MASTER

Analysis into intentional electromagnetic interference of electronic systems

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Preface

This thesis describes the results of my final MSc project in Electrical Engineering at the Eindhoven University of Technology (TU/e). The project has been initiated by Prof.Dr.Ir. A.P.M. Zwamborn of the Physics and Electronics Laboratory of the Netherlands Organization for Applied Scientific Research (TNO-FEL). The project is sponsored and supported by the Royal Netherlands Army (RNLA). The work has mainly been carried out at TNO-FEL located in The Hague. The measurements have been partly performed at the Center for Technology and Mission Support (CTM) of the RNLA located in Dongen.

At the begin of this thesis, I wish to express my gratitude towards a number of people who support me with this work. First of all to my supervisors Peter Zwamborn and Bart van Leersum. Peter for providing the possibility to perform my project at TNO-FEL, his support, ideas and many EM related discussions. Bart I want to thank for his detailed technical support within the field of High Power Microwave and 'Finite Difference Time Domain' method. Furthermore, I would like to express my gratitude to my employer CTM giving me the opportunity to get my MSc., Albert van Diemen of the Matlogco / IV&C for the financial support, advice and involvement during this project.

Of course, a pleasant environment is a great importance. Therefore I would like to thank the people at the 'EM security' group of TNO-FEL for helping, guiding and teaching me. In this respect I would especially like to mention Yolanda Barrell, Chris Clemens and Jan Rheebergen for their various contributions.

Last, but not least. I would like to thank my friends, my family and especially my girlfriend Sandra for her continuous support during the years I spent studying at the TU/e.

Disclaimer:
All material in this document is based on what is available in the public domain and public literature.
Abstract

Electronic systems are susceptible to electromagnetic wave interaction. These waves generate undesired currents and voltages in electronic circuits, which are added to the normal signals and at a certain level will interfere with the normal operation of the circuit. A new threat for military and civil systems, in the higher frequency spectrum, are high power microwave (HPM) and ultra wide-band (UWB) sources. These sources generate electromagnetic fields, ranging from kilowatts to gigawatts, in short pulses with fast rise times. Intense electromagnetic fields are known to cause damage in many electronic systems. In general, the frequencies range from 200 MHz up to 5 GHz. An attack with this kind of systems is an example of intentional electromagnetic interference (IEMI).

HPM and UWB waveforms are not easy to predict. Currently there is no standard for design, testing and countermeasures available. Research is the only way to provide sufficient insight into the coupling of these signals and interaction in electronic systems.

High power microwave energy can penetrate electronic systems through the "Front door" (antennas) or the "Back door" (cables and apertures). Permanent damage to electronic systems is generally caused by thermal effects or electrical breakdown in the circuits. It can be concluded that electronic is sensitive for the energy contents and / or peak value of the HPM field. To develop protection measures it is very important to know the sensitivity of electronic equipment for these kind of signals. When a trend is noted this can be used for setting up HPM test methods at appropriate field levels.

In order to study coupling into systems, appropriate models are used to model the housing of a real piece of 19" IT equipment to determine the possible resonant frequencies. The housing is modeled as an enclosure with two small apertures. The frequencies which are calculated analytically and defined in the frequency and time domain measurements are very close. The resonance frequencies calculated by using FDTD are a somewhat lower.

An existing system, on which the model is based, is tested for susceptibility. Some frequencies of interference are lower than the numerical and measured resonance frequencies. An important issue for future research is to investigate the cause of that difference.

When using plane-wave excitation within FDTD, by using the scattering formulation, it has been recognized that numerical errors were present. Further analysis is recommended. It is observed that the amplitude (of the resonance frequencies) depends on
the location in the enclosure.
FDTD simulation of the field inside the enclosure and measurements on the (empty) enclosure have shown that the electric field at discrete frequencies is significantly higher than outside the enclosure.
The most accurate way to measure the resonance frequencies is performed with a monopole with a length of less than 1/4 of the enclosure height. The measurements confirm that the resonance frequencies are not correlated to the angle of incidence.
To obtain reliable results it has been shown that performing dedicated measurements leads to the best approach. However, numerical evaluations are important to estimate an appropriate test-plan for measurements.
The experience with the time domain measurements with the impulse radiating antenna (IRA) shows that this is a relatively fast method to obtain results. It is recommended to investigate the possibility of detecting the resonance frequencies by means of reflection measurements in combination with the ground penetration radar (GPR) detection algorithm.
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List of abbreviations

AR  Autoregressive
DFT  Discrete fourier transform
EM  Electromagnetic
EMC  Electromagnetic compatibility
EMI  Electromagnetic interference
CW  Continuous wave
FDTD  Finite difference time domain
FT  Fourier transform
GPR  Ground penetrating radar
GPS  Global position system
HPM  High power microwave
HPEM  High power electromagnetics
ICNIRP  International commission on non-ionizing radiation protection
IEMI  Intentional electromagnetic interference
IRA  Impulse radiating antenna
IT  Information technology
LEMP  Lightning electromagnetic pulse
NEMP  Nuclear electromagnetic pulse
PEC  Perfectly electric conducting
PEL  Permissible exposure level
PRF  Pulse repetition frequency
RF  Radio frequency
RLNA  Royal Netherlands army
SAR  Specific absorbing rate
SATCOM  Satellite communication
STANAG  NATO standardization agreement
TDR  Time domain reflection
TE  Transverse electric
TEM  Transverse electromagnetic
TM  Transverse magnetic
TWT  Travelling wave tube
UWB  Ultra wideband
Chapter 1

Introduction

1.1 Electromagnetic interaction

We are surrounded by electromagnetic (EM) fields. The EM spectrum is full of transmitter signals and therefore the spectrum is a enormous multiplexer. These EM fields interact with humans and/or electronic systems.

Aspects of EM interactions are:

- possible health-hazards,
- unintended degradation of functionalities of electronic systems,
- intentional degradation or damage of the functionalities of electronic systems.

1.1.1 Possible health hazards

Exposure to RF at sufficiently high intensity can lead to effects associated with tissue heating and/or induce currents in the human body that can lead to shocks or burns depending on the frequency. The RF protection standards against detrimental health effects are based primarily on the specific absorption rate (SAR), expressed in W/kg. The standards take the frequency dependence of SAR into consideration, and give appropriate advice on the limitation of plane wave field intensity. Examples of standards are the NATO standardization agreement, STANAG 2345 and the recommendation of the International Commission on Non-Ionizing Radiation Protection (ICNIRP).

The basic dosimetric limit for RF exposure in the frequency range from 100 kHz to 6 GHz [2] is a whole-body SAR of 0.4 watts per kilogram (W/kg). This level incorporates a safety factor of 10 on the SAR of 4.0 W/kg, which has been assumed to be a threshold for potentially deleterious biological effects in people. Below 100 kHz, internal current density resulting in electro-stimulation of biological tissue is the basic dosimetric parameter. Above 6 GHz [2], the exposures are quasi-optical and power density is the applicable exposure parameter. The Permissible Exposure Level (PEL) is given in terms of measurable field components as a convenient correlation to the SAR.
1.1.2 Unintended degradation of functionalities of electronic systems

Electronic systems are susceptible to electromagnetic wave interaction. These waves generate undesired currents and voltages in electronic circuits, which are added to the normal signals and at a certain level will interfere with the normal operation of the circuit.

To avoid unwanted disturbance from e.g. transmitters, radar, lighting (LEMP) and nuclear electromagnetic pulse (NEMP), military and civil systems should satisfy electromagnetic compatibility (EMC) requirements. Examples for military systems are the STANAG 4245 for NEMP and the STANAG 4236 for LEMP. Other well know EMC military standard are the MIL-STD 464(A) and the MIL-STD 461(E). A example of a civilian standard for LEMP is the EN61000-4-5.

Nowadays, electromagnetic interaction is becoming more and more important due to:
1. increasing use of automated electronic systems, communication, navigation, medical, advanced military equipment, etc,
2. increasing susceptibility of electronic systems due to advanced technology, higher equipment and component densities,
3. increasing use of the EM spectrum.

Historically, the interaction of electromagnetic energy / fields and electronics was first investigated using the lightning electromagnetic pulse (LEMP). Against this threat, fuses, (cable) shielding and protection devices were developed. The first nuclear detonations acted as a highly intensive source of electromagnetic radiation. This nuclear electromagnetic pulse (NEMP) showed even more energy density per pulse at higher frequencies. Therefore, the development of faster protection devices and more dense shielding was required as an efficient protection to preserve the electronic systems.

1.1.3 Intentional degradation or damage of the functionalities of electronic systems

New threats for military and civil systems, in the higher frequency spectrum, are high power microwave (HPM) and ultra wide-band (UWB) sources. These sources generate electromagnetic fields in short pulses with fast rise times. In the last decade it has become clear that these pulsed electromagnetic fields can disturb, damage or even destroy modern electronics and decrease the reliability, availability and usability of an illuminated system e.g. an IT infrastructure.

Electromagnetic attacks may happen, not only in warfare circumstances but also in criminal terroristic actions aimed at the critical infrastructure of our modern society.

Intentional electromagnetic interference is defined as [27]: 'The intentional malicious generation of electromagnetic energy to introduce noise or high level disturbance in electrical or electronic systems'.

The effects can be degradation or destruction. In contrast to LEMP and NEMP field
strengths and waveforms for HPM and UWB are not yet predictable. These predictable waveforms of LEMP and NEPM have resulted in test levels, test methods and counter measures. Currently, for HPM and UWB no standard for design and testing is available and research is the only way to provide sufficient insight to the in coupling of these signals in electronic systems and interaction.

1.2 Goal of the project

To avoid intentional disturbance of critical systems it is important to harden these systems against HPM and UWB signals. The goal of the M.Sc. project is to gain more insight in terms of the sensitivity of electronic systems in general. The interpretation of the project description is very broad. An literature survey [10] and consultation with the supervisor led to a focus in the project to determine the resonance frequencies of electromagnetic fields which couple into a metal enclosure with apertures. Numerical models, such as the Finite Difference Time Domain technique (FDTD) is used to determine these frequencies.

The results of this project can be used to develop test methods to evaluate protection measures for electronic equipment in relation to HPM. Research at TNO-FEL within the field of High Power Electromagnetics (HPEM) [22] has been demonstrated that knowledge of the resonance frequencies of the enclosure of a PC system is very important if one wants to predict the (possible) vulnerable frequencies.

1.3 An overview of the report

This report starts, in Chapter 2, with a description of high power electromagnetic threats like lightning, nuclear electromagnetic pulses, ultra wide-band phenomena and high power microwaves. It also includes an overview of the technology of HPM sources, their technical aspects and the coupling into electronic systems in general. Chapter 3 explains the basics of the analytical method which is used to calculate the resonance frequencies and the FDTD method. This Chapter also gives a description of the enclosure with two apertures that is used in the analytical calculations and FDTD simulations. The results of the simulations are, in Chapter 4, compared to measurements in frequency and time domain. Conclusions and recommendations are given in the last two chapters.
Chapter 2

High power electromagnetic threat

2.1 High power electromagnetic environment

The High Power Electromagnetic (HPEM) environment can be defined as the region, in which the peak of the electric field exceeds 100 V/m [26]. This corresponds to a plane-wave free-space power density level of 26.5 W/m². The HPEM environment includes the lightning electromagnetic pulse (LEMP), nuclear electromagnetic pulse (NEMP), ultra wide-band pulse (UWB) and high power microwave signals (HPM). It also includes high power signals from sources such as radar or other transmitters in the vicinity of an electronic system. Figure 2.1 represents an amplitude relation between these sources as function of the frequency.

![Figure 2.1: high power electromagnetic environment](image)

General classes of HPM systems are narrow-band and (ultra) wide-band. These are typically found at higher frequencies > 200 MHz as shown in Figure 2.1. Lightning environments are variable, the maximum frequency goes up to 20 MHz [15].

The integration of the NEMP in military tactical scenarios led to ongoing concepts where energy pulse generators without nuclear detonation are considered and developed. Today these activities are gathered under the umbrella of HPM, which includes
sources, wave propagation, reaction of the electronics and system hardening against this threat.

It is nowadays well established that sufficiently intense electromagnetic signals in the approximate frequency range of 200 MHz to 5 GHz [15] can cause damage in many electronic systems. Such an environment can be:

- a single pulse with many cycles of a single frequency (an intense narrow-band signal that may have some frequency agility),
- a burst containing many pulses, with each pulse containing many cycles of a single frequency,
- an ultra-wide band pulse (spectral content from hundreds of MHz to several GHz),
- a burst of many ultra-wide band transient pulses, which can be radiated or conducted.

2.2 The electromagnetic interference problem

There are three essential elements to any electromagnetic interference (EMI) problem. There must be a source of an electromagnetic phenomenon, a receptor (or victim) and a coupling path in between. Each of these three elements must be present. In general, interference problems are solved by identifying at least the last two of these elements and eliminating one of them.

In case of a HPM attack the source is the HPM system, the propagation through the ether is the coupling path and the illuminated system (communication system, IT infrastructure, personal computer etc.), is the receptor.

2.3 High power microwave systems

High power microwave systems are sources of high power microwave energy ranging from kilowatts to gigawatts. The generation of microwave energy can be accomplished in many ways. In general a HPM system has three or four basic components.

1. power supply,
2. pulse forming network (PFN),
3. RF generator,
4. antenna.

![Figure 2.3: narrow-band HPM system.](image)

![Figure 2.4: wide-band HPM system.](image)

### 2.3.1 Power supply

The power supply converts the applicable power to (high voltage) DC and is connected to the pulse forming network. The applicable power can vary from mains to battery power. HPM is characterized by very short pulses with a high peak power. The average power is much less than the peak power. The required (average) power is then relatively low.

### 2.3.2 Pulse forming network

The pulse forming network converts the energy from the power supply into an appropriate sort of high-voltage pulse to drive a RF source for a narrow-band signal, or to drive an antenna directly to obtain a wide-band field. Sources exist, which can deliver gigajoules of energy and mega-Amperes of current. An example of a pulse forming network is the Marx generator, which was developed by E. Marx. Another example is the explosively driven flux compression generator.

