 3 of the setup as described in figure 3.7 also have an impact on the permittivity measurement accuracy. To assess the microstrip mismatch effect, the following, which is derived from (10), can be considered:

\[
E = \frac{\left\{1 - 2S_{11}S_{M11} + S_{M11}^2S_{11}^2 - S_{M12}^2S_{11}^2\right\}}{\left\{1 - 2S_{11}S_{C11} + S_{C11}^2S_{11}^2 - S_{C12}^2S_{11}^2\right\}}
\]

(11)

Performing a Monte-Carlo analysis on (11), the worst-case scenario, where the variation in magnitude and phase of \( E \) is maximized for a given \( S_{11} \) range, can be found. Using this approach, the plot of figure 3.16 was obtained, where we considered that \( S_{M11} = -18 \) dB, \( S_{C11} = -14 \) dB, \( S_{M12} = 1 - S_{M11} \) and \( S_{C12} = 1 - S_{C11} \), and the phases of these quantities were left for the Monte-Carlo optimization.

It is then noticeable from figure 3.16 that to minimize the impact of errors, a good matching is necessary since the variations vs. phase reduce for lower \( S_{11} \) values. In our case, we obtained a measured matching of the whole setup better than -13 dB on the band of interest. Considering
that this measurement takes into account the 3 sections of the uncovered setup, the matching
of sections 1 and 3 would be better than -16 dB. Also, the choice of \( S_{M1} = -18 \) dB for this
analysis is proper.

Our simulations showed a matching better than -28 dB in the whole band, which guarantees a
low variation on the magnitude and phase of \( E \) vs. \( S_{11} \) phase. Looking at figure 3.16 a), one
can see that the worst phase variation of \( E \) for \( S_{11} = -16 \) dB vs. \( S_{11} = -28 \) dB for any \( S_{11} \) phase
is of 8 degrees (7 deg. @ -16 dB and -1 deg. @ -28 dB). This translates into a maximum error
in the permittivity determination 0.032 for a high loss material and 0.07 for a low loss one. It
has to be noticed however, that this is the worst-case scenario and in general errors will be
much lower, given different combinations of the \( S_C \) and \( S_M \) phases. In figure 3.17 we show an
eexample for a case when \( S_C \) and \( S_M \) have a phase shift of 40 degrees. The variation in this case
reduces to a maximum of 3 degrees, which stands for a permittivity deviation of 0.012 for a
high loss material.

From figure 3.16 b), the worst amplitude variation of \( E \) for the considered case is of 1.25 dB,
which would correspond to a loss tangent maximum estimation error of 0.011 for a high loss
material. This result indicates that the covered microstrip line method leads to less accurate
measurements of the loss tangent. However, a couple of considerations need to be mentioned.
First, we are considering the worst scenario possible for the phases of \( S_C \) and \( S_M \). If we were to
consider the scenario of figure 3.17, then, the error reduces to 0.4 dB, which is equivalent to
0.0036 error on the loss tangent. For a low loss material, the second consideration is that \( S_C \)
and \( S_M \) will vary little in magnitude, and the error will consequently be reduced.

Measurements of different samples of the same low loss material confirm that the measured
permittivity variation despite varying dimensions of the sample, are very minimal (Table 3.5).
Fig. 3.16  a) Phase; b) Magnitude of $E$ as $S_{11}$ varies for the worst-case phase combination of $SM_{11}$ and $SC_{11}$ ($|SM_{11}|=-18$ dB, $|SC_{11}|=-14$ dB), leading to a wider spread in $E$. 
Finally, the errors due to the VNA measurement uncertainty need to be considered. The random measurement uncertainty is approximately +/- 1.5 deg., which corresponds to the ripple observable in figures 12 and 13. In our approach, simulation results are matched to the
best linear fitting curve, to minimize the impact of these random measurement errors. In the worst-case scenario, however, the error is of +/- 0.01 in the permittivity for a low loss material and +/-0.005 for a high loss material.

The uncertainty in the measurement of magnitude is +/- 0.01 dB for low loss materials, which translates into a tangent delta variation of +/- 0.0002. For higher loss materials, the random variation was of the same order. However, some of these materials exhibited a non-linear variation of $\Delta \alpha$ vs. frequency (see figures 3.3 and 3.13), which could not be replicated in simulations. In all scenarios, we used a linear regression line to match measurements and simulations, incurring in a higher error for materials that presented more variation. The recorded maximum deviation from the linear regression line was +/-0.15 dB, which would correspond to a tangent delta variation of about +/- 0.0014.

Therefore, from our simulations and theoretical analysis, we can conclude that the measured permittivity accuracy is mainly affected by the air gap model and the microstrip setup matching. Results indicate that the covered microstrip line method has accuracy on the order of +/- 0.05 in the permittivity determination but has lower accuracies in the measurement of the loss tangent.

In order to validate the accuracy values in the previous paragraphs, different samples of two materials were tested. First, we measured three different samples of CCL-HL832MG (CCL), with $\varepsilon_r = 3.54$ and $\tan\delta = 0.009$ according to the provider Mitsubishi Gas Chemical Co., with varying lengths and along the two main sample directions and obtained the permittivity for each of them. The test results of these measurements are in Table 3.5. The results show a relative permittivity variation of $\pm 0.01$ and a loss tangent variation of $\pm 0.0002$ across the
different measurements. The same results were found for another material, CCL2. The x and y directions for each sample were taken randomly, so variations between \( \varepsilon_{xx} \) and \( \varepsilon_{yy} \) cannot directly be related to anisotropy. Instead, differences can be attributed to measurement accuracy limitations. It is also noticed that the measured relative permittivity is significantly lower than that provided by the manufacturer and the loss tangent is greater. Measurement results, however, are verified in the next section.

<table>
<thead>
<tr>
<th>Material</th>
<th>Direction</th>
<th>( \varepsilon_r )</th>
<th>( \tan\delta )</th>
</tr>
</thead>
<tbody>
<tr>
<td>CCL (Sample 1)</td>
<td>Y</td>
<td>3.35</td>
<td>0.012</td>
</tr>
<tr>
<td>CCL (Sample 1)</td>
<td>X</td>
<td>3.36</td>
<td>0.0117</td>
</tr>
<tr>
<td>CCL (Sample 2)</td>
<td>Y</td>
<td>3.34</td>
<td>0.012</td>
</tr>
<tr>
<td>CCL (Sample 2)</td>
<td>X</td>
<td>3.36</td>
<td>0.0117</td>
</tr>
<tr>
<td>CCL (Sample 3)</td>
<td>X</td>
<td>3.36</td>
<td>0.0117</td>
</tr>
<tr>
<td>CCL2 (Sample 1)</td>
<td>X</td>
<td>2.31</td>
<td>0.016</td>
</tr>
<tr>
<td>CCL2 (Sample 2)</td>
<td>Y</td>
<td>2.29</td>
<td>0.018</td>
</tr>
</tbody>
</table>

Table 3.5. Accuracy analysis of complex permittivity measurements.

### 3.6. Results Validation

In order to have additional validation of the permittivity measurement results, a grid array antenna [2] was fabricated using the CCL-HL832MG packaging material. The designed antenna is presented in figure 3.18 and further details can be found in [42]. A picture of the fabricated prototype is shown in figure 3.19.

The simulated and measured return losses are compared in figure 3.20. We can observe that the measured and simulated curves using the measured permittivity of \( \varepsilon_r = 3.36 \) and \( \tan\delta = 0.012 \) agree very well. However, when the permittivity provided by the manufacturer is used, the resonance frequency is shifted about 2 GHz. Therefore, we can conclude that the proposed permittivity de-embedding procedure is accurate.
Fig. 3.18. Views and dimensions of the designed grid array antenna: a) Top view; b) Feeding detail; c) Side views.
3.7. Conclusions

In this chapter we have provided the measured complex permittivity at 60 GHz of a variety of common FR-4 and FR-5 type packaging materials. Accurate characterization of the electrical properties of packaging materials at 60 GHz, and at high frequencies in general, is of great
importance for robust EM design. From measurements, we have observed that permittivities at 60 GHz can vary largely from those at 10 GHz (which is the frequency generally provided by the manufacturers). Also, material thickness variations that relate to the material fabrication process have been detected. The complex permittivity variation versus temperature has also been analyzed. We showed that at low frequencies, this variation is negligible, but at mm-wave frequencies, it can represent up to a 5% variation in the relative permittivity and have a severe impact on the loss tangent.

The covered transmission line permittivity measurement method used to obtain the results has been explained in detail and its accuracy has been assessed.
Chapter 4

FULL WAVE ANALYTICAL
MODELING AND REALIZATION OF A
CLASS OF METAMORPHIC
MATERIALS

In this chapter, we present a realization of a metamorphic surface, consisting of variable impedance loaded planar loops over a ground plane. We derive a mathematical model for the reflection coefficient to an incident plane wave of an infinite surface of infinitesimally small loaded loops, and prove that at a given frequency, by continuous sweep of the load resistance from \(-\infty\) to \(\infty\), the surface can undergo all possible metamorphic states. We also corroborate the correctness of the mathematical formulation by comparison with numerical simulations. A practical physical implementation of such a surface at microwave frequencies with a circuitry that can realize the negative and positive resistance sweep is proposed. The surface is fabricated and measured results using a waveguide simulator are presented.

For the first time, a fully tunable metamorphic material that can undergo the states of PEC, PMC, Perfect Absorber and Perfect Amplifier has been successfully fabricated. This result is a major breakthrough and opens the door to a multitude of applications in antenna design and new tunable circuits and surfaces.
4.1. Plane Wave Reflection From an Infinite Array of Loaded Infinitesimal Loops Over a Ground Plane

In this section, the problem of a plane wave reflection from a two-dimensional array of infinitesimal impedance loaded planar loops located at a distance $h$ from a ground plane is analyzed. The ground plane lies in the XY plane and the loops normal $\hat{\mathbf{n}}$ is oriented along the $z$ direction. Initially we assume that the diameter of each loop in the array satisfies the restriction $2a<<\lambda$, where $\lambda$ is the wavelength of the incident wave and that the width $w$ of the loops is negligible. The loops are separated a distance $D_x$ and $D_y$ in the $x$ and $y$ directions respectively. In figure 4.1 we illustrate the structure under analysis.

![Diagram of a planar loop array over a ground plane](image)

Fig. 4.1. Infinite array of impedance loaded loops over a ground plane. Top and side views.

To compute the reflected wave we use image theory, so the equivalent problem is that of two infinite arrays of loops with their normal along the $z$-direction and separated a distance $2h$. 
The excitation field is in this case the original plane wave plus its image. Once we find the induced current in the loops we use array theory to find the reflected wave.

4.1.1. Plane Wave Incidence

The following two cases of plane wave incidence are examined:

Case I – TE Mode:

Let’s assume we have a plane wave with propagation vector defined by:

$$\vec{\Psi} = -\cos \phi_i \sin \theta_i \hat{x} - \sin \phi_i \sin \theta_i \hat{y} - \cos \theta_i \hat{z}$$  (2)

where $\phi_i$, $\theta_i$ define the angle of incidence as shown in figure 4.2.