**Marx generator**

A Marx generator is a convenient manner of obtaining very high voltages at very high powers. The basic idea behind the Marx generator is to charge capacitors in parallel and discharge them in series. In this instance, the capacitors are charged through a ladder network of resistors with spark gaps serving as the switching element. The lowest gap is allowed to break down due to overvoltage or it is triggered by an external source. As the voltage rises, the spark gaps eventually break down into conduction.
essentially connecting the capacitors in series. Ideally, the output of a Marx generator is the number of stages times the charging voltage. A schematic representation of the Marx generator is given in Figure 2.5(a).

**Flux compression generator**

An other way to generate a pulse is with an explosively driven flux compression generator, see Figure 2.5(b). Basically a coaxial flux compression generator consists of a tube of high power explosives inside a metal coil, through which an initial current flows. When the explosive detonates, an armature tube surrounding the explosive, usually made of copper, is forced into the metal coil surrounding the armature tube. This causes a short circuit. The explosion propagates continuously down the tube, forcing the circuit to shorten, until the peak current is produced and the explosive consumes the device. A high current is generated. The flux compression generator is an example of a single-shot device.

2.3.3 RF generator

Nearly all types of sources convert kinetic energy of electrons into microwave energy. These electron beam sources can produce a Continuous Wave (CW) or a pulsed narrow-band output waveform. The Radio Frequency (RF) generator is directly connected to an appropriate transmitting antenna. An short overview of generators can be found in Appendix A.

2.3.4 Transmitting antenna

Up to now, we have not asked ourselves how the EM energy is produced. The radiation or launching of waves into space is accomplished efficiently with the aid of conducting or dielectric structures called antennas. Theoretically, any structure radiate EM waves.
but not all structures can serve as efficient radiation mechanisms. An antenna may also be viewed as a transducer, which is used for matching the transmission-line or waveguide to the surrounding medium or vice versa. Figure 2.6 shows how an antenna is used to accomplish a match between the line or guide and the medium.

![Antenna Diagram](image)

Figure 2.6: antenna which is used as a matching device between the guiding structure and the surrounding medium.

The antenna choice depends on the output signal of the generator. For narrow band, signals horns in combination with a reflector are generally used. Ultra wide band pulse generators need an dispersion-free antenna, like a Impulse Radiating Antenna (IRA).

The radiation pattern of an aperture antenna is the Fourier transform [4] of the field distribution in the aperture. Properties for various antenna aperture distributions are given in Table 2.1, in which \( \lambda \) is the wavelength, \( a \) the (horizontal or vertical) dimension of the rectangular antenna and \( x \) is the location on that side.

<table>
<thead>
<tr>
<th>Type of distribution</th>
<th>Beamwidth ( \theta ) (degrees)</th>
<th>First sidelobe intensity ( k )</th>
<th>Efficiency factor</th>
</tr>
</thead>
<tbody>
<tr>
<td>uniform ( f(x) = 1 )</td>
<td>50.32 ( \lambda / a )</td>
<td>0.217</td>
<td>1.0</td>
</tr>
<tr>
<td>cosine ( f(x) = \cos(\pi x / 2) )</td>
<td>68.84 ( \lambda / a )</td>
<td>0.071</td>
<td>0.81</td>
</tr>
<tr>
<td>raised cosine ( f(x) = \cos^2(\pi x / 2) )</td>
<td>83.26 ( \lambda / a )</td>
<td>0.027</td>
<td>0.667</td>
</tr>
</tbody>
</table>

The \( \cos^2 \) distribution in the aperture gives the smallest side-lobes but this must be paid for with a lower efficiency. For a rectangular aperture, the radiation patterns in the two
perpendicular planes may be different, depending on the field over the aperture in that plane.

The *power gain* for a general aperture antenna is given by,

\[
G = \frac{4\pi A_e}{\lambda^2}, \quad (2.1)
\]
\[
A_e = kA. \quad (2.2)
\]

in which \( A_e \) is the effective aperture area, \( k \) is the efficiency factor and \( A \) is the cross-sectional area of the aperture.

The beamwidth \( \Theta \) can be approximated in radians as follows:

\[
\Theta = \frac{\lambda}{ka}. \quad (2.3)
\]

*Horn antenna*

The source for a high-frequency antenna becomes an open-ended waveguide or horn. Then the radiated electric field is directly related to the electric field in the aperture of the source (Huygens principle).

The beamwidth, efficiency and thus the gain are mainly dependent on the dimensions of the aperture, the frequency and aperture field distribution.

For a conventional open-ended rectangular waveguide horn, shown as in Figure 2.7, there is a cosine distribution of field in the horizontal plane. The horizontal radiation pattern is characterized by a beamwidth of \( 60.8\lambda/a \) and an efficiency factor \( k=0.81 \), whereas there is a uniformly distributed field in the vertical plane. The vertical radiation pattern has a beamwidth of \( 50.3\lambda/b \) and the efficiency factor is one. A flare at the end of the waveguide both increases the cross-sectional area of the antenna aperture and provides an impedance match between the waveguide and the intrinsic impedance of free space.

Horns flared in just one plane are called sectoral horns and those flared in both planes are called pyramidal horns. Various horns are illustrated in Figure 2.7.

The radiation patterns from a horn depend on the flare angle and the length of the horn. Too sharp a flare angle does little to improve the impedance mismatch, but introduces an unacceptable large phase error. An acceptably limit for phase error can be set by \( \delta < \lambda/8 \) [4].

The minimum length for the horn is given by

\[
\ell_{\text{min}} = \frac{a^2}{\lambda} + \frac{\lambda}{16} \quad (2.4)
\]
For large apertures, the horns that satisfy these conditions are long and bulky. Higher gain and narrower beam width are obtained using the reflector antennas, described in the next section.

In HPM applications, where power levels are possible in the giga-Watt range, precautionary measures must be taken to prevent air-breakdown occurring in the wave-guide and antenna. The nominal breakdown level for air is $\sim 1 \text{ MV/m}$.

A nice proposal for a horn antenna design can be found in [30]. Some details of the proposed horn can be found in Figure 2.8. The antenna consists of a vacuum wave-guide and horn. The horn aperture size is such that the power density and the peak electric field at the end of the horn enable a transition from vacuum to 1 atm $SF_6$ gas. Nominally, this means that the peak electric field at the horn aperture must be below 3 MV/m to avoid excessive electrical stress causing breakdown. The $SF_6$ container is large enough so that anywhere the peak electric field outside the container is of the order of 1 MV/m or less.
**Reflector antenna**

The easiest way to increase the area of an antenna (horn) is to use a reflector, designed according to the principles of geometrical optics. The antenna consists of two basic elements, a relatively small feed (horn) and a large reflecting surface. A parabolic reflecting surface has the useful property of transforming a diverging spherical wavefront into a parallel plane wave-front.

In a reflector antenna, the aperture of the reflecting surface controls the radiation pattern and hence the gain, beamwidth and sidelobe levels. The feed horn can be designed to produce an desired amplitude distribution of electromagnetic energy across the aperture of the reflector.

To eliminate the shadow of the feed horn and its support structure is to locate the focus point of the reflector away from the reflected beam. This is called an offset paraboloid antenna, as shown in Figure 2.9(a). Another system which removes the feed antenna horn from the front of the reflector is the Cassegrain system, as shown in Figure 2.9(b). The gain of a reflector antenna is still given by equation 2.1. Uniform illumination across the aperture of the reflector is the ideal situation, but for most systems having non-uniform illumination across the aperture, the efficiency factor, $k$, usually lies between 0.5 and 0.6 [4].

For high power signals the feed horn can for example be the one in Figure 2.8. Some configurations of reflector antennas are presented in Figure 2.9.

![Figure 2.9: some HPM reflector configurations [30].](image)

(a) a single reflector system fed by an offset feed horn.  
(b) a dual-reflector system.

**Impulse radiating antenna**

An UWB pulse is characterized by a very short rise time and a very slow decay. A UWB pulse can be only radiated efficiently by a dispersion free antenna, which is not a conventional antenna. The Impulse Radiating Antenna (IRA) is an example of a dispersion free antenna. The IRA at TNO-FEL is depicted in Figure 2.10. The IRA is useful for both transient and continuous wave illumination of facilities or systems.

Typical characteristics of IRA's are:

- produces a very narrow beam in the boresight direction,
• spectral response is reasonably flat from tens of MHz to several GHz,
  – low frequency limit is due to aperture size,
  – high frequency limit is due to source spectral fall-off (pulser rise time),
• radiates an impulse-like field in boresight direction.
  – with sub-nanosecond rise times,
  – E-field strengths ~4 kV/m at distances of in the order of hundreds of meters.

A detailed description of the IRA can be found in Appendix B.

![Figure 2.10: example of an impulse radiating antenna.](image)

### 2.4 Typical waveforms

This section classifies the waveforms that are produced by HPM sources into four different types of waveforms [20]. They all have advantages and disadvantages. Therefore, these four waveforms can occur in a hostile EM environment. The actual waveforms, which can be expected, might be a combination of these four waveforms. For comparison reason, the waveform types are shown in Figure 2.11 in a convenient way. The DFT of the time-signal is taken over the first 400 ns. The time signal is sampled with a time step of 0.01 ns, so the total number of samples is N=40000.

#### 2.4.1 Continuous wave

A HPM system that produces a CW waveform generally includes a resonant microwave tube. An example of a CW source is a microwave oven, of which the waveguide has been modified. The waveguide is connected to a horn antenna, which possibly feeds a parabolic disk. The typical frequency and power of this homemade HPM device are 2.45 GHz and 1 kW, respectively. The CW waveform can be characterized by two parameters:
- frequency [Hz],
- power [W].

The CW signal is presented in Figure 2.11(a).
This signal is most efficient when the frequency is tuned on the most vulnerable frequency of the target. All the HPM energy is concentrated into one frequency.

### 2.4.2 Narrow-band

A narrow-band (repetitive) waveform is presented in Figure 2.11(b). This is a typical waveform used for IEMI. The frequencies of available systems vary from roughly hundreds of MHz to tens of GHz. These sources are generally not tunable at all or just over a short frequency range. Peak powers can reach up to several GW. A useful pulse width consist of minimum of 100 cycles, because it takes approximately 100 periods of a sine wave to generate the maximum field in a cavity. The duty cycle is generally less than one percent.

The advantage of this waveform is that a low average power is used to generate efficient high peak power, high energy pulses used for upset or damage of electronic equipment. The highest peak powers are to be expected from these narrow band sources. This signal is most effective when the target is vulnerable for peak signal on the most vulnerable frequency.

The parameters characterizing an HPM source, which produces this waveform are:
- frequency [Hz],
- power [W],
- peak power [W],
- average power [W],
- pulse repetition frequency (PRF) [Hz],
- duty cycle [%],
- number of produced pulses.

### 2.4.3 Damped sinusoid

A damped sinusoid waveform can be generated by an impulsive source, by rapidly discharging an amount of slowly stored electrical energy directly into an antenna, without using an expensive vacuum tube.

The process of discharging electrical energy is limited in speed. Therefore, the frequency content of impulsive sources is limited to a certain maximum.

The explosively driven flux compressor is one example of a single pulse damped sinusoid waveform generator.

The side-lobes around \( f_o \) are lower than the narrow-band pulse. More energy is concentrated on \( f_o \).

Figure 2.11(c) shows the damped sinusoid waveform, with the following characteristic parameters:

20
• dominant frequency [Hz],
• peak power [W],
• damping rate [s⁻¹],
• energy in the pulse [J].

If the waveform is repetitive we also have:
• pulse repetition frequency (PRF) [Hz],
• average power [W],
• duty cycle [%].

2.4.4 (Ultra) wide-band

An UWB pulse presented in Figure 2.11(d) is characterized by a very short rise time and a relatively very slow decay. An UWB pulse can be radiated efficiently by a dispersion free antenna.

UWB pulses have the ability to excite many resonant modes in an object or target. Therefore the possibility that the pulse is able to excite a vulnerable frequency of the target is high. However, the power that is present in a small portion of the frequency range is relatively low and therefore less suitable to generate high field strengths in a resonant target for upset or damage. The behavior of UWB pulses in IEMI applications needs more research before a conclusion on the usability of UWB for that purpose can be drawn.

The ultra wide-band waveform can be characterized by:
• peak power [W],
• rise time [s],
• decay time [s],
• bandwidth [Hz].

2.5 Propagation

If a HPM source is to be used for illuminating a target with an intense microwave field, a number of propagation factors must be taken into account. For targets in the earth’s upper atmosphere these factors become less significant. However, when the target is located near earth’s surface, atmospheric refraction and absorption must be considered. Moreover, ground reflection and surface diffractions may reduce or enhance the illumination intensity significantly.

The most important propagation issues are shortly discussed in the next section.

2.5.1 Maximum propagation distance

The distance over which a target can be illuminated is firstly limited by the fact that the earth is round, as a result of which at a certain distance $d$ the source and target can disappear behind each other’s radio horizon. For standard atmospheric conditions the maximum propagation distance $d_{max}$ is given by [16]
(a) continuous Wave (CW), $f_0=1\ \text{GHz}$, CW in frequency domain is a delta-function

(b) narrow-band, $f_0=1\ \text{GHz}$, $t_d=200\ \text{ns}$

(c) damped sinusoid, $f_0=1\ \text{GHz}$, $t_d=60\ \text{ns}$

(d) (ultra) wide-band, $t_r=200\ \text{ps}$, $t_d=4\ \text{ns}$

Figure 2.11: example of different Waveforms in time-domain and frequency domain (single pulse).
in which \( h_z \) is the height of the transmitting antenna, \( h_o \) the height of the target (receiving antenna) and \( r_a \) the radius of the earth (6370 km).

For example, if a HPM source (antenna) is placed at a height of 3m, then a target at a height of 3m must be within a distance of \( d_{\text{max}} = 14 \text{ km} \).