Transverse electric waves (TE waves) are linearly polarized transverse to the plane of incidence, defined by wave vector $\vec{\Psi}$ and the normal vector to the surface $\hat{n}$. We can therefore find the expression for the incident electric field $\vec{E}_i$ as the cross product of the wave vector and the normal to the ground plane:
\[
\vec{E}_1 \times \vec{n} = -E_o (\sin \phi_i \hat{x} - \cos \phi_i \hat{y}) e^{jk\nu}.
\] (3)

Using Maxwell’s equations and considering that we are in a time sinusoidal field, the incident magnetic field takes the form:

\[
\vec{H}_1^i = \frac{j}{\omega \mu} \nabla \times \vec{E}_1^i = -\frac{E_o}{\eta_o} (\cos \phi_i \cos \theta_i \hat{x} + \sin \phi_i \cos \theta_i \hat{y} - \sin \theta_i \hat{z}) e^{jk\nu},
\] (4)

with

\[
\eta_o = \frac{\omega \mu}{k} = 120\pi \Omega.
\] (5)

The expressions for the image incident fields are given by

\[
\vec{E}_1^i = E_o (\sin \phi_i \hat{x} - \cos \phi_i \hat{y}) e^{jk\Lambda}
\] (6)

\[
\vec{H}_1^i = -\frac{E_o}{\eta_o} (\cos \phi_i \cos \theta_i \hat{x} + \sin \phi_i \cos \theta_i \hat{y} + \sin \theta_i \hat{z}) e^{jk\Lambda},
\] (7)

where

\[
\Lambda = -x \cos \phi_i \sin \theta_i - y \sin \phi_i \sin \theta_i + z \cos \theta_i.
\] (8)

**Case II – TM Mode:**

Transverse magnetic waves (TM waves) are linearly polarized transverse to the plane of incidence, defined by wave vector \(\vec{\Psi}\) and the normal vector to the surface \(\vec{n}\). Therefore,

\[
\vec{E}_2 = E_o ((\cos \phi_i \cos \theta_i) \hat{x} + \sin \phi_i \cos \theta_i \hat{y} - \sin \theta_i \hat{z}) e^{jk\nu},
\] (9)

\[
\vec{H}_2 = -\frac{E_o}{\eta_o} (\sin \phi_i \hat{x} - \cos \phi_i \hat{y}) e^{jk\nu}.
\] (10)
And the image fields are:

\[
\overrightarrow{E}_{2l} = -E_o (\cos \phi \cos \theta \hat{x} + \sin \phi \cos \theta \hat{y} + \sin \theta \hat{z})e^{jkA},
\]

(11)

\[
\overrightarrow{H}_{2l} = -\frac{E_o}{\eta_o} (\sin \phi \hat{x} - \cos \phi \hat{y})e^{jkA}.
\]

(12)

### 4.1.2. Induced current computations

The position of the \(mn^{th}\) element in the array is defined by the position vector

\[
\overrightarrow{r}_{mn} = x_{mn} \hat{x} + y_{mn} \hat{y},
\]

(13)

where

\[
x_{mn} = mD_x, \quad y_{mn} = nD_y; \quad -\infty < m, n < \infty.
\]

(14)

In order to compute the induced current in the \(mn^{th}\) element of the array of loops, the equivalent circuit for the receiving antenna, shown in figure 4, is considered

![Equivalent circuit for the loops.](image)

where \(V_{I,II}^{I,II}\) indicates the equivalent Thevenin voltage for the loop for the two polarizations and the impedances are given by
\[ Z_{a}^{I,II} = R_{a}^{I,II} + j \omega L_{a}^{I,II}, \quad (15) \]
\[ Z_{L} = R_{L} + j \omega L_{L} - j \frac{1}{\omega C_{L}}. \quad (16) \]

The current induced in the \( mn^{th} \) loop is

\[ I_{mn}^{I,II} = \frac{V_{mn}^{I,II}}{Z_{a}^{I,II} + Z_{L}}, \quad (17) \]

where the voltage \( V_{mn}^{I,II} \) is determined by utilizing

\[ V_{mn}^{I,II} = -\frac{d}{dt}(\varphi_{mn})^{I,II}. \quad (18) \]

The magnetic flux \( \varphi_{mn}^{I,II} \) through the loop is related to the total magnetic field incident on the \( mn^{th} \) element \( H_{i}^{I,II} \) through

\[ \varphi_{mn}^{I,II} = \mu_{0} \int_{S} (H_{i}^{I,II})^{I,II} dS, \quad (19) \]

with \( dS = dS_{xy} \hat{z} \), which is the loop area differential.

Taking into account that due to the image theorem the excitation field is the original plane wave plus its image, from equations (4) and (7) \( H_{i}^{I,II} \) is calculated to be

\[
\begin{align*}
H_{i}^{I,II} &= H_{i}^{I} + H_{i}^{II} \\
&= -\frac{E_{o}}{\eta_{o}} \left\{ \cos \theta_{i} e^{-jk(x \cos \theta_{i} \sin \theta_{i} + y \sin \theta_{i} \sin \theta_{i})} \left[ (e^{jkz \cos \theta_{i} + e^{-jkz \cos \theta_{i}})(\cos \varphi_{i} \hat{x} \\
+ \sin \varphi_{i} \hat{y}) + \sin \theta_{i} e^{-jk(x \cos \varphi_{i} \sin \theta_{i} + y \sin \varphi_{i} \sin \theta_{i})} (e^{jkz \cos \theta_{i} - e^{-jkz \cos \theta_{i} \hat{z}} \right) \right\}.
\end{align*}
\]

\[
(20)
\]
Focusing on the loop located at \( \vec{r}_{mn} = x_{mn}\hat{x} + y_{mn}\hat{y} + h\hat{z} \), and by using the variable changes \( \alpha = \cos \phi \sin \theta \) and \( \gamma = \sin \phi \sin \theta \), the magnetic current through it is
\[
(\vec{H}_{t_{mn}}') = -\frac{E_o}{\eta_o} \{2\cos\theta \cos(k\cos\theta)(\cos\phi\hat{x} + \sin\phi\hat{y})
+ j2\sin\theta \sin(k\cos\theta)\hat{z}]e^{-jk(x_{mn}\alpha + y_{mn}\gamma)}.
\]

We can now proceed to find \( \Phi_{mn}' \) using (19)
\[
\Phi_{mn}' = -j2\mu_o \frac{E_o}{\eta_o} \sin\theta \sin (k\cos\theta) \int \int_S e^{-jk(x_{mn}\alpha + y_{mn}\gamma)} dS_{xy}.
\]

Since we consider the radius of the loops to be much smaller than the wavelength (\( a<<\lambda \)) the following approximation applies, where \( dS_{xy} = \pi a^2 \),
\[
\Phi_{mn}' = -j2\mu_o \frac{E_o}{\eta_o} \pi a^2 \sin\theta \sin (k\cos\theta) e^{-jk(x_{mn}\alpha + y_{mn}\gamma)}.
\]

So, according to (18), \( V_{mn}' \) will be
\[
V_{mn}' = -2\omega \mu_o \frac{E_o}{\eta_o} \pi a^2 \sin\theta \sin (k\cos\theta) e^{-jk(x_{mn}\alpha + y_{mn}\gamma)}.
\]

In a similar fashion we can obtain \( V_{mn}'' \):
\[
\overline{H}_t'' = \overline{H}_t' + \overline{H}_{2t}'' = -2\frac{E_o}{\eta_o} \cos(k\cos\theta)(\sin\phi\hat{x} - \cos\phi\hat{y})e^{-jk(x\alpha + y\gamma)},
\]
\[
\overline{H}_{t_{mn}}'' = -2\frac{E_o}{\eta_o} \cos(k\cos\theta)(\sin\phi\hat{x} - \cos\phi\hat{y})e^{-jk(x_{mn}\alpha + y_{mn}\gamma)}.
\]
\[ \Phi_{mn}^{II} = \mu_0 \mu_o \int_S (\tilde{H}_{mn}^I)^{II} dS = 0. \] (27)

Therefore, for the second case of incidence, the equivalent Thevenin voltage is null and consequently there are no currents induced on the loops. As a result the structure will appear transparent to the electromagnetic wave.

The induced currents for the first case of incidence as given by (17) are

\[ I_{mn} = -2 \frac{1}{Z_a + Z_L} \frac{\omega \mu_o E_o}{\eta_o} \pi a^2 \sin \theta_i \sin (kh \cos \theta_i) e^{-jk(x_{mn}a + y_{mn}y)}. \] (28)

Since \( Z_a = R_a + jX_a \) and \( Z_L = R_L + jX_L \) then

\[ \frac{1}{Z_a + Z_L} = \frac{R - jX}{R^2 + X^2}, \] (29)

with \( R = R_a + R_L ; X = X_a + X_L \). Therefore

\[ I_{mn}^I = -2 \frac{\omega \mu_o E_o}{\eta_o} \frac{R_I - jX_I}{R_I^2 + X_I^2} \pi a^2 \sin \theta_i \sin (kh \cos \theta_i) e^{-jk(x_{mn}a + y_{mn}y)}. \] (30)

### 4.1.3. Reradiated Wave

In order to compute the electromagnetic field radiated by an infinite array of induced currents, it is preferable to work with the electric vector potential. In this case, the contribution of the \( mn^{th} \) element is given by [43]

\[ A_{(e)mn}^{II} = \frac{j \omega \mu \varepsilon R_{mn}^{II} e^{-jkR_{mn}}}{4\pi R_{mn}}, \] (31)

where the distance from the \( mn^{th} \) element to the observation point \( (x,y,z) \) is
\[ R_{mn}^2 = r_m^2 + (nD_y - y)^2, \]  
with

\[ r_m^2 = (z)^2 + (mD_x - x)^2. \]

The magnetic dipole moment for the \( mn^{th} \) element is:

\[ p_{mn}^{\scriptscriptstyle I,II} = I_{mn}^{I,II} \pi a^2 \hat{z} \]  
and therefore,

\[ \bar{A}_\scriptscriptstyle{(e)mn}^{I,II} = \frac{j \omega \mu \varepsilon a^2 e^{-jkR_{mn}}}{4} \frac{I_{mn}^{I,II}}{R_{mn}} \hat{z}. \]  
The total electric vector potential at \((x,y,z)\) from an infinite array is given by

\[ \bar{A}_\scriptscriptstyle{(e)} = \frac{j \omega \mu \varepsilon a^2}{4} \sum_{m=-\infty}^{\infty} \sum_{n=-\infty}^{\infty} I_{mn}^{I,II} e^{-jkR_{mn}} \hat{z}. \]

Considering that the currents can be written as

\[ I_{mn}^{I,II} = i^{I,II} e^{-jk(mD_x + nD_y)}, \]  
with

\[ i^{I,II} = \begin{cases} -2 \frac{\omega \mu_0 E_0 R_l - j X_l}{\eta_0} \frac{\pi a^2 \sin \theta_l \sin (kh \cos \theta_l)}{R_l^2 + X_l^2} & \text{for } \theta_l < \frac{\pi}{2} \frac{\pi}{2} \text{ \(0\)} \end{cases}, \]  
where the upper and lower expressions in brackets refer to incident polarizations I and II respectively, the electric vector potential now takes the form

\[ \bar{A}_\scriptscriptstyle{(e)} = \frac{j \omega \mu \varepsilon a^2}{4} i^{I,II} \hat{z}. \]
The factor $P$ in (39) is equal to

$$P = \sum_{m=-\infty}^{\infty} \sum_{n=-\infty}^{\infty} e^{-jk(mD_x\alpha+nD_y\gamma)} e^{-jkr_{mn}} R_{mn}$$

and can be rewritten as

$$P = \sum_{m=-\infty}^{\infty} e^{-jkmD_x\alpha} \xi_m,$$

where

$$\xi_m = \sum_{n=-\infty}^{\infty} e^{-jknD_y\gamma} e^{-jkr_{mn}} R_{mn}.$$  

Let’s now consider the following transform pair [44]

$$e^{-jk\sqrt{r_m^2+(nD_y)^2}} = \mathcal{F} \left\{ \frac{1}{2j} H_0^{(2)} \left[ r_m \sqrt{k^2 - k^2\gamma^2} \right] \right\},$$

where $\mathcal{F}$ denotes the Fourier transform operation. Using the shift theorem

$$G(nD_y - \gamma) = \mathcal{F} \{ e^{j\gamma} g(k\gamma) \}$$

where $G(nD_y) = \mathcal{F} \{ g(k\gamma) \}$, it follows that

$$\xi_m = \sum_{n=-\infty}^{\infty} e^{-jknD_y\gamma} \mathcal{F} \left\{ \frac{e^{j\gamma}}{2j} H_0^{(2)} \left[ kr_m \sqrt{1 - \gamma^2} \right] \right\}.$$  

If Poisson’s sum formula [45] is adopted i. e.
\[
\sum_{n=-\infty}^{\infty} e^{jn\omega_o t} G(n\omega_o) = T \sum_{m=-\infty}^{\infty} g(k + mT), \quad (46)
\]

with \( T = 2\pi/\omega_o \), \( \xi_m \) yields

\[
\xi_m = -j \frac{\pi}{D_y} \sum_{n=-\infty}^{\infty} e^{jky\left(\gamma - n \frac{\lambda}{D_y}\right)} H_0^{(2)}(k r_m \rho), \quad (47)
\]

with

\[
\rho = \left[1 - \left(\gamma - n \frac{\lambda}{D_y}\right)\right]^{1/2}. \quad (48)
\]

Upon reconsidering the expression for \( P \) i.e.

\[
P = -j \frac{\pi}{D_y} \sum_{n=-\infty}^{\infty} e^{jky\left(\gamma - n \frac{\lambda}{D_y}\right)} \sum_{m=-\infty}^{\infty} e^{-jkmD_x \alpha} H_0^{(2)}(k r_m \rho), \quad (49)
\]

the Fourier pair

\[
H_0^{(2)} \left[ k \rho \sqrt{(z)^2 + (mD_x - x)^2} \right] = \mathcal{F} \left\{ e^{jx\alpha} \frac{e^{-jz\sqrt{(kp)^2 - (k\alpha)^2}}}{\pi \sqrt{(kp)^2 - (k\alpha)^2}} \right\} \quad (50)
\]

and the shift theorem, \( P \) can be rewritten as

\[
P = -j \frac{2\pi}{D_y D_x} \sum_{n=-\infty}^{\infty} e^{jky\left(\gamma - n \frac{\lambda}{D_y}\right)} \sum_{m=-\infty}^{\infty} e^{jkx\left(a - m \frac{\lambda}{D_x}\right)} \frac{e^{-jz\sqrt{(kp)^2 - (k\beta)^2}}}{\pi \sqrt{(kp)^2 - (k\beta)^2}}, \quad (51)
\]

with

\[
\beta = a - m \frac{\lambda}{D_x}. \quad (52)
\]
Let’s now define

\[ r_z = \sqrt{(\rho)^2 - (\beta)^2} = \left[ 1 - \left( \alpha - m \frac{\lambda}{D_x} \right)^2 - \left( \gamma - n \frac{\lambda}{D_y} \right)^2 \right]^{1/2}. \]  

(53)

If we focus on the exponential term of (51) and define

\[ r^2 = \left( \alpha - m \frac{\lambda}{D_x} \right) \hat{x} + \left( \gamma - n \frac{\lambda}{D_y} \right) \hat{y} + r_z \hat{z} \]  

(54)

then

\[ P = -j \frac{2\pi}{k D_x D_y} \sum_{m=-\infty}^{\infty} \sum_{n=-\infty}^{\infty} \frac{e^{j k \vec{R} \cdot \vec{r}}}{r_z}, \]  

(55)

with

\[ \vec{R} = x \hat{x} + y \hat{y} + z \hat{z}. \]  

(56)

On denoting the array unit cell area as \( S = D_x D_y \) and the loop areas by \( A = \pi a^2 \), the electric vector potential takes the form, after combining equations (35) and (55),

\[ A_{(e)}^{(l,II)} = \frac{\omega \mu e A}{2 k S} \sum_{m=-\infty}^{\infty} \sum_{n=-\infty}^{\infty} \frac{e^{j k \vec{R} \cdot \vec{r}}}{r_z} \hat{z}. \]  

(57)

The electric field is derivable from

\[ E_{(e)}^{(l,II)} = -\frac{1}{\varepsilon} \nabla \times A_{(e)}^{(l,II)} = -\frac{j \omega \mu A}{2 S} \sum_{m=-\infty}^{\infty} \sum_{n=-\infty}^{\infty} \frac{e^{j k \vec{R} \cdot \vec{r}}}{r_z} \left( \left( \gamma - n \frac{\lambda}{D_y} \right) \hat{x} - \left( \alpha - m \frac{\lambda}{D_x} \right) \hat{y} \right), \]  

(58)

which can be rewritten in compact form as
\[
E_{II}^{\prime} = -j \frac{\omega \mu A}{2S} \sum_{m=-\infty}^{\infty} \sum_{n=-\infty}^{\infty} \hat{r} \times i_{II}^{\prime} z \frac{e^{jkR \cdot r}}{r_z}.
\] (59)

Now, in our case, taking into account image theory, we have two arrays placed at a distance \pm h from the ground plane. Therefore, the distance from the observation point to the arrays is now \( z' = z \pm h \) and the total scattered field becomes

\[
E_{II}^{\prime} = \frac{\omega \mu A}{S} \sum_{m=-\infty}^{\infty} \sum_{n=-\infty}^{\infty} \hat{r} \times i_{II}^{\prime} z \frac{e^{jkR \cdot r}}{r_z} \sin(khr_z).
\] (60)

### 4.1.4. Computation of the loops Radiation Resistance

In order to calculate the loop radiation resistance, we will use the expression for the total radiated power, given by

\[
P_r = \frac{1}{2} R_a |i_{II}^{\prime}|^2.
\] (61)

However, the radiated power can also be computed through the Poynting vector

\[
\vec{\phi} = \frac{1}{2} \text{Re} \left\{ E_{II}^{\prime} \times H_{II}^{\prime} \right\}.
\] (62)

If we now assume that \( D_x \) and \( D_y \) are chosen such that the \( m=n=0 \) is the only mode present

\[
E_{II}^{\prime} = -\frac{A \omega \mu}{S} i_{II}^{\prime} \sin(khr_z) (\gamma \hat{x} - \alpha \hat{y}) \frac{e^{jkR \cdot r}}{r_z} = E_{x}^{\prime II} \hat{x} + E_{y}^{\prime II} \hat{y},
\]

\[
H_{II}^{\prime} = \frac{k}{\omega \mu} \left\{ r_z E_{y}^{\prime II} \hat{x} - r_z E_{x}^{\prime II} \hat{y} + (-\alpha E_{y}^{\prime II} + \gamma E_{x}^{\prime II}) \hat{z} \right\}.
\] (63)

So the Poynting vector comes out to be
\[
\bar{\Phi} = \frac{1}{2} \frac{k}{\omega \mu} \text{Re}\{\left(\gamma E_x^{l,II} E_y^{l,II} - \alpha |E_y^{l,II}|^2\right) \hat{x}}
\]

\(- \left(\gamma |E_x^{l,II}|^2 - \alpha E_x^{l,II} E_y^{l,II}^*\right) \hat{y} - r_z \left(|E_x^{l,II}|^2 + |E_y^{l,II}|^2\right) \hat{z}\) \hfill (64)

The total radiated power can be found by integration of the Pointing vector on the loop surface

\[P_r = \int \bar{\Phi} \, d\mathbf{S} \approx \bar{\Phi} \mathbf{S}^\ast.\] \hfill (65)

Given that

\[\mathbf{S} = |\mathbf{S}| \hat{S},\] \hfill (66)

the radiated power is

\[P_r = \frac{S k}{2 \omega \mu} r_z \left(|E_x^{l,II}|^2 + |E_y^{l,II}|^2\right).\] \hfill (67)

If we now use (63), the radiated power can be rewritten as

\[P_r = \frac{\omega \mu k A^2}{2 r_z S} \sin^2(kh r_z) (\gamma^2 + \alpha^2).\] \hfill (68)

Equating terms in equations (61) and (78) we obtain the radiation resistance

\[R_a = \frac{\eta_o k^2 A^2}{\cos \theta_i S} \sin^2(kh \cos \theta_i) \sin^2 \theta_i,\] \hfill (69)

where we have used \(r_z = \cos \theta_i\) (derivable from (53)).
4.1.5. Reflection Coefficient

In the half space $y>h$, the reflection coefficient is given by

$$\Gamma^{l,II} = \frac{E_{1l,2l}^{l} + E_{1l,II}^{l}}{E_{1l,2l}^{l}},$$  \hspace{1cm} (70)

i.e. the total field for $y>h$ is given by the superposition of the image field plus the reradiated wave. In the assumption that $D_x$ and $D_y$ are chosen such that the $m=n=0$ is the only mode present and $R= R_a + R_L; X = X_a + X_L = 0$, the reflection coefficient at $z=h$ is given by:

$$\Gamma^l = e^{jk'h \cos \theta_l} \left(1 - \frac{2 \mu^2 \eta^2 \sin^2 \theta_l \sin^2 \theta_i}{\eta_0} \frac{S}{(R_a + R_L) \cos \theta_i} \right),$$

$$\Gamma^{II} = -e^{j2kh \cos \theta_i},$$ \hspace{1cm} (71)

It is interesting to note that for the case of normal incidence, $\theta_i = 0$, the reflection coefficient is 1, thus the array of loops appears to be transparent to the electromagnetic wave.

A deeper look at (71) reveals that under TE plane wave excitation, when $R_L = -R_a$, the reflection coefficient tends to infinity. In this case, the surface will act as a perfect amplifier. When $R_L = R_a$, the reflection coefficient is null, thus the surface behaves as a perfect absorber. As $R_L \to \pm \infty$, $\Gamma = -e^{j2kh \cos \theta_i}$ and, for small $2kh \cos \theta_i$ factors, the surface behaves as an active PEC or passive PEC. When $R_L = 0$, $\Gamma = 1$, so the surface will appear to be a PMC to the incident plane wave. Therefore, the proposed surface can undergo the different metamorphic states described at the beginning of the chapter.
In figure 4.4, we plot the magnitude and phase of the reflection coefficient as a function of the load resistance $R_L$ for an infinite array of infinitesimal loops being excited by a TE plane wave incident at $\theta_i = 45^\circ$ from broadside. The design parameters are $D_x=D_y=5$ mm, $a=1.4$ mm and $h=3.2$ mm. The operation frequency is 3 GHz. One can observe that such surface can undergo the different metamorphic states and that the perfect absorption point occurs at $R_L=R_a=8.3$ m$\Omega$ and the perfect amplification at $R_L=R_a=-8.3$ m$\Omega$. For comparison, simulation results of the metamorphic substrate using the commercial software Ansys HFSS are included. The material shows the response predicted by the theory, but the load resistance to obtain absorption or amplification is greater. This is due to the width assigned to the loops in the simulations, which makes them non-ideal elements.
4.2. Experimental Implementation

An experimental verification for the metamorphic material has been carried out in a rectangular waveguide simulator [46]. In the experimental setup, a metallic rectangular waveguide is terminated in a section of the array containing a single loaded loop over a
ground plane. The section of the array can represent an infinite array because the \( \text{TE}_{10} \) waveguide mode can be decomposed in two \( \text{TE} \) oblique incident plane waves that reflect on the side waveguide walls, and the waveguide perfectly conducting walls mirror an infinite series of images.