### 2.5.2 Two-ray model

For propagation paths with a length smaller than \( d_{\text{max}} \), the received signal level at the target is often described by the model shown in Figure 2.12.

\[
\frac{E}{E_0} = |(1 + \rho \exp(j\psi))|, \quad \psi = \frac{2\pi \Delta d}{\lambda},
\]

in which \( E_0 \) is the field strength of the direct wave at the target, \( \lambda \) is the wavelength, \( \psi \) the phase difference between the reflected and the direct wave, caused by the different pathlength \( \Delta d \) and \( \rho \) represents the complex reflection coefficient of the ground. In the
case that \( \theta \) is near zero \((d \gg h_1 \text{ and } h_2)\) then, for both polarizations, \( \rho = -1 \). When \( \rho = -1 \), equation 2.6 can be rewritten as \([16]\),

\[
\left| \frac{E}{E_0} \right| = \left| 2 \sin \frac{2\pi h_1 h_2}{\lambda d} \right|.
\]

We can see that the field strength for a certain \( h_1 \) is dependent on \( h_2, d \) and \( \lambda \). The dependency of \( h_2 \) led to the height gain pattern, presented in Figure 2.13.

![Figure 2.13: height gain pattern.](image)

I can demonstrate that each doubling of the length of the propagation path leads to an increase of the transmission loss of 12 dB (40 dB/dec).

For example, assume a HPM source operates at 1 GHz and the antenna is placed at a height of 3 m and the target is at a distance of 100 m. The first maximum will appear at a height of \( \frac{\lambda d}{2h_1} = 2.5 \text{ m} \) and the zero field (after ground-level) will appear at a height of \( \frac{\lambda d}{2h_1} = 5 \text{ m} \). At a distance of 1000 m the maximum will appear at a height of \( 25 \text{ m} \).

Obstacles in the path between the source and target can (partly) block the EM waves and as a consequence efficiency is reduced.

Note that the polarization has an effect on the ground reflection. Assuming a highly conducting ground. The electric field strength at ground-level will be zero for horizontally polarized fields. However, vertically polarized fields will be present and therefore vertically polarized microwave fields will have more effect on those targets close to that ground.

### 2.5.3 Building attenuation

In general, when electronic equipment is installed in buildings the wall provides an extra, frequency dependent, attenuation of the HPM signal. The electromagnetic coupling
through structures from the exterior to the interior depends on a number of factors. Radio frequency signals in and around structures may vary considerably from point to point, since the energy may arrive from different directions at different times due to multiple path propagation. The vector addition of signals from multiple paths may result in a deep null at one point or a sharp peak at another. A complete deterministic theoretical analysis of the RF penetration through structures is not easy feasible, which suggests the use of a geometric / statistical model. The attenuation of cinder-block constructed buildings generally falls in the range of 13 - 22 dB (200 MHz - 3 GHz) [31].

2.6 Coupling into electronic systems

High power microwave energy can penetrate electronic systems through the "front-door" and / or the "back-door". This is illustrated in Figure 2.14.

2.6.1 Front door coupling

When the high power microwave signal enters an antenna connected device, the coupling mechanism is called front-door coupling. The amount of coupled energy depends
very strongly on the antenna characteristics. When the operating frequency of the system (GPS, SATCOM) is very close to HPM relevant frequencies the first input stage will be more or less matched. Such systems are therefore highly vulnerable for incoming HPM signals.

2.6.2 Back door coupling

Back-door coupling is a much more complex phenomenon and occurs when high power microwave signals enter systems through openings and slits in the conducting enclosure of the system or though connected cables. This causes interference voltages and currents on wires inside.

2.7 High power microwave interactions with systems

The electromagnetic interaction problem is extremely complex, especially in the case of back door coupling. It is hard enough to calculate the generated field of the high power microwave source on the exterior of complex systems. A system can consist of thousands of wires going to numerous black boxes, through various cavities and structures, causing the problem to become intractable. However, even though detailed response characteristics of complex systems are complicated, some general trends can be summarized, based on various models that have been developed from a lot of experience [18].

The typical response of a system can be divided into three bands with certain characteristics [7] as shown in Figure 2.15.

![Figure 2.15: typical system response.](image)
The first frequency band, Static or Quasi-static region \( f < f_l \)
In the band, the EM waves are long comparing to the dimensions of the system. Apertures and small antennas are usually differentiators. Consider a wire behind an small aperture, which is usually modelled as a transmission-line with a series of voltage. The transfer-function for low frequencies is proportional to \( f \). The same proportionality applies to a small aperture that acts as a high-bandpass filter.

The second frequency band, Resonance region \( f_l < f < f_h \)
This is the resonance region. The transfer function varies up and down as one passes through the resonances. In general, the lowest frequencies can be associated with the major dimensions \( l \) of the system. As the frequency increases shorter wires and dimensions of interior cavities become more important. At higher resonance frequencies the dimensions of doors, windows, boxes of electronics, spacing between highest and rivets become dominant. The typical resonance region for systems goes up to several gigahertz, corresponding to the half- or quarter-wave resonance of the smallest characteristic dimensions involved in the system construction.

The third frequency band, High frequency region \( f > f_h \)
This is the integration region. The various wires in electronic systems carry electronic signals between the various boxes. If the wavelength is less than the distance between these wires and the ground-plane, the induced current is proportional to \( f^{-1} \). The wires act as integrators. At these high frequencies, apertures let in fields in a roughly frequency-independent manner, so the transfer function of exterior fields to wires should be roughly proportional to \( f^{-1} \).

In summary, the frequency spectrum can be divided as follows:

\[
\begin{align*}
  f < f_l & \sim \frac{c}{4l}, \\
  f_l < f < f_h & \sim \frac{5c}{l}, \\
  f_h > f &.
\end{align*}
\]

in which \( l \) is the dimension of the system and \( c \) is the speed of light.

In general, electromagnetic energy can enter a port of interest of a system through apertures, diffusion and connected transmission-lines. Detailed information about coupling of EM fields into transmission-lines can be found in [29].

2.8 Examples of HPM system Threats
High power microwave systems can attack any system that contains vital electronic sub-systems or components, like IT infrastructure, communication systems, motor management systems of vehicles, etc. High power microwave weapons can be placed
in mobile platforms like vehicles, ships and planes. An example is shown in Figure 2.16(a). In this example it is the intention to disturb the IT infrastructure. An example of a car stopper is given in Figure 2.16(d). Another threat is a suitcase with an HPM source and antenna inside which is commercially available, see Figure 2.16(c). The effects of HPM on electronics may result in upset or, at high levels of irradiation, even permanent physical damage. Upset (i.e. interference or disturbance) is caused by false in-band signals due to non-linear effects in the electronic components originating from envelope detection of the HPM. The upset may be temporary, i.e., the equipment returns to full function spontaneously after the irradiation, or a permanent failure of the function may occur, i.e. the equipment will require a manual restart or reset. Permanent damage is caused by thermal effects or electrical breakdown in the circuits. In this case the damaged component or equipment has to be repaired or replaced [3].

Figure 2.16: example of HPEM attacks.

2.8.1 Personal computer attack

The first practical example is the disturbance of a personal computer with HPM. In this situation a PC is located in a building and the HPM source and antenna are located in
a delivery van, as shown in Figure 2.16(a).
To get an idea of the power, which is needed to disturb (power down) the PC, a simple
calculation is made, based on the data below.

HPM source:
- source: Magnetron,
- antenna: Standard gain horn, \( G = 16.5 \) dBi,

Coupling path:
- medium: ether (air),
- distance: \( d = 10 \) m and \( 100 \) m between source and target,
- propagation: building attenuation of \( A_p = 22 \) dB [31].

Target:
- susceptibility: Pentium II 300 MHz, \( E_s = 100 \) V/m at \( f = 1400 \) MHz [22].

Solution:
The power needed to generate an electric field of 100 V/m at 10 m and 100 m can be
calculated by using:
\[
S = \frac{P_{G_t}}{4\pi d^2},
\]
with,
\[
E = \sqrt{S120\pi}.
\]
The needed power will be
\[
P_s = \frac{E^2 d^2}{30G_t} \approx 750\text{W}(10\text{m}), 75\text{kW}(100\text{m}).
\]
Due to the attenuation of 22 dB, the total source power \( P_s \) needed to disturb the PC is:
\[
P_s = P_{A_p} \approx 120 \text{ kW}(10\text{m}), 12 \text{ MW}(100\text{m}).
\]

2.8.2 Satellite receiver attack

The next example is the disturbance of a satellite receiver with HPM. The HPM source
is pointed at the side-lobe of the reflector antenna. The HPM source and antenna are
located in a delivery van of Figure 2.16(a).
To get an idea of the power needed to damage the Low Noise Converter (LNB), a
simple calculation is made, based on the data below.

HPM source:
- source: Pulsed CW Magnetron, \( f = 10 \) GHz (\( \lambda = 3 \) cm),
- antenna: reflector antenna gain \( G_t = 38 \) dBi,
- the antenna is pointed on the first side-lobe of the target.

Coupling path:
- medium: ether (air),
distance: \( d = 100 \text{ m} \) and \( 1000 \text{ m} \) between source and target,

- propagation: to simplify the problem the propagation effects are not taken into account.

Target:

- susceptibility LNB: \( P_s = 0.5 \text{ Watt} \),
- gain reflector antenna: \( G_t = 40 \text{ dBi} \),
- gain side-lobe reflector antenna: \( G_s = 20 \text{ dBi} \).

Solution:
The receiving power \( P_r \) of the receiver antenna is given by the radio-equation

\[
P_r = \frac{P_t G_t G_s \lambda^2}{4\pi d^2} \quad (2.16)
\]

The transmitting power \( P_t \) needed to result in a receiving power of \( P_r = 0.5 \text{ Watt} \) is

\[
P_t = \frac{P_r 4\pi d^2 4\pi}{G_t G_s \lambda^2} \approx 1.4 \text{ kW}(100\text{m}), 140 \text{ kW}(1000\text{m}) \quad (2.17)
\]

2.9 Conclusion of high power electromagnetic threat

High power microwave systems use sources with high power microwave energy ranging from kilowatts to gigawatts. General classes of HPM systems are narrow-band and (ultra) wide-band. The antenna that can be used depends on the output parameters of the generator. For narrow band signals, horns in combination with a reflector are generally used. Ultra wide band pulse generators need (ideally) a dispersion free antenna, like a Impulse Radiating Antenna (IRA). The lowest useful frequency depends, in general, on the dimensions of the antenna. To have a reasonable dimension versus gain ratio, the lower frequencies of HPM systems are typically found at 200 MHz. Ultra wide-band pulses have the ability to excite many resonant modes in an object or target. Therefore, the possibility that the pulse is able to excite a vulnerable frequency of the target is high. It is established that sufficiently intense electromagnetic fields up to approximately 5 GHz are known to potentially cause damage in many electronic systems via back door coupling.

High power microwave energy can penetrate electronic systems through the "front door" and / or the "back door". When the high power microwave signal enters an antenna connected device, the coupling mechanism is called front door coupling. Back door coupling is a much more complex phenomenon and occurs when signals enter systems through openings and slits in the conducting enclosure or though connected cables.

For targets in the earth's upper atmosphere, propagation factors are less significant. However, when the target is located near the earth's surface and the source is at high altitude, atmospheric refraction and absorption must be considered. When both the
HPM antenna and the target are at ground-level, reflection of the signal on the ground results in a height dependent fieldstrength pattern at the location of the target. This results in a practical ground-attack, which will be within a distance of hundreds of meters.

High power microwave systems can be placed in mobile platforms like vehicles, ships and planes. The effects of HPM on electronics may result in upset or, at high levels of irradiation, even permanent physical damage. Permanent damage is caused by thermal effects or electrical breakdown in the circuits.
Chapter 3

Simulation

The penetration of radiated high-intensity fields into cavities through apertures (back­
door coupling) is an important EMI issue as discussed in Chapter 2.6. Electronic equip­
ment is commonly housed in, conducting, boxes with openings. In such structures, an
external signal can penetrate through these apertures and directly couple energy into
the interior and generate currents and voltages in electronic circuits.

A piece of real equipment [19] has been modelled as a conduction enclosure with two
apertures. To get an idea of the resonance frequencies of the enclosure (the amplitude is
of less interest) these are calculated analytically as well as by a Finite Difference Time
Domain Method simulation. Finally, the analytical calculation and FDTD simulation,
are verified with measurements. For the FDTD simulation software EMFDTD 1.4 is
used [9], [22], [23].

3.1 Enclosure with aperture

The metal enclosure which is used as evaluation has the following inner dimensions:
440 × 290 × 42 mm. The aperture dimensions are 20 × 28 mm and 32 × 12 mm. The
gometry is illustrated in Figure 3.1.

Figure 3.1: geometry.
3.2 Cavity resonance frequencies

The enclosure of Figure 3.1 can be interpreted as a rectangular cavity resonator, as shown in Figure 3.2. We notice that the cavity is simply a rectangular waveguide, which has been shorted at both ends. Therefore, we expect to have standing waves and also transverse magnetic (TM) and transverse electric (TE) modes of wave propagation.

In the TE modes, the electric field is transverse (or normal) to the direction of wave propagation. The electric field $E_z = 0$ and the magnetic field $H_y = 0$.

For TM modes, the magnetic field is transverse (or normal) to the direction of wave propagation. The electric field $E_z \neq 0$ and the magnetic field $H_y = 0$.

Depending on how the cavity is excited, the wave can propagate in the $x$-, $y$-, or $z$-direction. In this case, we choose the $z$-direction as the "direction of wave propagation". In fact, there is no wave propagation, only standing waves can exist. A standing wave is a combination of two waves travelling in opposite position.

It can be shown that modes $TE_z, TM_z, TE_y, TM_y, TE_x$ and $TM_z$ satisfy the boundary conditions and are therefore appropriate modes (field configurations) for a rectangular waveguide. We will initially consider $TE_z$ and $TM_z$, which satisfy the following set of equations [5].