The metamorphic material design parameters are specified in Table 4.1. The Duroid 5880 substrate material from Rogers Corporation, with low loss and permittivity, has been chosen to provide the necessary height between the ground and the loops, since it is not possible to have loops over air. A WR-284 waveguide is used, with cross-sectional area of 72.136 \( \times \) 34.036 mm\(^2\). These cell dimensions ensure that only one mode propagates at the frequency of operation of the metamorphic material (\( f = 3 \) GHz). Also, the loop radius is chosen such as for a load resistance \( R_L = 0 \) \( \Omega \) the material behaves like an AMC at 3 GHz.

Note that the loop dimensions are still small compared to the wavelength, but we do not fall any longer under the assumption of infinitesimally small loops. This obeys practical fabrication reasons.

<table>
<thead>
<tr>
<th>Substrate Material</th>
<th>RT/d 5880</th>
</tr>
</thead>
<tbody>
<tr>
<td>( \varepsilon_{r1} )</td>
<td>2.195</td>
</tr>
<tr>
<td>( \tan\delta_1 )</td>
<td>0.001</td>
</tr>
<tr>
<td>( h )</td>
<td>3.175 mm</td>
</tr>
<tr>
<td>( D_y )</td>
<td>34.036 mm</td>
</tr>
<tr>
<td>( D_x )</td>
<td>72.136 mm</td>
</tr>
<tr>
<td>( a )</td>
<td>2.8 mm</td>
</tr>
<tr>
<td>( w )</td>
<td>0.4 mm</td>
</tr>
<tr>
<td>frequency</td>
<td>3 GHz</td>
</tr>
</tbody>
</table>

Table 4.1. Metamorphic material design parameters.

In order to obtain the desired resistance sweep range, including negative resistance values, we opted for loading the loops with tunnel diodes [47]. These devices, under the proper biasing conditions, have a region of operation where the resistance across their terminals is negative.
Moreover, they can also behave as a variable resistor ranging from low to high positive impedances as the biasing is varied.

To effectively bias the tunnel diode, the biasing network has to be integrated into the waveguide simulator. In figure 4.5 a schematic of the waveguide simulator and the metamorphic material unit cell are shown. It can be observed that the ring contains the tunnel diode and a capacitor that blocks the DC on the loop. Also, a couple of vias going from the loop to the bottom of the cell have been drilled in order to create the DC path. At the end of the vias, a couple of chokes are used to prevent RF leakage through the DC path.

![Waveguide simulator and unit cell of the designed metamorphic material.](image)

Fig. 4.5. Waveguide simulator and unit cell of the designed metamorphic material.
4.2.1. Simulation Results

Before proceeding to fabrication of the prototype, the setup described in the previous section, including the biasing network, has been simulated using Ansys Designer in combination with Ansys HFSS.

In figure 4.6 the simulation model is detailed. The biasing network has been included in Ansys Designer using ideal inductors and capacitors and the tunnel diode behavior is modeled through a variable resistor. The loop electromagnetic response is obtained using the full wave simulation software HFSS.

**HFSS Loop Model**

![Simulation model of the designed metamorphic material including biasing network.](image)

Simulation results, which corroborate the metamorphic behavior of the structure, are presented in figure 4.7. In this case however, the points of perfect absorption and perfect amplification
are not symmetrical with respect to the zero load resistance. This is an effect of the biasing network.

Fig. 4.7. Designed metamorphic material simulated reflection coefficient vs. load resistance: a) Magnitude, b) Phase.
4.2.2. Tunnel Diode Characterization

An important step is to assess the behavior of the tunnel diode vs. biasing, so the range of output resistances is known for the frequencies of interest.

The setup used for the tunnel diode characterization is depicted in figure 4.8. The tunnel diode was soldered between the center conductor of a coaxial connector and its ground. Then, the S-parameters as the biasing changed were measured with the vector network analyzer in the frequency range between 2 GHz and 4 GHz. To decouple the RF and DC paths, a Bias T was used. We also used a load resistor of 60 Ω and we constantly monitored the voltage across it to determine the current through the diode at all times. A 10 dB coupler was inserted in the setup to be able to monitor the tunnel diode frequency response in the spectrum analyzer and rest assured that it was not under oscillation condition.

Fig. 4.8. Tunnel diode characterization setup.
For the prototype, we have chosen the Aeroflex Metelics MDB5057-E28 beam lead tunnel diodes, which have a 0 V bias operation. In figure 4.9 their measured impedance versus bias voltage at two different but close frequencies is shown. The frequencies have been chosen a posteriori of obtaining the waveguide simulator metamorphic material experimental results, in order to better explain the results on the next section.

It is interesting to notice the evolution versus bias voltage of the real part of the impedance, since this determines the metamorphic behavior of the loop. The lowest impedance achieved is of -15 Ω, while the highest is around 100 Ω. Despite not being able to obtain very high negative values, the negative resistance indicates that we can potentially obtain amplification out of the metamorphic material. Moreover, we should be able to set the material in the AMC, PEC and perfect absorption points.

Another point of interest is the evolution of the reactance vs. bias voltage. As shown in figure 4.9, the reactance impacts the degree of metamorphism that one can obtain. If the reactance of the load device compensates that of the remaining circuit, such that the overall reactance is null, we can then obtain perfect amplification and perfect absorption. However, as the mismatch in reactance increases, the amplification decreases and the absorption is reduced.

For the tunnel diode, the reactance is not constant versus biasing, which may challenge our ability to achieve all the different metamorphic states at a single frequency. However, since reactance is a function of frequency, there can be different frequencies where the overall reactance of the tunnel diode plus the overall remaining circuit compensates, and we can therefore achieve the different metamorphic states in different frequencies.
Fig. 4.9. Tunnel diode measured impedance at two different frequencies.

4.2.3. Experimental Results

The top and bottom views of the fabricated prototype inside the waveguide are shown in figure 4.10. In order to reduce the impact of the vias for the DC path, a couple of 8.2 nH inductors have been placed on the top of the loop and two chokes have been placed below the ground plane.
Fig. 4.10. Top and bottom views of metamorphic material inside waveguide.

Using the measurement setup shown in figure 4.11, with the waveguide simulator and biasing network in place, the metamorphic material reflection coefficient versus bias voltage is measured. Results are shown in figure 4.12. It can be seen that at 2.89 GHz the material goes through a region where the impinging plane wave is amplified. This region corresponds to the region where the tunnel diode resistance is negative. Also, one may notice the moderate
amplification of the signal compared to theoretical results. This is due to the fact that tunnel
diode reactance does not fully compensate the loop reactance.

---

![Fig. 4.11. Metamorphic material measurement setup.](image)

There are also regions where the incoming wave is slightly absorbed. Again, the reason why
we did not obtain perfect absorption is due to the tunnel diode reactance. In the same figure,
we have plot the reflection coefficient obtained at 2.755 GHz. In this case, the tunnel diode
reactance fully compensates that of the remaining circuit and the material behaves like a
perfect absorber.
Finally, there are regions in where the wave is totally reflected. In this case the material behaves like a PEC or PMC.

In figure 4.13, we have marked over the theoretical reflection curve of the metamorphic material the resistance regions that we are able to sweep at both frequencies. From this plot and plot 4.12 it is evident that a metamorphic material that undergoes the PMC, PEC, Perfect Amplification and Perfect Absorption states is physically possible. It is now a matter of controlling the reactance of the circuit to obtain all the metamorphic states at a single frequency.
Fig. 4.13. Metamorphic material theoretical reflection coefficient vs. load resistance swept in measurements.

4.3. Applications

Metamorphic materials, and especially fully tunable ones like the class of metamorphic materials presented in this chapter, open up a wealth of new applications in antenna design. They have the potential to revolutionize the field of tunable and beamforming antennas, where traditionally phase shifters, switches, or varactors have been used to steer the antenna radiation patterns or achieve different modes of operation.

If the antenna ground plane is substituted by a layer of fully tunable metamorphic material, then, the radiation can be enhanced, suppressed, modulated, steered...to the needs of the user, which gives a lot more flexibility than current systems.
Moreover, metamorphic surfaces could be used to cover diverse object surfaces to obtain invisibility, or relay and amplification objects.

4.4. Conclusions

In this chapter a new class of fully metamorphic materials that can progressively be tuned from one metamorphic state to the others has been presented. This class of metamorphic surfaces can be obtained by loading an array of planar loops over a ground plane with a variable resistance. This new material can act as a PEC surface, PMC surface, Perfect Absorber or Perfect amplifier.

A mathematical model for the response of the metamorphic material to an incident TE and TM polarized plane wave has been obtained for the case of infinitesimal loops. Moreover, a prototype with practical dimensions has been fabricated and tested. To obtain a tunable resistance with positive and negative values a tunnel diode has been used as the loading element for the loops.

For the first time, we have shown that the proposed structure can undergo all the metamorphic states as the tunnel diode biasing voltage is swept. This finding represents a great breakthrough in achieving metamorphic and tunable antennas, materials and surfaces, and has a lot of potential applications.
Chapter 5

BEAMFORMING MM-WAVE GRID

ARRAY ANTENNAS

At millimeter-wave frequencies the wavelength is small enough to enable the integration of highly efficient antennas into the radio package, thus reducing size and cost. This benefit however, does not come for free. Mm-wave signals suffer from high path losses (for instance, there are 21 dB more path losses at 60 GHz than at 5 GHz), which limits the communication range to a few meters. Moreover, signals are highly attenuated by most construction materials and the human body. Therefore, to achieve reliable high data rate communication links we need to use directive antenna arrays with beamforming capabilities to exploit the best communication path possible.

Multiple Antennas in Package (AiP) designs have been reported [7-9, 48-49]. However, as they use only one antenna element [7, 8], the gain is not high enough to compensate the extra path loss of 60 GHz wireless communication links with respect to the 2.5 GHz and 5 GHz cases. Microstrip patch array antennas that provide the desired gain and allow beam scanning have also been proposed for 60 GHz applications [9, 48, 49]. These arrays require complicated feeding networks that introduce loss and they exhibit undesirable bandwidth limitations. Moreover, phase shifters are required to scan the beam, thus increasing the overall system cost.
In [50] a grid array antenna with a gain above 10 dB in the whole 60 GHz band was presented. This antenna structure is advantageous versus conventional arrays because the excitation is simplified to just one single point and is very broadband and directive. On the other hand, the application is limited to LOS environments and emitter and receiver have to be perfectly aligned to ensure a successful communication. Moreover, it was fabricated using an expensive low-temperature co-fired ceramic (LTCC) substrate, which could potentially delay market penetration.

In this chapter we present the design of different grid array antennas that overcome the aforementioned drawbacks. First we analyze a broadside radiating antenna and look for an optimal design in terms of the gain/area product. Then, we take advantage of the different radiation mechanisms of the grid antenna structure to obtain multiple selectable beams. All the antennas, fabricated on a low cost multi-layer substrate, satisfy the requirements of low cost and small size for mass production and are good candidates for different 60 GHz communication systems. The theory of operation, analytical model and experimental results are presented.

5.1. Grid Array Antenna Radiation Principle

The grid array antenna is composed of interconnected rectangular loops etched on a dielectric substrate backed by a metallic ground plane as shown in figure 5.1. This antenna concept was first introduced by Dr. Kraus [2] in 1964 but it was not until 1980 when Conti [51] reported its microstrip implementation.
The radiation properties of the structure are determined by the grid geometry. The loop dimensions and feed point determine the use of the structure as a traveling wave device or as a resonant radiating structure. Assuming a single feeding point and microstrip meshes with sides of a wavelength (L=λ) by half a wavelength (W=λ/2), the current on each short side is in phase. Using array theory and considering each short microstrip segment as a half wavelength radiating element, the radiation will be broadside and polarized parallel to the short sides of the mesh. The long sides of the loop support a full wavelength current element and therefore the radiation on the broadside direction is null. Moreover, the two long sides of the loop form a pair of anti-phase radiating elements with an array factor containing nulls on the antenna plane and the broadside direction. Therefore, the long segments of a single loop produce four distinct lobes at 45° from broadside in the x and y directions, with levels reduced about 6 dB relative to the co-polarized component. As more loops are added to the grid the magnitude of the cross-polarized peaks is reduced. The loop impedance and feed point determine the
amplitude distribution on the array and can be used to control the sidelobe levels and the antenna bandwidth.

If the antenna is fed from one end with \( W < \frac{\lambda}{2} \), the current on the short sides follows a phase progression and the maximum radiation occurs in the backward direction. On the other hand, if \( W > \frac{\lambda}{2} \), the radiation occurs in a forward direction. Figure 5.2 shows various radiation patterns that can be obtained by changing the grid antenna short sides’ length. The antenna can then be classified as a travelling wave antenna.

![Figure 5.2](image.png)

**Fig. 5.2.** E-plane grid antenna radiation patterns at 60GHz for \( L = 3.03 \) mm (microstrip effective wavelength) and \( W = 1.3 \) mm, \( W = 1.5 \) mm, \( W = 1.7 \) mm and \( W = 1.9 \) mm.

Furthermore, for a given geometry, if the input frequency is swept, the main beam can be scanned. In figure 5.3 the radiation patterns obtained at different frequencies for an edge fed grid antenna with \( L = 3.03 \) mm and \( W = 1.5 \) mm are shown.

![Figure 5.3](image.png)

**Fig. 5.3.** E-plane grid antenna radiation patterns at 57, 60 and 64 GHz for \( L = 3.03 \) mm and \( W = 1.5 \) mm.
5.2. Broadside Grid Antenna on Low Cost Packaging Substrate

5.2.1. Substrate Choice

To satisfy the requirements of low cost and chip packaging integration, we chose a multilayer substrate packaging process that uses the substrate CCL-HL832MG (CCL) from Mitsubishi Gas Chemical Co., with a permittivity of $\varepsilon_r=3.54$ and $\tan\delta=0.0091$ according to the provider. The package stack-up is shown in figure 5.4. The first and last layers are 125 µm thick and the core stack is 600 µm. With respect to the Low Temperature Co-fired Ceramics (LTCC) design presented in [50], this process is cheaper.

![Grid antenna substrate stackup](image)

Fig. 5.4. Grid antenna substrate stackup.
5.2.2. Antenna Design

Since we are not interested in the frequency scanning capability, assuming loops with sides of approximately one wavelength by a half-wavelength in the dielectric, the resulting radiation would be broadside to the grid and polarized parallel to the short sides of the loops. However, one of the main challenges when designing the grid array is keeping the maximum radiation in the desired direction for a wide frequency range, as required in our case. As frequency deviates from the designed center frequency, the short sides of the loop are not anymore half-wavelength and the beam starts to steer. On centering the feed in the grid, the frequency range in which the maximum radiation occurs in the broadside direction is spanned but as frequency is further increased or decreased the radiation pattern presents a split beam.

The designed grid array antenna with dimensions is presented in figure 5.5. The package has a size of 12x14 mm and is approximately 1 mm thick. The grid array, consisting of 18 microstrip loops of length \( l = 3.1 \text{ mm} \) and width \( w = 1.7 \text{ mm} \) is printed on the first metallic layer. The 3rd metallization layer provides the antenna ground plane, as well as the ground for the coplanar waveguide (CPW) feed line, allocated on the 4th metallic layer.

The CPW line, designed to be 50 \( \Omega \) and be probed with 200 \( \mu \text{m} \) pitch-to-pitch probes, has a center line width of 120 \( \mu \text{m} \) and a gap of 35 \( \mu \text{m} \) and is connected to a via of 125 \( \mu \text{m} \) diameter which brings the signal to layer 3. Another via of diameter 150 \( \mu \text{m} \) brings the signal from metal 3 to metal 2 and to complete the feeding path, a third via interconnects the last one to the antenna. A series of extra vias interconnecting the CPW ground plane with the antenna ground were added for matching purposes.
Fig. 5.5. Views and dimensions of the designed grid array antenna: a) top view; b) feeding detail; c) side views.

A picture of the fabricated antenna prototype is shown in figure 5.6.
5.2.3. Measurement and Simulation Results

In figure 5.7, the simulated and measured return losses are compared. Measurements were performed at our facilities at the University of California, Irvine, where we designed and built a probe-based mm-Wave band anechoic chamber. It can be observed that the measured curve is shifted in frequency with respect to the simulated one though they have similar features. This effect was identified as being due to the use of an inaccurate substrate permittivity value in the simulations. Using the covered transmission line complex permittivity measurement method described in chapter 3, a substrate sample was measured at 60 GHz, obtaining a permittivity of $\varepsilon_r=3.36$ and $\tan\delta=0.012$. The grid array was then re-simulated using these values and the new return losses obtained. In this case both measured and simulated curves agree very well, with some differences observed at the lower and higher frequencies due to the permittivity change versus frequency of the substrate and other fabrication tolerances, which were not been taken into account in the simulations. The antenna is well adapted (VSWR<2) from 52 GHz to 65 GHz, thus covering the desired 57-64 GHz frequency band.
Figure 5.7 shows the simulated and measured grid array return losses vs. frequency.

Figure 5.8 shows the simulated and measured co-polarized radiation patterns in the XZ and YZ cuts at 57, 60 and 64 GHz. It can be seen that the maximum radiation is tilted approximately 3° from broadside in the YZ cut, mainly due to the asymmetry of the antenna on this plane. Comparing the simulated radiation patterns for both substrate permittivities, it is observed that the sidelobes are more significant at 57 GHz for $\varepsilon_r=3.36$ thus the antenna gain will be lower than for $\varepsilon_r=3.54$. At 64 GHz however, the antenna with a substrate permittivity of $\varepsilon_r=3.54$ has higher secondary lobes and thus will have less gain. Both antennas show approximately the same sidelobes level at 60 GHz. The simulated and measured radiation patterns agree very well, except at 57 GHz, where higher secondary lobes have been measured.

The antenna has small cross-polarization radiation components. Simulation and measurements have shown that the cross-polarized components are approximately 20 dB lower than the co-polarized ones.
Fig. 5.8. Grid array antenna simulated with substrate permittivity $\varepsilon_r=3.36$ and $\varepsilon_r=3.54$, and measured co-pol radiation patterns at 57, 60 and 64GHz.
In figure 8.9 the simulated broadside gain for both permittivities (manufacturer provided and measured) is presented. A shift of about 2 GHz between both curves is observed, corresponding to the same shift detected when measuring the return losses.

![Simulated broadside gain vs. frequency.](image)

5.3. Gain vs. Area Considerations

The grid array antenna can be classified as a travelling wave antenna given that the energy is radiated as the fields travel through the wire mesh. Since leakage occurs over the whole periodic structure, the whole mesh constitutes the antenna’s effective aperture. However, if the leakage rate is so great that the power has effectively leaked away before reaching the furthermost opposite meshes to the feed point, the antenna’s effective aperture is smaller than the physical antenna size.

Given a sufficiently extended grid area, a large attenuation constant would imply a short effective aperture, so that the radiation pattern would have a large beamwidth. Conversely, a low attenuation constant would result in a long effective aperture and a narrow beamwidth.
pattern. Therefore, the antenna substrate will play an important role in the achievable antenna beamwidth and gain.

With the previous considerations in mind, once a choice has been made for the antenna substrate, a valid requirement is to determine the optimal grid size. A figure of merit to look into is the Gain/Area ratio. Since power is radiated continuously along the wire mesh, the aperture field has an exponential decay. As the number of meshes increases so does the antenna gain. However, given the tapered field distribution, the meshes furthest from the feed point contribute with little radiated power. This results in a larger physical antenna area needed to increase the gain by the same amount. Therefore, there exists a tradeoff between gain and area.

In figure 5.10 the Gain/Area ratio for different size grid antennas on the substrate previously described is given. It can be observed that the optimal designs for antennas with a width of 3 loops are the 3 and 7 loops in length. The choice of one design or another will then depend on the desired beam characteristics and physical available area. If we choose a width of 2 loops the optimal is obtained with 9 loops in length. It is here important to notice that the Gain/Area figure of merit graph is a function of the substrate properties. Therefore, the antenna designer will need to perform a similar study to find the optimal for his/her application.

Given the decay in the Gain/Area figure of merit, if sufficient area is available, one may consider other options rather than increasing the mesh size. For instance, it may be a better option to array different optimal size grid arrays and feed them through a corporate feed or any other technique. Figure 5.11 illustrates a possible configuration with 4 grid antennas that could potentially give a higher gain than an equivalent size single grid antenna. If the higher
gain is achieved or not may again depend on the attenuation constant of the fields as they travel through the mesh and ultimately on the substrate.

Fig. 5.10. Gain/Area figure of merit for different sizes of the grid in number of loops.

Fig. 5.11. Array of 4 equal size optimal grid array antennas.
5.4. Beam Selection Grid Antenna

There have been numerous studies to enhance and optimize the radiation characteristics of the grid antenna [51-54] and extensions have been proposed to obtain circular polarization [55-56], but none of these have exploited the traveling wave nature of the antenna to obtain different beams by feeding the antenna at different points.

If the antenna is fed from one end and the loop dimensions are \( L = \lambda \) and \( W < \lambda/2 \), we have previously argued that the maximum radiation will be on a backward direction, let us say at an angle \( \alpha \) from broadside. Therefore, intuitively, if we have a second feeding point located at the opposite side of the antenna, we will obtain a second beam pointing at \(-\alpha\).

In this section we exploit both the traveling wave and resonant radiation nature of the grid antenna to obtain a beam selection antenna. Feeding the antenna on both opposite sides and at the center and applying different phases to each excitation, the beam can be scanned to 5 different directions.

5.4.1. Substrate Choice

To satisfy the requirements of low cost and chip packaging integration, an organic substrate, with a permittivity (as provided by the manufacturer) of \( \varepsilon_r = 4.16 \) and \( \tan\delta = 0.012 \), was used. The same package stack-up as shown in figure 5.4 was used, but with slightly thinner layers. The outer substrate layers are 65 \( \mu \)m thick meanwhile the inner stack is 435 \( \mu \)m in total.
5.4.2. Antenna Design

The designed beam selection grid antenna consists of 18 microstrip loops of length $L=3.03$ mm, width $W=1.4$ mm and microstrip line width of $50 \, \mu$m printed on the first metallic layer (M1). M3 serves as the antenna ground plane as well as the ground for the coplanar waveguide (CPW) feed lines, allocated on M4. The antenna is fed at 3 points, located at opposite edges and at the center of the grid, by vias of $100 \, \mu$m diameter which interconnect the CPW lines with the antenna. The CPW lines, designed to be $50 \, \Omega$ and probed with $200 \, \mu$m pitch-to-pitch probes, have a center line width of $70 \, \mu$m and a gap of $40 \, \mu$m. A series of extra vias interconnecting the CPW ground plane with the antenna ground were added for matching purposes. The design parameters were optimized to obtain a VSWR<$2$ from 57 to 64 GHz.