3.2.1 Transverse Electric Modes

Since transverse electric modes for a rectangular cavity are derived in a similar manner to those of a rectangular waveguide, they satisfy equations 3.1-3.3 [5].

\[
\begin{align*}
\mathcal{E}_x &= -\frac{1}{\epsilon} \frac{\partial F_z}{\partial y} \\
\mathcal{E}_y &= \frac{1}{\epsilon} \frac{\partial F_z}{\partial x} \\
\mathcal{E}_z &= 0
\end{align*}
\]

\[
\begin{align*}
\mathcal{H}_x &= -j \frac{1}{\omega \mu \epsilon} \frac{\partial^2 F_z}{\partial x \partial z} \\
\mathcal{H}_y &= -j \frac{1}{\omega \mu \epsilon} \frac{\partial^2 F_z}{\partial y \partial z} \\
\mathcal{H}_z &= -j \frac{1}{\omega \mu \epsilon} \left(\frac{\partial^2 F_z}{\partial z^2} + \beta^2 F_z\right)
\end{align*}
\]

in which $F(x, y, z)$ is a scalar potential function, and it represents the $z$ component of the vector potential function $\mathbf{F}$. The potential $F$, and in turn $F_z$, must satisfy the wave
\[ \nabla^2 F_z(x, y, z) + \beta^2 F_z(x, y, z) = 0. \]  \hspace{1cm} (3.4)

in which \( \beta^2 = \beta_x^2 + \beta_y^2 + \beta_z^2 = \omega^2 \mu \varepsilon. \) The parameter \( \beta \) is known as the wave number [rad/m].

Since the boundary conditions on the bottom, top, left and right wall corresponds to a perfectly conducting wall / boundary, the \( F_z(x, y, z) \) function for the rectangular cavity can be written as:

\[ F_z = A_{zmp} \cos(\beta_x x) \cos(\beta_y y) \sin(\beta_z z), \]  \hspace{1cm} (3.5)

in which,

\[
\begin{align*}
\beta_x &= \frac{m \pi}{a}, & m &= 0, 1, 2, \ldots \\
\beta_y &= \frac{n \pi}{b}, & n &= 0, 1, 2, \ldots \\
\beta_z &= \frac{p \pi}{c}, & p &= 1, 2, 3, \ldots 
\end{align*}
\]

Thus for each mode the dimensions of the cavity in each direction must be an integer number of half wavelengths of the wave in that direction.

Substitution of 3.5 into 3.4 results in equation 3.6. \( (f_r)^{TE}_{mnp} \) represents the resonant frequency for the \( TE_{mnp} \) mode and is obtained from:

\[ (f_r)^{TE}_{mnp} = \frac{1}{2 \pi \sqrt{\mu \varepsilon}} \sqrt{ \left( \frac{m \pi}{a} \right)^2 + \left( \frac{n \pi}{b} \right)^2 + \left( \frac{p \pi}{c} \right)^2 }, m = n \neq 0. \]  \hspace{1cm} (3.6)

If \( c \geq a \geq b \), the mode with the lowest order is the \( TE_{101} \) mode.

### 3.2.2 Transverse magnetic modes

In addition to \( TE_{mnp} \) modes inside a rectangular cavity, \( TM_{mnp} \) modes can also be supported and they satisfy equations 3.7 - 3.9 [5].

\[
\begin{align*}
\mathcal{E}_x &= -j \frac{1}{\omega \mu \varepsilon} \frac{\partial A_z}{\partial x}, & \mathcal{H}_x &= \frac{1}{\mu} \frac{\partial A_z}{\partial y}, \\
\mathcal{E}_y &= -j \frac{1}{\omega \mu \varepsilon} \frac{\partial A_z}{\partial y}, & \mathcal{H}_y &= \frac{1}{\mu} \frac{\partial A_z}{\partial x}, \\
\mathcal{E}_z &= -j \frac{1}{\omega \mu \varepsilon} \left( \frac{\partial^2}{\partial x^2} + \beta^2 \right) A_z, & \mathcal{H}_z &= 0.
\end{align*}
\]

\[ \nabla^2 A_z(x, y, z) + \beta^2 A_z(x, y, z) = 0. \]  \hspace{1cm} (3.10)

The modes can be derived a similar manner to \( TE_{mnp} \) field configurations. The vector potential \( A_z(x, y, z) \) will be

\[ A_z = B_{zmp} \sin(\beta_x x) \sin(\beta_y y) \cos(\beta_z z), \]  \hspace{1cm} (3.11)
in which,
\[
\begin{align*}
\beta_x &= \frac{m\pi}{a}, \quad m = 1, 2, 3 \ldots \\
\beta_y &= \frac{n\pi}{b}, \quad n = 1, 2, 3 \ldots \\
\beta_z &= \frac{p\pi}{c}, \quad p = 0, 1, 2 \ldots 
\end{align*}
\]

By substitution of 3.11 in 3.10. The resonance frequencies can be calculated by:
\[
(f_r)^{TM}_{mnp} = \frac{1}{2\pi\sqrt{\mu\varepsilon}} \sqrt{(\frac{m\pi}{a})^2 + (\frac{n\pi}{b})^2 + (\frac{p\pi}{c})^2}.
\]

Note that the equations for \(TE_{mnp}^x\) and \(TM_{mnp}^z\) are the same. Only the conditions for \(n\), \(m\) and \(p\) are different.

We can see that the first resonant frequency is still at mode \(TE_{101}\). This is also called the dominant frequency.

Applying equation 3.12 to the enclosure of Figure 3.1 we can estimate the resonance frequencies. The results are shown in Figure 3.3 and are listed in Table 4.1. The resonance frequencies density increase at higher frequencies.

\[\text{Figure 3.3: resonance frequencies of the enclosure.}\]

### 3.3 Basic Finite Difference Time Domain Method Algorithm

For complex configurations such as structure with several apertures and wires where no (simple) analytical models exist, can be solved with a numerical approach using for example Finite Difference Time Domain (FDTD). The FDTD method is based on the Maxwell equations. Consider a source free, non magnetic, linear, homogeneous
and isotropic medium, characterized by $\sigma$, $\epsilon$ and $\mu_0$. The Maxwell's curl equations are given by

$$\nabla \times \mathbf{E} = -\mu_0 \partial_t \mathbf{H}, \quad (3.13)$$

$$\nabla \times \mathbf{H} = \epsilon \partial_t \mathbf{E} + \sigma \mathbf{E}. \quad (3.14)$$

in which $\mathbf{E}$ is the electric field, $\mathbf{H}$ the magnetic field, both in time domain.

The fundamental idea of the FDTD method is to discretize space and time and approximate the differentials as finite differences. Following Yee's approach [35], the grid points for the $\mathbf{E}$-field and the $\mathbf{H}$-field are chosen at the center of the ribs of the Yee-cell so as to approximate the conditions to be as accurate as possible. The various grid positions are shown in Figure 3.4.

The grid point in the solution region is defined as

$$F(i, j, k) = F^n(i, j, k), \quad (3.16)$$

where $\Delta x$, $\Delta y$ and $\Delta z$ are the space increments, $\Delta t$ is the time increment, and $i$, $j$, $k$ and $n$ are integers.

Note that in Figure 3.4 not all field components exist at the same physical location, a shift of $\frac{1}{2}$ cell is used. This is called a staggered grid. The vector $\mathbf{E}^n = (E_z^n, E_y^n, E_x^n)$
and $\mathbf{H}^n = (\mathbf{H}_x^n, \mathbf{H}_y^n, \mathbf{H}_z^n)$ are defined as

$$E_x^n(i+\frac{1}{2},j,k), \quad H_x^n(i,j+\frac{1}{2},k), \quad H_y^n(i+\frac{1}{2},j,k), \quad H_z^n(i,j,k).$$

(3.17)

(3.18)

(3.19)

Using the second-order accurate central difference approximation, the derivation $\partial_x$ to space follows from

$$\frac{\partial x E^n(i,j,k)}{\Delta x} = \frac{E^n(i+\frac{1}{2},j,k) - E^n(i-\frac{1}{2},j,k)}{\Delta x} + O(\Delta x^2).$$

(3.20)

The differentiation with respect to time $\partial_t$ is obtained from

$$\frac{\partial t E^n(i,j,k)}{\Delta t} = \frac{E^n(i,j,k) - E^n(i,j,k)}{\Delta t} + O(\Delta t^2).$$

(3.21)

in which $O(\Delta x^2)$ and $O(\Delta t^2)$ indicate the magnitude of the truncation error.

### 3.3.1 FDTD total field equations

From equation 3.16 - 3.21, an explicit difference approximation is obtained to approximate Maxwell equations (equation 3.13 and equation 3.14) by

$$\frac{-\mu_0}{\Delta t} (\mathbf{H}^{n+\frac{1}{2}} - \mathbf{H}^{n-\frac{1}{2}}) = (\nabla \times \mathbf{E})^n,$$

(3.22)

$$\frac{\varepsilon}{\Delta t} (\mathbf{E}^{n+1} - \mathbf{E}^n) + \frac{\sigma}{2} (\mathbf{E}^{n+1} + \mathbf{E}^n) = (\nabla \times \mathbf{H})^{n+\frac{1}{2}}.$$

(3.23)

This can be re-written as

$$\mathbf{H}^{n+\frac{1}{2}} = C_{1H} \mathbf{H}^{n-\frac{1}{2}} - C_{2H} (\nabla \times \mathbf{E})^n,$$

(3.24)

$$\mathbf{E}^{n+1} = C_{1E} \mathbf{E}^n + C_{2E} (\nabla \times \mathbf{H})^{n+\frac{1}{2}},$$

(3.25)

in which

$$C_{1H} = 1, \quad C_{2H} = \frac{\Delta t}{\mu_0},$$

$$C_{1E} = \frac{2\varepsilon - \sigma \Delta t}{2\varepsilon + \sigma \Delta t}, \quad C_{2E} = \frac{2 \Delta t}{2\varepsilon + \sigma \Delta t},$$

(3.26)

(3.27)

are constants for homogeneous medium. These are the FDTD total field equations.
3.3.2 FDTD scattered field equations

For a known plane-wave source the incident electric fields $E_i$ and magnetic fields $H_i$ are known. Due to the linear behavior of the media the total field $E$ consists of the superposition of the scattered fields $E_s$ and incident field $E_i$. The incident field is the field that would be present without the object and is, in general, defined analytically.

\[
E = E_i + E_s, \quad (3.28)
\]
\[
H = H_i + H_s. \quad (3.29)
\]

Using the Maxwell equation 3.14 this can be re-written as

\[
\nabla \times (E_i + E_s) = -\mu_0 \partial_t (H_i + H_s), \quad (3.30)
\]
\[
\nabla \times (H_i + H_s) = \varepsilon \partial_t (E_i + E_s) + \sigma (E_i + E_s). \quad (3.31)
\]

The equations of the incident and scattered field are

\[
\nabla \times E_i = -\mu_0 \partial_t H_i, \quad (3.32)
\]
\[
\nabla \times H_i = \varepsilon_0 \partial_t E_i, \quad (3.33)
\]
\[
\nabla \times E_s = -\mu_0 \partial_t H_s, \quad (3.34)
\]
\[
\nabla \times H_s = \varepsilon \partial_t E_s + (\varepsilon - \varepsilon_0) \partial_t E_i + \sigma (E_i + E_s). \quad (3.35)
\]

The scattered field can be re-written as

\[
H_s^{n+\frac{1}{2}} = C_1 H_s^{n-\frac{1}{2}} - C_2 (\nabla \times E_s)^n, \quad (3.36)
\]
\[
E_s^n = C_1 E_s^{n-1} + C_2 (\nabla \times H_s)^{n-\frac{1}{2}} - C_2 (\nabla \times H_s)^{n+\frac{1}{2}}. \quad (3.37)
\]

The scattered field, the total field minus the incident field, is calculated by using FDTD. The total field can be obtained by adding the analytical incident field to the calculated scattered field.

3.3.3 Lumped impedance

Lumped impedance (resistor, capacitor and inductor) in FDTD algorithm can be implemented by adding a lumped electric current density term $J_L$ [23] to the right hand side of equation 3.14

\[
\nabla \times H = \varepsilon \partial_t E + \sigma E + J_L, \quad (3.38)
\]

and in discrete form at $t = (n + \frac{1}{2}) \Delta t$ it will be

\[
\frac{\varepsilon}{\Delta t} (E^{n+1} - E^n) + \frac{\sigma}{2} (E^{n+1} + E^n) + J_L^{n+\frac{1}{2}} = (\nabla \times H)^{n+\frac{1}{2}}. \quad (3.39)
\]
Consider the case of a z-directed lumped element located in free-space, between the grid points \((i, j, k)\) and \((i, j, k + 1)\), i.e. at the field component \(E_z(i, j, k)\). Assuming that the lumped current \(I_{Lz}\) is uniformly distributed, the corresponding current density \(J_{Lz}\) that flows between the points \((i, j, k)\) and \((i, j, k + 1)\) is given by

\[
J_{Lz} = \frac{I_{Lz}}{\Delta x \Delta y}.
\]  

(3.40)

The positive direction of the lumped current \(I_{Lz}\) is the same as it is for the \(E_z\) field component. \(I_{Lz}\) is related to the voltage \(V_s\), between the points \((i, j, k)\) and \((i, j, k + 1)\), through the I-V relationship of the lumped element to be modelled, which can be written in a general form as \(I_{Lz} = f(V_s)\). Assuming that \(E_z\) is constant along the side of the cell between \((i, j, k)\) and \((i, j, k + 1)\) and equal to \(E_z\), the voltage change across the side of the cell can be written as

\[
V_z(i, j, k) = - \Delta z E_z(i, j, k).
\]  

(3.41)

The I-V characteristic of z-directed resistor \(R\), connected between the nodes \((i, j, k)\) and \((i, j, k + 1)\) is given by Ohm’s law

\[
I_{Lz}^{n+\frac{1}{2}}(i, j, k) = \frac{V_z^{n+\frac{1}{2}}}{R}.
\]  

(3.42)

The voltage is related to the electric field \(E_z\) by

\[
V_z^{n+\frac{1}{2}} = - \Delta z E_z^{n+\frac{1}{2}}(i, j, k) + E_z^{n-\frac{1}{2}}(i, j, k) / 2.
\]  

(3.43)

According to the preceding, the lumped current density can be determined by the resistor value, the cell size \((\Delta x, \Delta y, \Delta z)\) depending on resistor direction and the electric field in the cell.

### 3.3.4 Stability

To ensure accuracy of the computed results, the cell size must be taken as a small fraction (e.g. 0.1 \(\lambda\)) of either the expected minimum wavelength or the minimum scatterer dimensions. Once the cell dimensions have been chosen, the stability of the time stepping algorithm in FDTD scheme is ensured by the Courant criterion equation 3.44 [32].

\[
\Delta t_{max} \leq \frac{1}{c \sqrt{\frac{1}{\Delta x^2} + \frac{1}{\Delta y^2} + \frac{1}{\Delta z^2}}},
\]  

(3.44)

in which \(c = c_0(\varepsilon_r \mu_r)^{-\frac{1}{2}}\) denotes the maximum wave phase velocity. In most cases \(c = c_0\) as the region outside the object is usually chosen to be free-space (\(\varepsilon_r = \mu_r = 1\)). The maximum time step is \(\Delta t_{max}\) and \(\Delta x, \Delta y\) and \(\Delta z\) are the cell dimensions. This criterion is only valid when the target region consists of loss-less, locally homogeneous material.
3.3.5 Absorbing Boundary Conditions

As noted previously the computational FDTD domain is divided into Yee cells. For an efficient computation, the size of the computational domain is restricted by the memory and CPU-speed of the computer. The main problem that arises from the bounded calculation space is the calculation of the field at the boundary. In most cases the entire exterior of the FDTD calculation domain is taken to be free space. This means in practice that waves travelling towards a boundary should not be reflected at that boundary. In literature some methods are described on how to handle these boundaries. One of these methods uses the Absorbing Boundary Conditions (ABC) developed by Mur [25]. This method is implemented in EMFDTD 1.4. Better performing methods however, for example Berenger's Perfect Matched Layers (PML), exist. An important future step to gain accuracy of the results might be implementing PMLs.

3.3.6 Gaussian Pulse

For all FDTD simulation in this report a Gaussian pulse is used as plane wave excitation, see also Figure 3.5. The advantage of a gaussian pulse is that its shape remains the same after a Fourier transform. The time domain function is defined as

\[ E(t) = \exp \left[ - \left( \frac{t - t_1}{\tau} \right)^2 \right], \tag{3.45} \]

in which \( \tau \) corresponds to the pulse duration and \( t_1 \) the time instant needed for the amplitude to reach its maximum. The frequency-domain pulse is given by

\[ E(\omega) = \tau \sqrt{\pi} \exp \left( -\left( \frac{\omega \tau}{2} \right)^2 + i\omega t_1 \right). \tag{3.46} \]

The unit of \( E(\omega) \) in the frequency domain is density [V/m] / Hz.

Figure 3.5: the shape of the Gaussian pulse in the time and frequency domain.

3.4 Discrete Fourier Transform

The output of FDTD is the amplitude of the electric field (voltage) in discrete time domain steps. A Fourier Transform (FT) algorithm can be used to calculate the various
frequency components and the amplitude of the electric field (voltage). The continuous Fourier transform pair is given below.

\[
F(\omega) = \int_{-\infty}^{\infty} F(t) \exp(-j\omega t) dt, \quad (3.47)
\]

\[
F(t) = \frac{1}{2\pi} \int_{-\infty}^{\infty} F(\omega) \exp(j\omega t) d\omega. \quad (3.48)
\]

The \(E\) and \(H\)-field are calculated at \(t = n\Delta t\). So at lumped elements the associated current and the voltage can be calculated at certain points \(n\Delta t\) in the time domain only, i.e., the current and the voltage are "sampled" in time. At this stage the frequency components can be derived from the time samples using a Discrete Fourier Transform (DFT). Initially a function \(F(t)\), which is sampled according to \(F(t) = F(n\Delta t)\) is defined. Here \(n = 1, ..., N\), in which \(N\) is the total number of time samples.

The Discrete Fourier Transform (DFT) can be calculated using Equation 3.49 and yields a sampled signal \(F(m\Delta\omega)\), \(m = 1, ..., M\), in which \(M\) is the number of samples in the frequency domain. With \(\Delta\omega\Delta t = 2\pi/N\), the DFT can be defined as:

\[
F(m) = \sum_{n=1}^{N} F(n) W_N^{(n-1)(m-1)}, \quad (3.49)
\]

\[
F(n) = \frac{1}{N} \sum_{m=1}^{N} F(m) W_N^{(n-1)(m-1)}, \quad (3.50)
\]

where

\[
W_N = \exp\left(-\frac{2\pi j}{N}\right), \quad (3.51)
\]

in which \(n = 1, ..., N\) and \(m = 1, ..., N\).

When a Fourier Transform is used, care needs to be taken to choose a sample frequency, which is more than twice the highest frequency in the source signal, i.e. over two sample points per wavelength. This is known as Nyquist's criterion.

### 3.5 Test box

Before starting the FDTD calculations on the enclosure with apertures, shown in Figure 3.1, the algorithm of EMFDTD 1.4 is tested with a plane wave incident on a small box of \(11 \times 11 \times 11\) in a total FDTD space of \(31 \times 31 \times 31\) elements. The box has an empty space inner space of \(9 \times 9 \times 9\) elements. In Figure 3.6 the test box is shown. The
dimensions of one cell are $2 \times 2 \times 2$ mm. The time step from the stability criterium is automatically set by EMFDTD to 3.85167 ps. The boundary absorbing conditions are 'second order Mur' [25].

Figure 3.6: geometry testbox with grid locations.

Figure 3.7: gaussian plane-wave.

3.5.1 Completely closed box

The first test is to illuminate the box, without the apertures, with a Gaussian plane wave of 100 V/m peak as shown in Figure 3.7. For this test we take 501 time steps. In Figure 3.8 the results are presented. The first graph in Figure 3.8(a) shows the field in the middle of the calculation space, without presence of the box. Figure 3.8(b) gives the field in the center of the massive PEC box. The electric field is zero to numerical accuracy. The last three graphs at the bottom show the field in the case the box is filled with air. The resonance frequency is approximately 9.6 GHz. Increasing the thickness of the walls on the outside to a total of three cells makes no difference, (see Figure 3.8(d)), but increasing the thickness on the inside of the wall (see Figure 3.8(e)) increases the frequency.

Inside the box the total field inside must be 'zero', which means that the scattered field must be equal to $E_s = -E_i$. In this situation the calculated field inside the box has
an amplitude of approximately 1.5 V/m. This is 36 dB below the value of the incident wave. The level of the field in the box is constant; there is no attenuation. The same exercise is done with TNO FEL-FDTD software. The results where the same when using a plane-wave. When using a point-source excitation, there was no problem, the field inside was numerically 'zero'. Probably there is an accuracy problem with the scattering field FDTD formulation or implementation. A possible solution to avoid this problem is to divide the FDTD space into two FDTD spaces as presented in Figure 3.9. Another way to avoid the problem is to calculate the field distribution over the apertures and use this as excitation in FDTD. Further investigation of this error and the impact is recommended but lies outside the scope of this project.

3.5.2 Aperture in the box

The next evaluate step is to determine the amount of empty cells around the box. The less cells needed, the shorter the calculating time. For this test the box has an aperture of 5 x 3 cells.

The number of cells on the front of the box is set to 10, to avoid reflected signals with the boundary that could be illuminating the aperture. In Figure 3.10 the $E_x$-field in the center of the box is shown to evaluate the influence of the number of cells around the box. The calculation time of the different test configurations is presented in Table 3.1. The field in the center of the box is quite the same when using 5 and 10 cells. With 2 cells a small difference is observed. While using 1 cell there is a large error. Instability is the result of placing an object too close to the absorbing boundary.

Table 3.1: calculating time needed for 5000 time steps on a PENTIUM 4, 2600 MHz, 1024MB.

<table>
<thead>
<tr>
<th>Cells</th>
<th>Time</th>
</tr>
</thead>
<tbody>
<tr>
<td>10 cells in front, 10 around (31 x 31 x 31)</td>
<td>88 s</td>
</tr>
<tr>
<td>10 cells in front, 5 around (21 x 31 x 21)</td>
<td>12 s</td>
</tr>
<tr>
<td>10 cells in front, 2 around (15 x 31 x 15)</td>
<td>4 s</td>
</tr>
<tr>
<td>10 cells in front, 1 around (13 x 31 x 13)</td>
<td>3 s</td>
</tr>
</tbody>
</table>

3.6 Simulation results of the enclosure with aperture

In this section the results of the FDTD simulation of the enclosure, which was shown in Figure 3.1 are presented. The fields inside the enclosure are calculated. Furthermore, the induced voltage over the 50 $\Omega$ load, terminating a monopole sensor with a length of 1 and 2 cm is calculated, and subsequently compared to measurements.
Figure 3.8: closed test box, field strength $E_z$ at the center of the FDTD space.
Figure 3.9: Possible solution to avoid the FDTD scattering problem.

An FDTD model of the enclosure with a cell size of 2 mm has been used. Hereby a reasonable number of cells is obtained to include the apertures. For a resolution of 10 cells per wavelength, the maximum frequency is 15 GHz. This is large enough to account for the maximum frequency of interest of 5 GHz. The perfectly conducting walls of the enclosure are chosen to be one cell thick. As a result of Chapter 3.5.2, 5 empty cells are placed between the walls of the enclosure and the computational space and at the apertures side 10 empty cells are added. A monopole sensor is located at the centre of the enclosure. The terminator is modelled as one cell with a resistance of 50 Ω. The total computational space is $232 \times 162 \times 33$ cells (in total $1240272$ cells). The absorbing boundary layer is modelled by the second order Mur equations. In Figure 3.11 a drawing of the enclosure with the monopole is shown.

A total of 100 000 time steps are performed, which limits the minimum frequency to $t_{\text{max}}^{-1} \approx 5$ MHz. In Chapter 3.6.1 the results of the simulated field strength in the middle of the enclosure, without the monopole sensor, are presented. In Chapter 3.6.2 the computed voltage over the terminating resistor of the monopole sensor is given.

3.6.1 Field-strength inside the enclosure

Figure 3.12 presents the results of the calculated field strength, without the monopole sensor, at several positions in the enclosure. The graphs on the top denote the field strength, given in time domain. The frequency domain is obtained by using a DFT, and is presented in the graphs in the middle. At the bottom the transfer-ratios are shown. The transfer-ratio is calculated by division of the electric field in the inside and the incident field $E_i$ in frequency domain. The resonance frequencies are visible. The amplitude of the field is strongly dependent on the position inside the enclosure, see Figure 3.12(b). Although the incident plane-wave has a $E_z$ component only, scattering on the edge of the apertures results in a relatively small $E_x$ and $E_y$ field, starting form about 3.4 GHz. This is caused by the height of the enclosure. After the plane-wave
Figure 3.10: field strength $E_z$ at the center of the FDTD space. In above situations the number of empty cells at the front-side of the object is constant at 10 cells.
passes the enclosure there is no longer an external source. Therefore, the field strength in the enclosure will decay to zero. The field in the enclosure will be reradiated through the apertures.
After 100000 time steps the field is not yet zero. By applying DFT there will be some aliasing in the high frequency range. This is, in this situation, no problem because the frequency of interest is limited to 5 GHz. Another result of 'cutting' the time signal is that the amplitude in the frequency domain has not reached the final value.

3.6.2 Monopole sensor
Two simulations are done with a monopole, which has a length of 1 cm (5 cells) and 2 cm (10 cells) respectively. Both have a thickness of 2 mm (1 cell). Figure 3.13 presents the results of the calculated voltage over the 50 \( \Omega \) terminator. Note that the voltage levels of the 2 cm monopole are about twice the level of the 1 cm version. The voltage transfer-ratio of the monopole difference is not a factor two higher for all frequencies. The reason for this is that the field distribution along the height of the enclosure is not uniform at higher modes.

3.7 Conclusions of simulations
When using plane-wave excitation it has been recognized that numerical errors were present with two FDTD software packages. The field in a completely closed PEC box filled with air is not equal to 'zero'.

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Figure 3.12: electric field strength.

(a) middle of enclosure, (b) $E_z$ at several locations.

Figure 3.13: Monopole antenna voltage.

(a) Monopole 1 cm (b) Monopole 2 cm
The scattering at the edges of the apertures results in electric field components in the \( x \) and \( y \) directions. When comparing the resonance frequencies of the analytical solutions of a closed box and the simulation with the EMFDTD software, one observes that the calculated frequencies with FDTD are approximately 5\% lower in frequency, (see Table 4.1 in Chapter 4.3). In the graphs there is no information about the modes of the resonance frequencies. The fact that there is a field in the middle of the enclosure when the integers \( m \), \( n \) and \( p \) are odd is used to determine the modes of the responses. To verify if errors have been made in the FDTD program, it is verified that the speed of light is correctly implemented in the EMFDTD software. This is illustrated in Figure 3.14.

The resonance frequencies in the middle of the enclosure of the field, detected with the 1 cm monopole, are equal to the resonance frequencies without the monopole. Some frequencies are a bit lower with the 2 cm monopole, which is probably caused by antenna and casing interaction. The detected levels are higher, caused by the larger effective antenna area.

The FDTD simulation is a time consuming process. A general simulation of resonant structures takes 83 hours for 100 000 steps. To reduce the simulation time, smarter approaches are possible. The number of time steps can be reduced by applying Prony extrapolation [17] [21] for example. The number of cells can be reduced by applying different cell sizes, depending on variations in the structure (and maximum frequency).
(a) straight ahead propagation time needed for a gaussian plane-wave in an empty space. The FDTD space is $11 \times 501 \times 11$ elements, one cell is 2 mm square. In total 1000 time steps are taken, one step is $3.85167 \cdot 10^{-12}$. The propagation time from the beginning to the end is $3.33551 \cdot 10^{-9}$. Over a length of 1 m, the speed is $299.8 \cdot 10^6$ m/s.