Figure 5.12 presents the antenna design and figure 5.13 shows detailed views of the feeding. It is important to notice the simplicity of the overall design, and specifically of the feeds. The package has a size of $11.75 \times 15$ mm and is approximately $0.5$ mm thick.

Fig. 5.12. 60 GHz band beam selection grid antenna design.
5.4.3. Simulation Results

Feeding properly each port with 0°, 90°, 180° or 270° phase shifts (corresponding to different combinations of the in-phase and quadrature outputs which a 60 GHz chip can provide), up to 5 different beams can be obtained. Table 5.1 summarizes the different beam characteristics obtained by the given feeding configuration. Figure 5.14 shows the simulated gain vs. frequency for beam the 5 different beams. The simulated in-band gain difference is below 3dB for all beams.

Table 5.1. Grid Antenna Beams and Feed Configuration
5.4.4. Measurement Results

Given that at present a 60GHz chip capable of providing the required signal and phase at each antenna feed point is not available, a series of test prototypes with a single feed as shown in figure 5.15 have been designed. The relative length among the three lines is controlled through the meander lines so the desired relative phase between feed points is achieved.
60 GHz band beam selection grid antenna dummy prototype.

The feed network consists of a 50 Ω CPW line whose impedance is transformed to approximately 20 Ω by using a quarter wavelength tapered transition. The line then splits into three 50 Ω CPW lines that bring the signal to each feeding point. The relative length among the three lines is controlled through the meander lines so the desired relative phase between feed points is achieved.

It has to be noticed that given the extra length of the side feeding paths, the signal will suffer higher attenuation at feed points F1 and F2 with respect to F3. Therefore, the current distribution on the antenna will be slightly different than that of the antenna in figure 5.12 and the radiation patterns will have some distinct features. However, simulations show that the main beam direction remains unchanged.

A picture of one of the fabricated prototypes is presented in figure 5.16.
Table 5.2 presents a summary of the maximum measured gain for each beam. It can be observed that there is a significant deviation between the predicted simulated gain and the measured value. This can be attributed to the fact that the measured final permittivity of the substrate was $\varepsilon_r = 3.65$ and $\tan\delta = 0.02$ and therefore the design dimensions were not optimal and in addition we had increased losses. Measurements have also shown that the overall array performance has shifted up in frequency due to the permittivity change. The measured VSWR is lower than 2 in the range from 58GHz to 66GHz.

<table>
<thead>
<tr>
<th>Beam #</th>
<th>1</th>
<th>2</th>
<th>3</th>
<th>4</th>
<th>5</th>
</tr>
</thead>
<tbody>
<tr>
<td>Max. Meas. Gain (dB)</td>
<td>11.3</td>
<td>12</td>
<td>10.3</td>
<td>8.7</td>
<td>9.2</td>
</tr>
</tbody>
</table>

Table 5.2. Measured Grid Antenna Beams Gain

Figure 5.17 shows the different measured beams obtained at 62 GHz. In spite of the resulting poor design due to the permittivity change, the 5 beams can be distinguished.
5.5. Conclusions

The feasibility of integrating a 60 GHz band grid array antenna with gain higher than 11dB on a low-cost packaging substrate has been demonstrated. Measurements show a bandwidth from 52 GHz to 65 GHz (VSWR<2) and broadside radiation patterns from 57 to 64 GHz. Simulated and measured return losses however, present approximately 2 GHz frequency deviation. The difference between design substrate permittivity and the real substrate permittivity was identified as the cause of such frequency shift. The Gain/Area figure of merit for different designs has also been evaluated.

A variant of the grid array antenna that allows beam selection has been proposed for low-cost 60 GHz devices. Simulations show that 30° beam scanning with high gain over the entire 60 GHz band can be achieved. The antenna has been fabricated in an organic low-cost substrate. The substrate uncontrolled permittivity change explains the poorer performance of the prototype. However, the beam scanning capability has been confirmed experimentally.
Chapter 6

APPLICATIONS TO INDOOR POSITIONING SYSTEMS

High accuracy indoor wireless localization systems are receiving a lot of interest as a wealth of applications such as indoor autonomous navigation, object and body motion tracking for gaming, asset management, etc. are gaining momentum. Moreover, knowing the location of people in indoor environments can be very advantageous for certain markets. For instance, in the retail sector, the whereabouts of shoppers in shopping malls is of great importance in determining the clients’ responses to marketing strategies, understanding consumer behavior and in general gathering statistical data that could help boost the revenues. Recently, a wealth of companies that track individuals inside the premises of shopping centers, retail stores, sport arenas and museums have emerged. Their solutions range from WI-FI triangulation to proximity sensing or using cell-phone signals to find the customers’ position. However, the accuracy of all these systems is in the range of a few meters, which is not sufficient in some cases.

In academia, radio frequency (RF) based indoor positioning systems have been the focus of numerous studies [57-60]. In [57] a summary of different positioning systems is given. Significant attention has been devoted to the system architecture design in order to obtain higher positioning accuracies. Proposed competing methods include frequency modulated continuous wave (FMCW), impulse based ultra-wideband (UWB) and carrier-based UWB
technologies. However, no much attention has been given to the receiving antenna system configuration. The vast majority of positioning systems use omnidirectional receiving antennas for simplicity. Nevertheless, the antenna could play an important role in filtering some of the multipath in rich scattering environments and therefore contribute to increase the positioning accuracy. Figure 6.1 illustrates the effect of using directional antennas vs. omnidirectional antennas in rich scattering environments.

![Diagram of scattering surfaces with omnidirectional and directional antennas](image)

**Fig. 6.1.** Omnidirectional vs. directional antenna multipath filtering.

In the present work, we study the impact of using phased-array antennas on the positioning and tracking precision of multiple targets in a realistic indoor environment. The positioning architecture is a 60 GHz carrier-based ultra-wide band (UWB) system. The 60 GHz band has been chosen because it allows for the design of very compact high-gain phased-array antennas. Also, the positioning accuracy is proportional to the signals bandwidth in a CDMA (Code Division Multiple Access) system. Therefore, the available bandwidth at the mm-Wave band allows for greater positioning capabilities. Figure 6.2 illustrates the bandwidth-resolution relation in a CDMA system, which is given by equation (1):
\[ R_s = \frac{c}{BW \sqrt{T \frac{C}{N_0}}} \]  

(1)

where \( c \) is the speed of light, \( BW \) the code bandwidth, \( T \) the measurement duration and \( C/No \) the carrier to noise density ratio (68 dB-Hz @ threshold).

Moreover, future portable electronic devices are expected to integrate 60 GHz technology, making the proposed tracking system a viable option for commercialization.

### 6.2. Mm-Wave Positioning System Description

#### 6.2.1. Architecture of the 3D Positioning system

In figure 6.3 the system architecture of a 3D high resolution positioning system is shown. The preferred configuration includes at least four stationary RF receivers defining sensors which are placed at known locations around the area of interest. Their positions should minimize the
geometric dilution of precision (GDOP) [61]. A stationary RF transmitter defining a reference marker is placed on a known location inside the tracking zone, which should be at least within the reception range of all receivers. A number of other markers are placed on objects or carried by people to be tracked or positioned. Each of the markers transmits a unique identification code that is identified at the receivers. The processing unit computes each marker position using the combined information from each receiver.

Fig. 6.3. 3D high resolution positioning system architecture.

For the purpose of the current study, the markers’ channel access method is similar to that of an asynchronous code division multiple access (CDMA) system. Each marker has an m-sequence pseudo-noise (PN) code generator to create its unique identification code $c(t)$. Each code has a length of $N_c$ chips and bandwidth $BW$ and is statistically uncorrelated with other codes. The code is then modulated and up-converted to the carrier frequency of 60 GHz. Figure 6.4 shows a block diagram of the structure of the markers (60 GHz Transmitters) and receivers.
The receivers down-convert, demodulate and sample the signals. Then, they use a bank of correlators with the same m-sequence code generators and variable time delay $\tau$ to compute the pseudo time of arrival (Pseudo-TOA) for each of the markers, given by the peak of the correlation function between the delayed code $c(t-\tau)$ and the received code $c_r(t-\Delta\tau)$ [60]. The processing unit then combines the information from each receiver and applies a triangulation technique to determine the position of each marker.

### 6.2.2. Positioning Algorithm

In a time of arrival (ToA) positioning approach, the receivers and transmitters need to have precisely synchronized clocks and label the packets with timestamps in order for the time of flight to be discerned at receiver nodes. In our approach, and emulating real scenarios, since the various markers and sensor receivers do not usually share a common time reference, the
extracted ToA measurements are called Pseudo-ToA and a direct triangulation algorithm cannot be applied to find the markers position. Instead, a double time difference of arrival (DTDOA) triangulation technique [62, 63] is used.

Pseudo-ToA measurements are converted to pseudo-range measurements by considering that rays travel at light speed. They are pseudo-range, as opposed to range, because they contain a time term. Let’s consider that A designates a marker, R a reference marker and i refers to an arbitrary receiver. At the n\textsuperscript{th} measurement time, R is located at \((x_R, y_R, z_R)\) with clock \(T_R(n)\) and marker A is located at \(r_A(n)=(x_A(n), y_A(n), z_A(n))\) with clock \(T_A(n)\). The sensor i is located at known position \(s_i=(x_{si}, y_{si}, z_{si})\) with clock \(T_i(n)\). Pseudo-ranges at sensors are given by:

\[
P^R_i(n) = \sqrt{(x_R - x_{si})^2 + (y_R - y_{si})^2 + (z_R - z_{si})^2} - c(T_R(n) - T_i(n)) = |r_R - s_i| - c(T_R(n) - T_i(n))
\]

\[
P^A_i(n) = \sqrt{(x_A(n) - x_{si})^2 + (y_A(n) - y_{si})^2 + (z_A(n) - z_{si})^2} - c(T_A(n) - T_i(n)) = |r_A(n) - s_i| - c(T_A(n) - T_i(n))
\]

(2)

To eliminate the sensor time term, pseudo-range measurements are subtracted between the markers and the reference markers to obtain single differences.

\[
P^A_i(n) - P^R_i(n) = |r_A(n) - s_i| - |r_R - s_i| - c(T_A(n) - T_R(n))
\]

(3)

Subtracting between the single differences of two different sensors, the marker related clock terms can be eliminated:

\[
\delta P^A_{ij}(n) = P^A_i(n) - P^R_i(n) - P^A_j(n) - P^R_j(n) = |r_A(n) - s_i| - |r_R - s_i| - |r_A(n) - s_j| + |r_R - s_j|
\]

(4)
Finally, the double differences are processed to determine the markers positions relative to the reference marker position using a typical triangulation algorithm. Since we have 3 unknowns corresponding to the marker $A$ position coordinates, measurements from 4 sensors are required to form the 3 independent double difference equations.

The least squares solution for marker $A$ position relative to the reference tag is given by

$$r_A(n) = r_0(n) + \delta r_A,$$

where $r_0(n)$ is an approximate solution for the marker $A$ position vector and

$$\delta r_A = (H_n^T H_n)^{-1} H_n^T \Delta z,$$

with

$$\Delta z = \begin{bmatrix}
\delta P_{01}^{AR}(n) - |r_0(n) - s_o| + |r_R - s_o| + |r_0(n) - s_1| - |r_R - s_1| \\
\delta P_{12}^{AR}(n) - |r_0(n) - s_1| + |r_R - s_1| + |r_0(n) - s_2| - |r_R - s_2| \\
\vdots \\
\delta P_{M,M+1}^{AR}(n) - |r_0(n) - s_M| + |r_R - s_M| + |r_0(n) - s_{M+1}| - |r_R - s_{M+1}|
\end{bmatrix},$$

and

$$H_n = \begin{bmatrix}
ax_{01}^{A} & ay_{01}^{A} & az_{01}^{A} \\
ax_{12}^{A} & ay_{12}^{A} & az_{12}^{A} \\
\vdots & \vdots & \vdots \\
ax_{M,M+1}^{A} & ay_{M,M+1}^{A} & az_{M,M+1}^{A}
\end{bmatrix},$$

$$ax_{ij}^{A} = \frac{x_0(n) - x_{si}}{|r_0(n) - s_i|} - \frac{x_0(n) - x_{sj}}{|r_0(n) - s_j|},$$

$$ay_{ij}^{A} = \frac{y_0(n) - y_{si}}{|r_0(n) - s_i|} - \frac{y_0(n) - y_{sj}}{|r_0(n) - s_j|},$$

$$az_{ij}^{A} = \frac{z_0(n) - z_{si}}{|r_0(n) - s_i|} - \frac{z_0(n) - z_{sj}}{|r_0(n) - s_j|}.$$
For improved accuracy, the equations are iterated as follows:

- Compute the measurement vector $\delta z$.
- Compute the linearized observation matrix $H_n$.
- Compute the error vector $\delta r$.
- Update the marker A estimated position vector $r_A$.
- Repeat steps using the updated position vector as the new estimated position vector, until computed error is sufficiently small.

In [63] the mathematical formulation of the process is given in detail. Figure 6.5 presents a block schema of the DTDOA triangulation method steps.

![Block diagram](image)

**Fig. 6.5.** Double time difference of arrival triangulation method block diagram.

### 6.2.3. Simulator Description

The system previously described has been programmed in C# in order to conduct a series of simulations. In figure 6.6 the schema of the different blocks configuring the program is presented. The scenario geometrical description, antenna radiation patterns and carrier frequency are used as inputs to a ray-tracing tool which computes the channel impulse response. This, along with the unique codes identifying each marker and the signal bandwidth
$BW$ and sampling frequency are used to generate the signals received at each sensor receiver. Finally, pseudo-ToAs for each marker and receiver pairs are computed and the positioning algorithm described in the previous section is applied to obtain the estimated position of each marker.

![Block schematic of the 3-D Positioning system simulator](image)

**Fig. 6.6.** 3-D Positioning system simulator block schematic.

Figure 6.7 shows the program’s user interface and the following paragraphs offer a more detailed description of the inputs and each simulation block along with its outputs.

![User interface of the 3D Positioning simulator](image)

**Fig. 6.7.** 3D Positioning simulator user interface.
INPUTS

Geometry description

This text file describes the position and nature of all scatterers. Each scatterer is defined by the following quadric equation:

\[ ax^2 + by^2 + cz^2 + 2fyz + 2gzx + 2bxy + 2px + 2qy + 2rz + d = 0. \quad (10) \]

Examples of quadric surfaces include the cone, cylinder, ellipsoid, paraboloid, sphere, plane, etc. A combination of multiple quadric surfaces defines the topology of a room and all the scattering objects inside it. Reflection and transmission coefficients for each surface are also considered. The following code shows the description of an infinite PEC wall:

```
Plane
1 0 0 -10 # => x-10=0
1 #=> Reflection Coefficient=1
0 #=> Transmission Coefficient=0
```

Antennas Description

These text files contain the description of each marker, the reference marker and all the sensor receivers. The antennas are described by their location in space, their radiation pattern and the polarization. The following code shows the description of an antenna:

```
%% Antenna Description Form.
Position Reference Antenna (x,y,z)
4 4 1
Delta theta and delta phi in deg.
90 0.1
Radiation pattern (theta coordinate (deg), phi coordinate(deg), power (dB))
90 0 10
....
```
Antennas ID Codes

This text file contains a set of Gold Code sequences identifying each of the markers (or transmitting antennas). In this case, the codes have been generated by using Matlab. With the use of 2 preferred m-sequence generators of degree $r$ and a fixed non-zero seed in the 1st generator, $2^r$ Gold codes are obtained by changing the seed of the 2nd generator. $2^r-1$ Gold codes are available to identify a maximum of $2^r-1$ markers. In figure 6.8 we show the schematic of a gold code sequence generator.

SIMULATION SCENARIO

The information from the input files is then combined to obtain a spatial representation of the scenario under surveillance. In figure 6.9 a 2D scenario with 4 receivers, 1 marker and a reference marker in free space is shown for reference.
SIMULATION MODULES

Ray Tracing

This simulation module generates a set of initial rays originating from each of the markers. Each ray is represented by equation (11) given below

\[ R_{inc}(t) = R_o + R_d = (x_o, y_o, z_o) + (x_d, y_d, z_d)t. \]  

Each ray has initial power proportional to the markers radiated power in the direction of the ray, which is given by the antenna radiation defined in the description text files. Then, the module determines whether there is an intersection between each ray and every object in the scenario by substituting the ray equation (11) into the quadric equation (10), to obtain:

\[ A_q t^2 + B_q t + C_q = 0 , \]  

with
\[ A_q = A x_o^2 + B y_o^2 + C z_o^2 + D x_o y_o + E x_o z_o + F y_o z_o, \]
\[ B_q = 2 A x_o x_d + 2 B y_o y_d + 2 C z_o z_d + D (x_o y_d + y_o x_d) + E x_o z_d + F (y_o z_d + y_d z_o) + G x_d + H y_d + I z_d \]

and

\[ C_q = A x_o^2 + B y_o^2 + C z_o^2 + D x_o y_o + E x_o z_o + F y_o z_o + G x_o + H y_o + I z_o + J. \] (13)

From these equations we proceed to calculate \( t \). For \( A_q = 0 \)

\[ t = -\frac{C_q}{B_q} \] (14)

When \( B_q^2 - 4A_qC_q < 0 \) there is no intersection. Otherwise, \( t_o \) is determined to be

\[ t_o = \frac{-B_q - (B_q^2 - 4A_qC_q)}{2A_q} \] (15)

If \( t_o > 0 \), we are done. If not, we proceed to find \( t_1 \) as:

\[ t_1 = \frac{-B_q + (B_q^2 - 4A_qC_q)}{2A_q} \] (16)

Once \( t \) is found, we can calculate the intersection point \( R_i \) and the normal \( N \) to the surface at \( R_i \). From these, we can obtain the reflected ray direction:

\[ R_{reflected} = 2N(R_{inc}N) - R_{inc}. \] (17)

The reflected rays power is adjusted according to the Friis path loss formula to account for the power loss as it travelled from the origin to the intersection point. The reflectivity of the surface is also taken into account to adjust the ray power according to the properties of the object, as defined in the geometry description document.
The reflected rays then become part of the set of rays to analyze and go under the same treatment as the initial rays. A ray becomes part of the solution when it is received at a sensor. A ray stops to be considered when its power is below the system sensitivity or it does not reflect on any object. Figure 6.10 shows an example of the ray tracing solution for a simple scenario.

![Ray Tracing Solution Diagram](image)

**Fig. 6.10.** Example of the ray tracing solution for a simple scenario.

The ray tracer output is a file containing each ray captured by the receiver. For each ray, the following information is given:

- Receiver identifier
- Ray travelled distance from the origin
- Ray power
- Last ray reflection point coordinates
- Transmitting antenna identifier

Figure 6.11 shows an example of the output file.
Fig. 6.11. Ray tracer output file example.

The program also allows for visualization of the ray tracing solution. Figure 6.12 shows the solution rays for the free space scenario presented in figure 6.9. It can be seen that since there are no reflecting objects, only the line of sight rays are part of the solution.

Fig. 6.12. Ray Tracer graphical solution for the scenario of figure 6.9.

**Channel Impulse Response**

From the Ray Tracer output the Channel Impulse Response for each receiver can easily be calculated. This can be done by transforming the rays travel distance to a time value. If we
then group the ray times by the receiver, we can obtain the channel impulse response for each receiver. Figure 6.13 shows the channel impulse response for the case of figure 6.9.

![Channel Impulse Response Diagram](image)

**Fig. 6.13.** Channel impulse response as seen by each receiver for the scenario of figure 6.9.

**Receiver Signal Generation**

Given the channel impulse response, the Gold Code sequences identifying all the markers and the signal bandwidth, the baseband signal received at each sensor can then be generated. This is done by properly summing up in the time domain the signals emitted by each marker. Figure 6.14 presents the received signal at receiver R3 from figure 6.9.
Pseudo-ToA Calculus

Up to now we have simulated the transmitters and the channel and we have recreated the signals that the receivers would sense. Now, we need to process the received signals appropriately to finally retrieve the position of our markers.

Given the baseband signals received at each sensor, the objective is to estimate the channel impulse response. To do so, the received signals are sampled and convoluted with the known Gold Code sequences of each marker. The convolution peaks are taken as the Pseudo-ToA, which will later be used in the triangulation module.

Figure 6.15 shows the convolution of the sampled received signal by R1 and the code of marker 1.
The estimated channel impulse response for the scenario of figure 6.9 is given in figure 6.16. It can be observed that since we just had the line of sight signals and low interference between the two markers, the estimated channel impulse response is very close to the simulated one. This means that the position of the marker can be retrieved with very low error.

Fig. 6.16. Estimated channel impulse response as seen by each receiver of the scenario depicted in figure 6.9.
**Triangulation**

Given the channel impulse response, with the pseudo-ToA for each marker and receiver pair, we can now apply the triangulation algorithm described in section 6.2.2 to find the markers position.

### 6.3. Positioning Accuracy Analysis

#### 6.3.1. Omnidirectional vs. Beamforming System Accuracy

The purpose of this section is to determine whether the use of beamforming antennas at the receivers has an impact on the positioning accuracy in comparison to the case where omnidirectional antennas are used. Without loss of generality, a realistic 2D indoor scenario is considered. An 11 x 10 m² room, defined by walls of $\Gamma=0.5$ reflection coefficient is assumed. Two highly reflective ($\Gamma=0.9$) cylinders are also present in different corners of the scenario (see figure 6.17). This could well be the representation of an exposition room, gallery or conference room, where the cylinders account for furniture or a sculpture. Notice that only 3 receivers are needed for 2D positioning.
For the first case under consideration, both the markers and receiving sensor antennas are omnidirectional. Each marker emits a unique $N_c=511$ symbols code, with a bandwidth $BW=2$ GHz. The sampling frequency is 6 GHz. For the second case under consideration, the receiving sensor antennas are beamforming antennas. These antennas have a synthetic pattern of 25 deg. beamwidth with a gain of 10 dB and -4.8 dB gain elsewhere, so the total equivalent isotropically radiated power (EIRP) is equal to that of the omnidirectional antennas.