(b) diagonal propagation time needed for a gaussian plane-wave in an empty space. The FDTD space is $355 \times 355 \times 11$ elements, one cell is 2 mm square. In total 1000 time steps are taken, one step is $3.85167 \cdot 10^{-12}$. The propagation time from the beginning to the end is $3.45492 \cdot 10^{-9}$. Over a length of 1.0001 m, the speed is $299.8 \cdot 10^6$ m/s.

Figure 3.14: numerical FDTD propagation time.
Chapter 4

Measurements

To verify the numerical results of Chapter 3, measurements in frequency (narrow-band) and time domain (broadband) have been performed. Therefore, a copy of the conducting enclosure, which was described in Chapter 3.1, is constructed of aluminum (thickness of 3 mm). All of the side joints were soldered together. The dimensions of the enclosure are scaled up two times to ensure measurement accuracies as reasonable as possible. As a consequence all (resonance) frequencies will be half of the values determined in the simulation. A wire probe (monopole) with a 1.78 mm diameter of 2 cm (one quarter of the height) or 4 cm lengths was mounted, on a N-connector, at the bottom of the box to measure the electric field inside the box. To determine the frequency response of the monopoles, a ground-plane is constructed with the same dimensions as the enclosure.

4.1 Frequency domain

4.1.1 Setup

The measurements in the frequency domain were performed in a full anechoic chamber of the Center for Technology and Mission support of the Royal Netherlands Army. The dimensions of this chamber are 7 x 4 x 4 m. The walls, including the floor, are covered with ferrite tiles (30 MHz - 1000 MHz) and partly with absorbing cones (1 GHz - 18 GHz).

The measurements are performed from 200 MHz to 1800 MHz (400 MHz to 3600 MHz for comparison with the simulation) with a EMI-analyzer. To achieve reasonable dynamic range the resolution bandwidth was set at 100 kHz with corresponding step-size of 100 kHz, the measurement detector was set on RMS (Root Mean Square). The output power of the tracking-generator was set to the maximum value (≈ -5 dBm).

To minimize the impact of the impedance mismatch at the wire probe, a 10 dB attenuator was placed just after the wire probe or field measurement antenna. For a reasonable impedance match (50 Ω) at the EMI-receiver, the internal attenuator at the input was set at 10 dB. The transmitting antenna is a broadband horn antenna (EMCO 3106)
and is vertically polarized. A horn is used instead of a logperiodic antenna because of the relatively flat and smooth frequency response. In LabView®, a program is written to read the compressed data (from 16000 to 920 points) of the EMI-receiver. The enclosure or monopole is placed on a turntable at a height of 1.10 m and 3 m from the transmitting antenna. The measurement setup is presented in Figure 4.1.

![Figure 4.1: measurement setup frequency domain.](image)

(a) enclosure and horn antenna. (b) monopole on a ground-plane.

**Figure 4.2: pictures of the measurement setup in the anechoic chamber.**

### 4.1.2 Field strength measurement

The goal of the first measurement is to determine the field strength at the location of the enclosure. Therefore the enclosure was replaced by a small biconical antenna (PBA10200). The free-space 3 m distance antenna-factors (dB/m) were derived from the calibration-report [1]. The measured field strength is in the range of 90 - 98 dBµV/m \((\approx 32 - 79 \text{ mV/m})\) and is shown in Figure 4.3. The relation between the electric field, measurement level and antenna-factor is
\[ E[ dBV/m] = U[ dBV] + AF[ dB/m] \]  

(4.1)

Figure 4.3: field strength at a distance of 3 m (exclusive cable losses).

4.1.3 Monopole sensor on a ground-plane

A monopole wire sensor inside the enclosure will be used for field measurements inside. Before this can be done it is necessary to know the antenna factor of the monopole in order to calculate the field strength inside the enclosure from the measured voltage. The setup is the same as in Figure 4.1. A picture of the setup is shown in Figure 4.2(b). Only the enclosure is replaced by the monopole sensor on the ground-plane. The antenna factor can be calculated by subtracting the field strength (Figure 4.3) from the measured voltage. The results of the 2 cm and 4 cm monopoles are presented in Figure 4.4.

(a) measured level (exclusive cable losses), (b) antenna factor,

Figure 4.4: monopole antennas.
4.1.4 Completely closed enclosure

The first measurement with the enclosure is to verify the fidelity of the enclosure, connected attenuator and cable. For this, the apertures were covered with copper tape. Only noise was measured, as expected.

4.1.5 Enclosure with apertures

Now we know for certain that the enclosure is constructed properly, we can measure the effect of the apertures. The setup is shown in Figure 4.1 in which the apertures are facing the transmitting antenna. The measured voltage levels of both monopoles are shown in Figure 4.5.

![Figure 4.5: measured voltage at the terminals of the monopole antennas in the enclosure (exclusive cable losses).](image)

Figure 4.5 displays the electric field transfer ratio of the 2 cm and a 4 cm monopoles. The amplitudes of the resonance frequencies measured with both monopoles do not have the same level. It is noted that the monopole exhibit some coupling with the enclosure, which changes the previously determined antenna factor. Some resonance frequencies measured by the 4 cm monopole are a few MHz lower as can be seen in Table 4.1. The 4 cm monopole has more coupling with the enclosure [13]. Finally, it is recognized that some resonance frequencies are actually measured, while the associated cavity modes should be zero in the middle of the enclosure. This is caused by inaccuracies in placing the monopole some field is detected.

4.1.6 Azimuth incident angle

By rotating the enclosure, the effect of a changing incident angle on the amplitude and resonance frequencies is measured. The results of the different illumination angle of the incident plane-wave is presented in Figure 4.7. As expected resonances are uncorrelated to angle of incidence. It is remarkable that when the enclosure is illuminated at the back-side (180°), the measured levels are relatively close (3 dB - 12 dB) to the
4.2 Time domain

The measurements in the time domain were performed at the covered ground penetration radar (GPR) test-site at TNO-FEL. Normally, the site is used for mine detection research with Ultra Wide-band techniques.

4.2.1 Setup

The measurements are performed with a time domain pulse generated with a pulse generator mounted at the backside of the IRA. Figure 4.8 presents the setup. The 50 Ω output of the generator is connected to the antenna. The trigger-input is connected to output A (high voltage) of the pulse delay generator using a coaxial cable. The bias-tee and step-diode will enhance the leading or trailing edge pulse by a factor of 20. The monopole antenna of the enclosure is connected to a 10 dB attenuator and a coaxial cable which is connected to the input of the digital oscilloscope. An adjustable attenuator is placed in front of the input to protect the input for voltage levels higher than the maximum input level of ±1 V. During the measurement the attenuator is adjusted to get a maximum display readout. The trigger input is connected via a step-diode to output \( T_o \) of the pulse delay generator. The coax-cable between the monopole and digital oscilloscope is protected with ferrite to eliminate common-mode currents on the cable.

The pulse delay generator activates the pulse generator and trigger input of the digital oscilloscope. The pulse repetition rate (PRF) is set to 800 Hz.
Figure 4.7: different angles of incident plane-wave, measured with a 2 cm monopole.
The digital oscilloscope has a resolution of 8 bits, therefore the obtainable signal to noise ratio (s/n) amounts to 48 dB. There is a delay between the time that the pulse leaves output $T_o$ and the receiving of the measured signal on the display of the digital oscilloscope. This delay is a summation of all delay times in the circuit. The total delay is

$$td_l + td_{pulse} + td_{wire} + td_{air} + td_r + td_{scope} - td_{trig} \approx 133\text{ns}. \quad (4.2)$$

The delays of the attenuators, bias-tee and step-diode are unknown. For the first measurements the delay time of the digital oscilloscope will be set at 133 ns and the actual unknown delay will be found by measurements.

By using the averaging function of the digital oscilloscope the signal-to-noise ratio will increase.

### 4.2.2 Output signal pulse generator

The original plan was to measure the electromagnetic field strength in the time domain with a B-dot sensor. Unfortunately the sensor turned out to be defective and was not available within a reasonable time. To get an idea of the frequency contents of the electromagnetic field, the output signal of the pulse generator is measured with the digital oscilloscope by using attenuators, in total 86 dB. The time domain signal and its DFT are presented in Figure 4.9. The rise time of the pulse is approximate 200 ps and the duration is 4 ns. The maximum useful frequency is around 2 GHz.

### 4.2.3 Monopole on Ground-plane

To obtain an idea of the time and frequency response of the monopole wire sensor inside the enclosure a measurement with the monopole on a ground-plane has been performed. The setup is the same as in Figure 4.8, only, instead of the enclosure, the monopole sensor is placed on the ground-plane. The results of the 2 cm and also a 4 cm monopole are presented in Figure 4.12. The graphs show that the starting point of the received signal is around 133 ns, as expected. Between the brackets marks the attenuation setting before the input is given. The averaging function of the digital oscilloscope is set at 16. This means that the result is the average over the last 16 measurements. Unfortunately, the antenna-factor (AF) and thus the field transfer-ratio in the enclosure could not been calculated because it was not possible to measure the field strength at the ground-plane.

### 4.2.4 Completely closed enclosure

The first measurement with the enclosure is performed to verify the fidelity of the setup. For this, the apertures where covered with copper tape.

With the attenuator set at 0 dB, a very small signal is measured and is presented in Figure 4.11. The voltage level $< 0.01$ V. The voltage density $< 1.5 \cdot 10^{-10}$ V/Hz and has a average $< 0.5 \cdot 10^{-10}$ V/Hz.
distance \(d = 6\) m

\[s = 3.33 \text{ ns/m}\]

\[t_{\text{air}} = 30\text{ ms}\]

\[r_{\text{enclosure}} = 12\text{ m}\]

\[t_{\text{I1}} = 49.4\text{ ns}\]

\[I = 8\text{ m}\]

\[t_{d_{\text{td}}} = 32.9\text{ ns}\]

\[r_{\text{fen-He}} = 0.8\text{ m}\]

\[C_{\text{trig}} = 4\text{ ns}\]

(a) setup.

(b) front side.

(c) back side.

Figure 4.8: IRA measurement setup.
4.2.5 Enclosure with apertures

The results of the time domain measurement on the enclosure with apertures are presented in Figure 4.12. To improve the signal-to-noise ratio, the average function of the digital oscilloscope is increased to 64. As expected the level of the time signal goes to zero exponentially. The voltage level of the first resonance frequency for example measured with the 2 cm monopole is in the noise. The results with the 4 cm monopole are better, due the fact that the induced signal is higher.

4.2.6 Reflection measurement with second IRA

The signal in the enclosure attenuates gradually. This is caused by the losses due to the finite conductivity (aluminium). Also, the signal will re-radiated through the aperture at the resonance frequencies of the enclosure and if possible will be measurable with a second receiving IRA.

To test this, the coax cable connected to the monopole is now connected to the receiving IRA. The setup is shown in Figure 4.13.

Firstly, the receiving IRA will receive the direct signal from the transmitting IRA and secondly the reflection of the enclosure at a distance of 6 m. Both antennas are vertically polarized. As expected, the amplitude of the direct signal is much larger than the reflection. To protect the input of the digital oscilloscope, the external attenuators in front of the input are maximized. Figure 4.15 shows the measured signal. The direct pulse is measured at 121 ns and reflection at 156 ns. The circuit time, including both IRAs, is 112 ns. Time delay between both IRA's is thus 9 ns (=2.6 m) and the reflection by the enclosure is delayed 44 ns =(12.6 m). This is of course the summation of the distance of the enclosure between the transmitting and the receiving IRA, both
approximately 6 m.

The resolution of the reflected signal is very low compared to the direct signal. To solve this problem a coax switch is placed in front of the input, which connects the input after the direct pulse has past. The control input of the switch is connected to output \( B \) of the pulse delay generator, which controls the switch. The setup is presented in Figure 4.14.

Figure 4.16 shows the measured reflected signal with the enclosure placed at 6 m distance in front of the antennas. In the situation without the enclosure reflection of the metal structure placed at the left site of the test range is detected. This fixed structure is somewhat visible in Figure 4.8(c).

The level of reflection is too high to recognize the resonance frequencies directly. A simple improvement is made by subtracting the measurement without enclosure from the one with enclosure. The result is presented in Figure 4.17. Peaks are visible (around 345, 565 and 855 MHz). These could correspond to resonance frequencies of the enclosure, but this is too little to validate this claim.

TNO-FEL has developed a detection algorithm for Ground Penetrating Radar (GPR) based on autoregressive modelling to determine mine signature. This algorithm is applied to the measurement data. Results can be found in Appendix C.

### 4.3 Conclusions of measurements

The results of the detected resonance frequencies of the enclosure of the frequency and time domain measurements are listed in Table 4.1. The frequencies are scaled up...
two times for comparison to the simulation results. The detected frequencies with both methods are close to each other. Some frequencies measured with the largest monopole are a few MHz lower. Probably this monopole has induces coupling to the enclosure.

The frequency measurement technique (EMI-receiver) is time consuming but has a high sensitivity. As a consequence more resonance frequencies are detected. For the time domain measurement this is opposite.

The electric field transfer ratio measured in frequency domain of the field on the outside to the field at the resonance frequencies goes up 30 dB. This leads to the conclusion that the field inside the enclosure is significantly higher than outside the enclosure at discrete frequencies. The measurements confirm that the resonance frequencies are not correlated to the angle of incidence. Illumination of the enclosure from the back-side shows that the measured levels are relatively close to the case of front-side illumination.

In the time domain, the results show that field is 'ringing' in the enclosure and attenuates gradually. The measured levels with the largest monopole are the highest and result in a better signal to noise ratio, because of a higher antenna efficiency.