The algorithm to compute the position of the markers is the same in both cases. However, additional intelligence has been added for the BF case, so the receiver beams are scanned in the antennas plane. For each beam position configuration, pseudo-ToAs are calculated and the peak of the correlation value is stored. The triangulation is then performed with those pseudo-ToA values which had the highest correlation.
In table 6.1, statistical data about the positioning accuracy of both systems in the previously described scenario is given. The data has been obtained for 26 different cases, where 5 markers in each case were randomly distributed in the room space. The number of detections, 43.8% in the omnidirectional case vs. 49.2% in the BF case, refers to the number of markers for which the triangulation algorithm converged and a position could be calculated. The algorithm does not converge when some of the pseudo-ToAs are greatly erroneous, meaning that they differ a lot from the theoretical value. This can be due to two different factors, namely multipath or the near-far problem [8] inherent to asynchronous CDMA systems. Meanwhile we will tackle the near-far problem in the next paragraphs, we can conclude that the BF system increases the detection percentage by 5.4%, mainly due to the antenna filtering some of the multipath.

<table>
<thead>
<tr>
<th>Positioning Accuracy in Indoor Scenario</th>
<th>Omni. Antennas</th>
<th>BF Antennas</th>
</tr>
</thead>
<tbody>
<tr>
<td># Detections</td>
<td>43.8 %</td>
<td>49.2 %</td>
</tr>
<tr>
<td>&lt; 10 cm error detections</td>
<td>23 %</td>
<td>76.6 %</td>
</tr>
<tr>
<td>Mean Error</td>
<td>3.01 m</td>
<td>0.78 m</td>
</tr>
</tbody>
</table>

Table 6.1. Indoor Scenario positioning results summary.

Taking into account only the detected markers, the mean positioning error for the BF case is 0.78 m while for the omnidirectional case it increases to 3 m. Furthermore, if we consider the detected markers that have been positioned with an accuracy of 10 cm or better, for omnidirectional systems the percentage is 23% but it is 76.6% for the BF case. This represents a 53.6% improvement. Figure 6.18 provides a graphical representation of the absolute positioning error distribution for the different markers that are detected. Meanwhile for the omnidirectional case the majority of the markers are positioned with errors greater than 1 m, for the BF case most of the markers are located with precisions better than 10 cm. The mean
error increases to 0.78 m for this case due to the large errors of the markers that are wrongly positioned.

Fig. 6.18. Absolute positioning error distribution for the detected markers with: a) Omnidirectional receiving antennas; b) Beamforming receiving antennas.
From previous statistical results it is apparent that the use of BF antennas at the receiving sensors can increase the positioning accuracy of location systems in rich multipath indoor environments. This is because the antenna effectively filters out some of the multipath.

### 6.3.2. Impact of the CDMA Near-Far Problem

As pointed out earlier, the missed detections can be due to multipath or to the near-far problem [64] inherent to asynchronous CDMA systems. (Recall that the near-far problem is a condition in which a receiver captures a strong signal and thereby makes it impossible for the receiver to detect a weaker signal). In this section we determine which portion of the missed detections for both omnidirectional and BF antenna positioning systems is due to each effect.

To do so, the same antenna configurations analyzed in section 6.3.1 are now considered, but their placement is in free space (no walls or reflecting objects are present). The positioning results are shown in table 6.2. It is observed that the number of detections is close to 50% for both the omnidirectional and BF cases. The conclusion is therefore that for our random marker distributions 50% is the detection maxima that we are able to achieve in our scenario.

<table>
<thead>
<tr>
<th>Positioning Accuracy in Free Space</th>
<th>Omni. Antennas</th>
<th>BF Antennas</th>
</tr>
</thead>
<tbody>
<tr>
<td># Detections</td>
<td>49.2 %</td>
<td>50 %</td>
</tr>
<tr>
<td>&lt; 10cm error detections</td>
<td>60.9 %</td>
<td>76.9 %</td>
</tr>
<tr>
<td>Mean Error</td>
<td>1.19 m</td>
<td>0.82 m</td>
</tr>
</tbody>
</table>

Table 6.2. Free space positioning results summary.

Figure 6.19 provides a graphical representation of the absolute positioning error distribution for the different markers that are detected.
Fig. 6.19. Absolute positioning error distribution for the detected markers with: a) Omnidirectional receiving antennas; b) Beamforming receiving antennas.

Results show that now we have much more similar error distributions, with 60.9% of the markers positioned with less than 10 cm error in the omnidirectional case compared to 76.9%
in the BF case. Also, the mean positioning errors are now closer. This can be explained by the fact that the use of BF antennas can mitigate the near-far problem in the case that different beams capture the signals from different markers. However, the advantage is very limited and just offers about 15% improvement in the positioning accuracy.

Taking the free space cases as the best marker detection possible, the benefit of using BF antennas in the positioning accuracy is clear. For the omnidirectional cases, in free space, 49.2% of the markers can be detected, while this number drops to 43.8% in an indoor environment. This implies that the multipath worsens detection by 5.4%. In BF systems the drop is just 0.8%. A further look into positioning accuracies gives some more insight into the BF benefits. Omnidirectional systems have an average positioning error of 1.19m in free space, but it increases to 3 m in indoor environments due to the multipath. If BF antennas are used, the average error in indoor environments drops to 0.78 m, which is very close to the best achievable performance given by the free space case. The same occurs if we look at the percent detections with a precision higher than 10 cm. With the use of BF antennas, we are able to remove the multipath scattering effect.

6.3.3. Multi Target Positioning Example

In this section, we provide an example of the positioning accuracy including both antenna systems for the scenario previously considered. Figure 6.19 shows graphically the calculated position of each of the markers for a particular case in which the 5 markers were detected for the BF and omnidirectional cases.
Table 6.3 summarizes the exact positioning results. Notice that the given positions are calculated with respect to the reference marker, placed at (5, 5).

<table>
<thead>
<tr>
<th></th>
<th>Theoretical Position (m)</th>
<th>Calculated Position (m)</th>
<th>Absolute error</th>
</tr>
</thead>
<tbody>
<tr>
<td>Omni Antennas</td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>M1</td>
<td>(2.1, -3.8)</td>
<td>(0.15, -3.42)</td>
<td>1.98 m</td>
</tr>
<tr>
<td>M2</td>
<td>(-0.9, 0.9)</td>
<td>(-3.85, 0.37)</td>
<td>3.08 m</td>
</tr>
<tr>
<td>M3</td>
<td>(-2.9, -2.7)</td>
<td>(-2.02, -1.14)</td>
<td>8.74 m</td>
</tr>
<tr>
<td>M4</td>
<td>(-0.3, -1.2)</td>
<td>(-2.43, -2.2)</td>
<td>2.3 m</td>
</tr>
<tr>
<td>M5</td>
<td>(0.3, 0.8)</td>
<td>(-2.37, 0.01)</td>
<td>2.78 m</td>
</tr>
<tr>
<td>BF Antennas</td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>M1</td>
<td>(2.1, -3.8)</td>
<td>(2.1, -3.78)</td>
<td>0.02 m</td>
</tr>
<tr>
<td>M2</td>
<td>(-0.9, 0.9)</td>
<td>(-0.86, 0.88)</td>
<td>0.045 m</td>
</tr>
<tr>
<td>M3</td>
<td>(-2.9, -2.7)</td>
<td>(-2.86, -2.73)</td>
<td>0.05 m</td>
</tr>
<tr>
<td>M4</td>
<td>(-0.3, -1.2)</td>
<td>(-0.28, -1.27)</td>
<td>0.073 m</td>
</tr>
<tr>
<td>M5</td>
<td>(0.3, 0.8)</td>
<td>(0.34, 0.78)</td>
<td>0.045 m</td>
</tr>
</tbody>
</table>

For this particular case, the mean error using omnidirectional antennas is 3.78 m vs. 0.05 m for the BF case. It is also noticeable that for the BF case all markers are detected with a precision better than 10 cm.
6.3.4. Multiple Target Tracking Example

In this section, we will consider the 2D scenario of figure 6.21 to compare the tracking accuracy of the system with omnidirectional receiving antennas vs. directional ones. An 8 x 7 m$^2$ room defined by walls of $\Gamma=0.5$ reflection coefficient is considered. Three highly reflective ($\Gamma=0.8$) cylinders are also distributed inside the premises. The reference marker is placed roughly at the center of the room and 2 markers following random trajectories are to be tracked.

In figure 6.22 the tracking results, when BF antennas of 25 deg. beamwidth and 10 dB gain are used, are shown and compared to those obtained with omnidirectional receiving antennas. Results show that due to the rich multipath, when using omnidirectional receiving antennas, the algorithm is having trouble in tracking both markers. Positioning errors of a few meters
are common and even locations outside the room are given. On the other hand, when beamforming antennas are used, the tracking accuracy increases considerably.

Fig. 6.22. Tracked path of two targets moving in a realistic indoor environment when using omnidirectional (left figure) and BF (right figure) receiving antennas.

6.4. Conclusions

In this chapter we have analyzed the benefit of using phased arrays as the receiving antennas of indoor location systems. A 60 GHz code division multiple access carrier-based ultra-wide band positioning system for the simultaneous location of diverse markers has been considered.

The positioning error in both free space and indoor environments has been assessed when the receivers are equipped with omnidirectional or beamforming antennas. In the second case, the receiving sensors are doted with phased arrays that scan the beams through space. Using extra intelligence in the positioning algorithm, the markers position can be calculated.

Results show that beamforming systems can effectively mitigate the multipath effects in indoor spaces and achieve positioning accuracies close to those achievable in free space.
scenarios. An increase over 50% on the number of detections with positioning accuracies
greater than 10cm is achieved with BF systems with respect to omnidirectional ones. This
confirms that the performance of omnidirectional systems is severely degraded by multipath
and multiple access interference. However, by using BF antenna systems, we can recover the
positioning performance that one can achieve in the best-case scenario, free space.

The near-far problem of CDMA systems has also been addressed to determine its effect on the
positioning accuracy separately from that of the use of different antenna systems.

The 60 GHz band is recommended for such location systems given the available bandwidth
and the antenna size reduction that make high-gain arrays feasible.
Chapter 7

CONCLUSIONS

A method to accurately determine the complex permittivity of packaging and substrate materials at the millimeter-wave band has been proposed. The validity of the measurement results obtained with “The covered transmission line method” has been established and its accuracy deeply analyzed. A wealth of measurements has been provided as a reference for mm-wave antenna designers.

We have also studied a metamorphic material, consisting of variable resistance loaded loops over a ground plane. A mathematical formula for the reflection coefficient of an array of resistively loaded infinitesimal loops has been obtained. A complete set of metamorphic states: Perfect Electric Conductor, Perfect Magnetic Conductor, Perfect Absorber, and Perfect Amplifier can be achieved by such a material. A prototype of this novel material has been fabricated and demonstrates the feasibility of obtaining a tunable artificial intelligent substrate for both circuit and antenna applications.

Millimeter-wave frequencies allow and, in most cases require, the integration of beamforming antenna systems into small areas to compensate for the extra path losses at 60 GHz compared to lower frequencies. We have presented different prototypes of grid antennas that can achieve high gains and are suitable for mm-wave applications. Moreover, a unique beam-selection grid array with multiple feeds that can achieve up to 5 different beams has
been fabricated and measured. The importance of the accurate characterization of the substrate material has been demonstrated.

We have also studied the impact of the antennas in mm-wave indoor positioning and target tracking systems. Using beamforming antenna systems, the accuracy on the determination of various target positions is comparable to that of an equivalent free space scenario. This represents a major improvement compared to the accuracy achievable with current omnidirectional antenna systems.

Future work should address the use of the variable resistance loaded loop metamorphic material in beamforming antenna systems and explore other applications. A mm-wave indoor positioning system making use of beamforming antenna systems should be implemented and theoretical results here presented corroborated by measurements.
Bibliography