The reflection of the enclosure is detectable with the reflection setup using both IRA, although there is a lot of reflection 'noise' of other structures close to the setup. A simple improvement is made by subtracting the measurement without enclosure from the one with enclosure. In the result, peaks are noticed that could be resonance frequencies of the enclosure, but there is too little information to prove this claim. The GPR detection algorithm gives some preliminary promising results.

Figure 4.11: voltage 4 cm monopole antenna, closed enclosure (attenuator 0 dB).
(a) monopole 2 cm (attenuator 3 dB).
(b) monopole 4 cm (attenuator 7 dB).

Figure 4.12: monopole antenna in enclosure with apertures (relative measurement).

Figure 4.13: reflection measurement setup.
Figure 4.14: IRA measurement setup.

(a) without enclosure (attenuator 27 dB). (b) with enclosure (attenuator 27 dB).

Figure 4.15: received direct signal from transmitting IRA and reflection of the enclosure at a distance of 6 meter (relative measurement).

(a) without enclosure. (b) with enclosure.

Figure 4.16: enclosure at a distance of 6 meter (relative measurement).
Figure 4.17: subtracting time domain of 4.15(b) - 4.15(a).
Table 4.1: overview of the resonance frequencies up to 2 GHz, the locations of \( \text{mnp} \) are shown in Figure 4.18.

<table>
<thead>
<tr>
<th>Mode</th>
<th>Analytical MHz</th>
<th>FDTD field monopole</th>
<th>Frequency monopole short MHz</th>
<th>Time monopole short MHz</th>
</tr>
</thead>
<tbody>
<tr>
<td>( mnp )</td>
<td></td>
<td>MHz</td>
<td>MHz</td>
<td>MHz</td>
</tr>
<tr>
<td>101</td>
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<td>201</td>
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<td>-</td>
</tr>
<tr>
<td>102</td>
<td>1089</td>
<td>-</td>
<td>-</td>
<td>-</td>
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<td>-</td>
<td>-</td>
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<td>-</td>
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<td>1458</td>
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<td>-</td>
<td>-</td>
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<td>402</td>
<td>1712</td>
<td>-</td>
<td>-</td>
<td>-</td>
</tr>
<tr>
<td>501</td>
<td>1781</td>
<td>1688</td>
<td>1688</td>
<td>1688</td>
</tr>
<tr>
<td>303</td>
<td>1858</td>
<td>1758</td>
<td>1758</td>
<td>1758</td>
</tr>
<tr>
<td>502</td>
<td>1994</td>
<td>-</td>
<td>-</td>
<td>-</td>
</tr>
</tbody>
</table>

Figure 4.18: locations of \( \text{mnp} \).
Chapter 5

Conclusions

Highpower microwave systems use sources of high power microwave energy ranging from kilowatts to gigawatts. It is established that sufficiently intense electromagnetic fields can cause damage proportional in many electronic systems. In general, the typical frequency range is from 200 MHz up to 5 GHz. When both the HPM antenna and the target are at ground level, the reflection of the field at the ground results in a height dependent level pattern. As a consequence, systems will be vulnerable within a distance of hundreds of meters.

High power microwave energy can penetrate electronic systems through the "Front door" (antennas) or the "Back door" (cables and apertures). Permanent damage to electronics is caused by thermal effects or electrical breakdown.

In order to study coupling into systems, appropriate models are used to model the housing of a real piece of 19” IT equipment to determine the resonance frequencies. The housing is modeled as an enclosure with two small apertures. The frequencies which are calculated analytically and defined in the frequency and time domain measurements are very close (< 1.5 %). The resonance frequencies calculated by using FDTD are a little bit lower (5 %).

It has been recognize that numerical errors are present in FDTD (scattered field formulation) results for the case of plane wave incidence. In the situation of a closed box, the calculated field inside the box has an amplitude of approximately 1.5 V/m. This is 36 dB below the value of the incident wave.

It is observed that the amplitude (of the resonance frequencies) depends on the location in the enclosure. The simulation is a time consuming process. A general simulation of resonant structures takes 83 hours. However, FDTD shows to be a suitable tool for complex structures for which analytical answers are absent. Analytical approaches are useful for simple structures but can be used for a quick analysis.

Although the apertures are very small compared to the dimensions of the cavity, the amplitudes of lower frequency cavity resonances are relatively high.

FDTD simulation of the field inside the enclosure and measurements on the enclosure shown that the electric field at discrete frequencies is significant higher than outside the (empty) enclosure. This field enhancement amounts to 30 dB maximum.
It is recognized that some resonance frequencies are actually measured, while the associated cavity modes should be zero in the middle of the enclosure. This is caused by inaccuracies in placing the monopole some field is detected.

The most accurate way to measure the resonance frequencies is performed with a monopole with a length of at least less than 1/4 of the enclosure height. The measurements confirm that the resonance frequencies are not correlated to the angle of incidence. Illumination of the enclosure from the back-side shows that the measured levels are relatively close to the case of front-side illumination.

The used frequency measurement technique (EMI-receiver) is time consuming but has a high sensitivity, so more resonance frequencies are detected. For the time measurement (oscilloscope) with an Impulse Radiating Antenna this is opposite.

An existing system, on which the model is based, is tested for susceptibility in the frequency range of 200 - 1000 MHz. It is remarkable that some frequencies of interference are lower (~ 15%) than the numerical and measured resonance frequencies. This is in contrast with the measurements on Personal Computers [22], in which it was comparable. A possible explanation is that equipment has a high density of printed circuit boards (PCB's) and components. The propagation speed through PCB's is less than through air, which can result in lower resonance frequencies. For more results the reader is referred to [19].

To obtain reliable results it has been shown that performing dedicate measurements leads to the best approach. However, numerical evaluations are important to estimate an appropriate test-plan for measurements.

The first results obtained from time domain measurements with the IRA indicate that this might be a promising tool for analyzing the resonance frequencies of complex systems. The results of the processed IRA reflection measurement data with GPR detection algorithm give some preliminary promising results.

Until now, the best measure to avoid disturbance of electronic systems is by applying the basic EMC design rules.
Chapter 6

Future research and recommendations

The effects of HPM on electronics may result in upset or, at high levels of irradiation, even permanent physical damage. Permanent damage is generally caused by thermal effects or electrical breakdown in the circuits. It can be interpreted that electronic is sensitive for the energy contents and/or peak value of the HPM field. To develop protection measures it is very important to know the sensitivity of electronic equipment for continuous wave variants (AM, PM) and pulsed signals with short pulse widths. It is advisable to test electronic systems with these kind of signals. When a trend is noted this can be used for setting up HPM test methods at appropriate field levels.

FDTD is widely used for determining shielding efficiency of structures. It has been noted that for plane wave excitation numerical errors were present with two FDTD software packages using the scattered field formulation. The field in a completely closed PEC box filled with air is not equal to 'zero'. Therefore, it is most advisable to analysis this. A possible solution to avoid this problem is to divide the FDTD space into two FDTD spaces. The outside space will be the scattering formulation an the inner space the total field formulation. Another way to avoid the problem is to calculate the field distribution over the apertures and use this as excitation in FDTD.

FDTD simulations have shown that the level of the resonance frequencies depends on the location in the enclosure. The magnitude of the induced currents and voltages in electronics depends on the frequency and level. Therefore a new approach of the problem is advisable using statistical electromagnetics [12].

The used FDTD algorithm is straightforward. The simulation is a time consuming process. The duration can be reduced by reducing the number of time steps by means of applying Prony's extrapolation method [17]. It has been shown that the relative deviation between the extrapolation and the FDTD results is less than 2% [21]. The time required will be reduced to 8 hours when computing 10,000 steps and extrapolating them to 100,000 by using Prony's method.

The number of cells can be also reduced by applying different cell sizes in the grid, depending on variations in the structure and the maximum frequency.

An important issue for research is to investigate the cause of the difference in the simulated and measured resonance frequencies as well as the lower frequencies of inter-
ference found with the real equipment. The first step can be to measure the resonance frequencies inside the equipment and compare those to the resonance frequencies in the (empty) enclosure. The next step could be to verify whether or not the system is more susceptible to those resonance frequencies than other frequencies.

The experience with the time domain measurements with the IRA antenna shows that this is a relatively fast method to obtain results. An future application of the time domain measurement with the IRA is to measure the attenuation of facilities (buildings, shielded rooms, etc.).

It is recommended to investigate the possibility of detecting the resonances frequencies by means of reflection measurements in combination with the GPR detection algorithm.
Appendix A

An short overview of electron beam sources

Below, an overview of electron beam sources is given. More detailed information can found in [30].

Klystron
The klystron is an evacuated electron tube used as a high-frequency oscillator or amplifier. In the Klystron, an electron beam is velocity modulated to produce large amounts of power up to tens of MW's.
In general, the klystron provides highly pulsed and continuous wave (CW) power with medium bandwidth limitations (2-15 %). Relatively high gain (up to 70 dB) and efficiencies, up to 70 %, can be achieved.
The main features are the electron, the collector, the input and output cavities and the beam accelerator potential as shown in Figure A.1(a).
Generally, the electron beam is formed from thermionic electrons accelerated by an anode to cathode voltage. The electron velocities are modulated as the electron beam passes the grid, which is connected to the input cavity. The different velocities produce variations in the electron number densities (beam intensity). This drives the second grid with periodic signals, which is rich in harmonics. The output cavity acts as a high Q resonant oscillator. The result is a sinusoidal output at the resonance frequency of the cavity.

Magnetron
The magnetron is a widely used device for power RF applications. It is more efficient than the Klystron; up to 80 %. Relativistic magnetrons have achieved efficiencies in the order of 30 % in the band from 500 MHz to 10 GHz. Conventional magnetrons are available up to several tens of kW's for CW signals to GW's (phase locking) for peak signals. In Figure A.1(b) an illustration of a multi cavity magnetron oscillator is presented. The features of the magnetron are the coaxial cathode, the multi cavity anode, an axial DC magnetic field and a circulating electronic cloud.

Super Reltron
The super reltron is a klystron-like device which can operate without the need of an externally applied magnetic field. High peak powers can be achieved without the need
of a very large magnet and, therefore, a bulky microwave system.

**Gyrotron**
The gyrotron is a relatively new type of microwave device, which is operated in the mm-band, normally between 30 to 300 GHz. The basic idea is to amplify or generate coherent electromagnetic radiation by radiative emission from a relativistic electron beam.

**Vircator**
The virtual cathode oscillator, or vircator, is a high-power source capable of operating within the frequency range of a few hundred megahertz to tens of gigahertz. This device operates only in the pulsed mode but is relatively broadband, tunable and has a very low efficiency. The advantage of a vircator is the very simple design. The most important disadvantage is the short lifetime. It is a typical one-shot device.

**Travelling wave tube (TWT)**
The Travelling-Wave Tube (TWT) is an amplifier of microwave energy. It accomplishes this through the interaction of an electron beam and an RF circuit, known as a
slow wave structure. The term "slow wave" is derived from the fact that the RF wave velocity as it travels down the circuit is much less than that of light in free space. As the electron beam travels down this interaction region, an energy exchange takes place between the particles and the RF circuit wave. As an example, if one was to apply five watts of RF energy to the input of a TWT RF circuit, they may find one hundred watts at the output RF terminal. TWT provides gain to the applied signal.

There are three basic components to any TWT, or linear beam, device. They are the electron gun, the slow wave circuit, and the collector. Any or all of these major components can range from a very complicated to a simple design. The choice is based upon performance requirements and customer specifications. Central to the operation of a linear beam device is the electron gun, from which the electron beam is generated. The source of the electrons is a component known as the cathode. This component is typically heated to a temperature in the range of 760 to 1100 degrees Celsius, depending on the nature of the cathode, and via thermionic emission and the application of a high voltage bias the electrons are drawn down the tube. This voltage, known as the cathode voltage, may range in value from several thousands to several hundreds of thousands of volts. There are numerous types of cathodes that have been used in microwave tubes since the early days of the magnetron. These tubes employed what are known as oxide coated cathodes. In this case a mix of carbonates, usually barium, strontium, and calcium, are sprayed onto a nickel surface, and once heated, break down into oxides and provide the source of electrons for the vacuum device. TWTs manufactured at the Williamsport operation utilize the oxide cathode and have been very successful in the field. Tubes with this particular source of electron beams have demonstrated life performances in excess of one hundred thousand hours of continuous operation.

The second major component of the TWT, or linear beam device, is the slow wave structure (SWS). This assembly can be compared to a bandpass filter in the classic microwave engineering sense. Over a particular band of frequencies, which can range as high as two or more octaves (one octave equals a doubling of frequency), the SWS supports the RF signal. There are numerous types of slow wave structures, helical, coupled-cavity, ring-and-bar and many other types in this class. The frequency at which the device operates controls the geometry, or size of the structure. Also RF power handling capabilities become important when selecting which type of SWS to use. The RF wave then travels down the SWS and an interaction, or energy exchange, takes place between it and the electron beam. One of the most important features of the SWS is
that it must control the velocity of the RF wave such that it matches that of the beam. This is a characteristic known as synchronism and is very important to the operation of the device. After the energy has been extracted to the circuit the beam enters a region, or assembly, known as the collector. The function of this device is exactly as the name implies and it collects the spent beam. There are various collector configurations used in linear beam devices. Some of these include single-stage grounded collectors and multiple stage collectors. The driving concept behind the selection of the used collector is efficiency and power supply considerations. For instance in the TV broadcast market where power consumption is a premium, multi-stage depressed collectors provide significant savings in energy. Some tubes with this type of collector configuration have reached basic electronic efficiencies of over 65 percent.
Appendix B

Design and construction of an IRA

The paraboloidal reflector antenna fed by a pyramidal horn has found wide-spread application in radar and communication engineering. The reflector antenna also has very useful characteristics when it is fed by two or four conductor transmission lines. A dispersion-less wideband antenna, with a nearly flat radiating spectrum, is desirable for short pulse applications. The reflector IRA employs a paraboloidal reflector fed by TEM lines [8, 34] and is an example of an aperture antenna. The radiated field from an aperture antenna consists of a spatial integration of the aperture fields over the aperture, while the temporal behavior of the aperture field is differentiated in the far field. In a practical situation, the illuminating field or the aperture field is a double exponential waveform and the radiated field then becomes impulse-like with a very large bandwidth ratio.

![Figure B.1](image_url)

Figure B.1: An illustration of a reflector fed by a pair of coplanar conical TEM lines

The reflector IRA under consideration consists of a paraboloidal reflector fed by two pairs of coplanar feed plates as illustrated in figure B.1. Coplanar feed plates are chosen over the more conventional facing-plate geometry to minimize the aperture blockage effects. To reduce the aperture blockage, the feed plates are required to be narrow,
which results in feed impedances of several hundreds of Ohms, the 400 \( \Omega \) lines results in a net feed impedance of 200 \( \Omega \). The aperture area should be as large as practically possible, since the far field is proportional to the square root of this area for a constant voltage at the feed. The magnitude of the far field is proportional to the aperture area for a constant aperture field. The pulse generator has to be of the differential type to avoid common mode DC currents on the feed plates, which could distort the desired features in the far field. The driving voltage is \( V(t) = \left( V_0/2 \right) u(t) \), where \( V_0 \) is generator output Voltage. Since \( E_{\text{far}} \propto \partial V/\partial t \), it is desirable to maximize this rate of rise of the incident field or the voltage pulse.

Two identical reflectors were manufactured with the following characteristics.

- one piece, spun aluminum, paraboloidal surface,
- diameter \( D = 900 \text{ mm} \),
- focal length \( F = 337.5 \text{ mm} \),
- profile accuracy \( \leq 1.5 \text{ mm} \),
- \( f_d = F/D = 0.375 \).

The desired rise time of 100 ps implies an upper 3 dB frequency of 3.5 GHz which gives rise to wavelengths in air of 85.7 mm. The surface tolerance of the reflector is small compared to the shortest wavelength and is therefore acceptable.

Next, we look at an estimation of boresight waveforms. For analysis purposes, one could consider a single (two conductor) coplanar feed, although in practice we used two such feed lines connected in parallel for a more uniform illumination of the reflector (see fig. B.1). When the reflector IRA was originally proposed [6], the boresight radiation was predicted to consist of a feed step followed by an impulse-like behaviour. It was also shown that the total area under these two parts of the radiated waveform (i.e. prepulse plus impulse) is zero. This means, that there is no DC component in the radiated waveform consisting of the prepulse and the impulse. Furthermore, this implies that the portion of the radiated waveform after the impulse must have a net zero area in itself. The post impulse portion consists of diffracted signals from the feed plate and the circular rim of the paraboloidal reflector. A more recent analysis [14] has extended this result by chronologically considering the various temporal elements of the boresight radiation, which is illustrated in figure B.2. Let us assume that the voltage pulse generator is switched on at \( t = 0 \), and the observer is at a distance \( r(=z) \) to the right of the focal point of the paraboloid. These temporal elements are:

- Prepulse: feedstep \( E_{y1}(r, t) \).
- Main pulse: impulse \( E_{y2}(r, t) \).
- Postpulse:
  - feed plate diffraction: \( E_{y3}(r, t) \) (actually this consists of two parts originating from the plate edge (3a) and the plate itself (3b)).
  - edge diffraction from circular rim of the reflector, \( E_{y4}(r, t) \).
- Entire pulse constraints: low-frequency dipole moment radiation and no radiation at zero frequency (DC).
The far field $\vec{E}_f(\mathbf{r}, t)$ is given by,

$$\vec{E}_f(\mathbf{r}, t) = \sum_{i=1}^{\text{area}} E_y(r, t),$$

where

$$E_y(r, t) = E_y(t) + \text{low-frequency radiation from dipole moments resulting in time-integral constraints on entire pulse.} \quad (B.1)$$

The results of eqn. B.1 are illustrated in fig. B.2. The time-integral constraints imposed by the low-frequency radiation are,

i) the complete first-time integral of the radiated waveform must be zero,

ii) the second-time integral must be proportional to the late-time dipole moments.

**Feed plates, terminating and matching**

The feedplates form a parallel combination of $2 \times 400 \Omega$ conical TEM lines, giving the IRA an input impedance of $200 \Omega$. Matching to the $50 \Omega$ pulse source is done by means of a balun. Suitable high-frequency, high-voltage, and high-impedance cables are difficult to find. Finally, a $100 \Omega$ flexible cable with a PTFE dielectric was chosen. The latter is bound to introduce some dispersion, however the cable can be replaced by a better one in the future. Termination of each feed plate at the dish edge consists of a net DC impedance of $200 \Omega$. The insulating spacer is a high density polyethylene slab, the dimensions of which are optimised experimentally. Capacitance of the slab is
trimmed by drilling holes (tuned with help of TDR measurements at the antenna terminal). The terminating resistor network, consists of a series-parallel network (5 by 5) of 1 W, 200 Ω carbon composite resistors of 10% tolerance. Thus total heat dissipation is not a problem. Since the length of the resistor chain is ≥10 cm, the voltage stand-off (in the order of 10 kV/cm) is not seen as a problem either.

**Far field of the IRA**

If an aperture antenna of diameter $D$ is illuminated by a CW field of frequency $f$, then the far field is determined by the distance $r$ such that,

$$r \geq \frac{2D^2}{\lambda},$$

which is obtained by requiring that $\Delta r \leq (\lambda/16)$ where $\Delta r$ is the path difference of the edge ray and the central ray from aperture to observer. For pulsed antennas one can define a clear time $t_c = \Delta r/c$ and require that $t_c < t_{\text{rise time}}$ of the incident pulse. It is seen that

$$t_c = \frac{1}{c} \left[ \sqrt{\frac{D^2}{4} + r^2} - r \right].$$

Requiring that $t_c < t_{\text{rise time}}$, it is observed from table B.1 that the observation point

<table>
<thead>
<tr>
<th>$r$ (in m)</th>
<th>$r + \Delta r$ (in m)</th>
<th>$\Delta r$ (in mm)</th>
<th>$t_c = \Delta r/c$ (in ps)</th>
</tr>
</thead>
<tbody>
<tr>
<td>1.0</td>
<td>1.096</td>
<td>96</td>
<td>320.0</td>
</tr>
<tr>
<td>2.0</td>
<td>2.050</td>
<td>50</td>
<td>166.6</td>
</tr>
<tr>
<td>2.5</td>
<td>2.540</td>
<td>40</td>
<td>133.3</td>
</tr>
<tr>
<td>3.0</td>
<td>3.033</td>
<td>33</td>
<td>110.0</td>
</tr>
<tr>
<td>3.5</td>
<td>3.529</td>
<td>29</td>
<td>96.6</td>
</tr>
<tr>
<td>4.0</td>
<td>4.025</td>
<td>25</td>
<td>83.3</td>
</tr>
<tr>
<td>5.0</td>
<td>5.020</td>
<td>20</td>
<td>66.6</td>
</tr>
<tr>
<td>8.0</td>
<td>8.012</td>
<td>12</td>
<td>40.0</td>
</tr>
<tr>
<td>10.0</td>
<td>10.010</td>
<td>10</td>
<td>33.3</td>
</tr>
</tbody>
</table>

has to be at a distance greater than 8 m, to be in the far field, by a comparison of the clear time with the rise time of 100 ps. Russian researchers [24] have worked out near and far field expressions of the E-field on the bore-sight of an IRA. Using these expressions, we have estimated the on-axis E-field at various distances in the near and far field of the antenna. Although 8 m is the minimum distance to be in the far field, the actual experiments of illuminating targets can be performed at shorter distances in the near field. The pulse is broader and has more low-frequency components in the near field. In fact, very near the antenna, for example at the feed point, the electric field is similar to the voltage pulse, which is a double exponential waveform. The differentiation occurs in the far field and the transformation from near to far field is gradual and not abrupt.
This transformation can be observed experimentally by measuring the electromagnetic field on-axis, as one moves away from the focal region. These values have been experimentally verified during initial measurements and compared favorably to the above table. Expressions for the radiated spectrum are available in literature, as mentioned earlier. Calculations for the TNO-FEL IRA results in the values tabulated in table B.3. Measurements indicate that due to practical restrictions the actual performance is not quite as good as theoretically expected.

Table B.2: Near field and far field characteristics of the IRA (bore-sight)

<table>
<thead>
<tr>
<th>quantity</th>
<th>near field</th>
<th>far field</th>
</tr>
</thead>
<tbody>
<tr>
<td>distance $R$</td>
<td>5 m</td>
<td>10 m</td>
</tr>
<tr>
<td>clear time</td>
<td>66 ps</td>
<td>33.3 ps</td>
</tr>
<tr>
<td>prepulse</td>
<td>-255 V/m</td>
<td>-127 V/m</td>
</tr>
<tr>
<td>prepulse duration</td>
<td>2.25 ns</td>
<td>2.25 ns</td>
</tr>
<tr>
<td>impulse peak</td>
<td>5.9 kV/m</td>
<td>3.8 kV/m</td>
</tr>
<tr>
<td>impulse duration</td>
<td>85 ps</td>
<td>60 ps</td>
</tr>
</tbody>
</table>

Table B.3: Calculated bore-sight values for a distance $R = 10$ m

$E = 2.85 \cdot 10^{-7}$ V/m (+/- 12%)  
$f_{t} (3dB) = 80$ MHz  
$f_{u} = 10$ GHz  
BW-ratio = 125  
decades = 2.097

($E = magnitude \ of \ the \ electric \ field \ spectrum$)
Appendix C

Reflection measurement data processed with GPR detection algorithm

The detection algorithm is normally used for detection of buried mines with a ground penetrating radar (GPR). The algorithm is applied to the measurement data of the reflection measurement.

The technique has also been successfully employed in speech processing. The algorithm is based on autoregressive (AR) modelling [33] of the object (mine) signature. This AR model extracts the eigenmodes from the scattered signal of an object and determines the optimum number of poles describing the time series that represents the object signature. More information on this subject can be found in [28].

This algorithm is applied to the measurement reflection data. The input of this algorithm is the time signal and the minimum and maximum number of poles. In this case the minimum number is set on 1 and the maximum on 100 poles. The results of the raw time data of the measurement with and without the enclosure are shown in Figure C.1. The left graph show the time signals, the graph on the right are the roots of the AR-polynomial. At the bottom the frequency response of the time and the poles are presented. The optimal number of poles are chosen. It is noted that the positions of the poles are different. This means that there is a difference in the time signals. The maximum frequency of the DFT is about 10 GHz. However, the maximum useful frequency of the measurement falls around 2 GHz. From 2 GHz the measurement data is noise. The number of poles is used to fit the noise. An improvement can be made to process the data with a low-pass filter.

The results of the subtracted signal of Figure 4.17 is presented in Figure C.2. Three frequencies (375 MHz, 900 MHz and 1500 MHz) are detected as can be seen in the frequency domain graph.
Figure C.1: processed data with GPR algorithm.
Figure C.2: processed subtract signal data with GPR algorithm.
### Appendix D

#### Measurement equipment list

**Table D.1: Frequency domain measurements equipment**

<table>
<thead>
<tr>
<th>Description</th>
<th>Type</th>
<th>Manufactory</th>
<th>Serial number</th>
</tr>
</thead>
<tbody>
<tr>
<td>Anechoic chamber</td>
<td>FALC 3m</td>
<td>Comtest</td>
<td></td>
</tr>
<tr>
<td>EMI receiver</td>
<td>ESAI</td>
<td>R&amp;S</td>
<td>804.8910.52</td>
</tr>
<tr>
<td>Double Ridged Horn</td>
<td>3106</td>
<td>EMCO</td>
<td>9910-2760</td>
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<td>Biconical</td>
<td>PBA10200</td>
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<td>K193/02</td>
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<td>10dB attenuator</td>
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<td>Coax-cable 1m</td>
<td>-</td>
<td>-</td>
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<td>Coax-cable 2m</td>
<td>-</td>
<td>-</td>
<td>24142</td>
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</table>

**Table D.2: Time domain measurements equipment**

<table>
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<th>Description</th>
<th>Type</th>
<th>Manufactory</th>
<th>Serial number</th>
</tr>
</thead>
<tbody>
<tr>
<td>Oscilloscope</td>
<td>54750</td>
<td>Hewlett Packard</td>
<td>-</td>
</tr>
<tr>
<td>Pulse generator</td>
<td>PBG2</td>
<td>Kentech</td>
<td>J9606181</td>
</tr>
<tr>
<td>IRA antenna (transmit)</td>
<td>200 MHz - 4 GHz</td>
<td>TNO-FEL</td>
<td>-</td>
</tr>
<tr>
<td>IRA antenna (transmit)</td>
<td>200 MHz - 4 GHz</td>
<td>TNO-FEL</td>
<td>-</td>
</tr>
<tr>
<td>Trigger delay generator</td>
<td>DG535</td>
<td>SRS</td>
<td>-</td>
</tr>
<tr>
<td>fast transition time unit</td>
<td>DG535 OPT-04A</td>
<td>SRS</td>
<td>-</td>
</tr>
<tr>
<td>fast transition time unit</td>
<td>DG535 OPT-04A</td>
<td>SRS</td>
<td>-</td>
</tr>
<tr>
<td>Bias tee</td>
<td>DG535 OPT-04C</td>
<td>SRS</td>
<td>0639</td>
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<td>Time gate switch</td>
<td>ZYSW-2-50DR</td>
<td>Micro Circuits</td>
<td>-</td>
</tr>
<tr>
<td>DC Power supply</td>
<td>-</td>
<td>-</td>
<td>-</td>
</tr>
<tr>
<td>10dB attenuator</td>
<td>23-10-34 DC-18 GHz</td>
<td>Weinschel corp.</td>
<td>BD0765</td>
</tr>
<tr>
<td>20dB attenuator</td>
<td>-</td>
<td>-</td>
<td>-</td>
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<tr>
<td>20dB High voltage attenuator</td>
<td>-</td>
<td>-</td>
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