Frank Sabath · Eric L. Mokole *Editors*

Ultra-Wideband, Short-Pulse Electromagnetics 10



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Carl Edward Baum

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In Memoriam of Carl Edward Baum (February 6, 1940–December 2, 2010)

We will bear in remembrance Carl's personality A huge colorful bunch of stories Carl's fair-minded remarks and answers His knowledge in music and history

A friend

Preface

When the first volume of the *Ultra-Wideband*, *Short-Pulse Electromagnetics* book series was published in 1993, the terms ultra-wideband (UWB) and short pulse (SP) were acronyms for challenging technologies. In 1992, the DARPA Ultra-Wideband Radar Review Panel defined UWB by the need for special techniques to overcome challenging problems facing conventional systems and technologies when attempting to operate over a broad range of frequencies.

Since then notable progress in UWB and SP technologies has been achieved. As a result, wideband systems are now being used for an increasingly wide variety of applications. UWB radar systems are used for collision avoidance, concealed object detection, mine detection, and oil pipeline inspections. In the communication area, the need for increasing bandwidth boosted the development of UWB communication systems including the impulse radio. Many high-power electromagnetic (HPEM) environments are generated, employing short-pulse technology. With the advent of HPEM sources capable of interrupting and/or damaging sensitive electronics, there has been an increasing interest in protecting critical infrastructure and systems. Recently, the literature has reported the usage of SP techniques in microwave tomography systems for biomedical applications.

Through the whole development of UWB and SP technologies, the *Ultra-Wideband*, *Short-Pulse Electromagnetics* series of books provided new and stateof-the-art information on the tendencies and current achievements in UWB- and SP-related technologies, analyzing methodologies, theoretical models, and time domain data processing. The objectives of the *Ultra-Wideband*, *Short-Pulse Electromagnetics* book series are:

- To focus on advanced technologies for the generation, radiation, and detection of UWB and SP signals
- To report on developments in supporting mathematical and numerical methods, which are capable of analyzing the propagation of UWB and SP signals as well as their scattering from and coupling to targets and media of interest
- To describe current and potential future applications of the UWB and SP technology

"Ultra-Wideband, Short-Pulse Electromagnetics 10" (UWB SP 10) contains articles which present recent developments in the areas UWB and SP technology, components, application, numerical analysis, modeling, and electromagnetic theory. The initial set of contributions was selected from presentations at the UWB SP 10 and UWB SP 11 conferences that were held in conjunction with AMEREM 2010 in Ottawa (ON), Canada, and EUROEM 2012 in Toulouse, France. The goal of the volume editors was to cover the complete range of aforementioned topics with articles of deep technical content and high scientific quality. Where we felt that there were gaps in coverage, selected authors were invited to contribute additional articles to complete the overall picture. Therefore, we hope that this book contains something of interest for every scientist and engineer working in the area of ultra-wideband and short-pulse electromagnetics.

With a slight variation on a tradition in the *Ultra-Wideband*, *Short-Pulse Electromagnetics* (UWB SP) series, particularly those which are related to EUROEM conferences, a frontispiece displays the picture of a renowned scientist. In the opinion of the editorial board, the title of the UWB SP 9 book completed the series consisting of Maxwell, Hertz, and Einstein.

Therefore, we decided to use the frontispiece of UWB SP 10 to start a new series. In important respects, this volume is special. It establishes ties between North America and Europe as it contains contributions from both AMEREM 2010 and EUROEM 2012. A human tie between both the AMEREM and EUROEM series was Carl E. Baum, who passed away in December 2010. AMEREM 2010 was the last conference of this series that he attended, and EUROEM 2012 took place without Carl in the audience or delivering numerous presentations.

The frontispiece of this UWB SP 10 book displays a picture of Carl Baum during his acceptance speech of the honorary doctoral degree from the Otto-von-Guericke University in Magdeburg. This 10th volume in our series honors Carl, a remarkably creative engineer who introduced innumerable new concepts in mathematics, electromagnetic theory, and system design, many of which remain the standards of excellence today. From his earliest designs in electromagnetic pulse sensors and simulators to the latest developments in high-power microwave and ultra-wideband antenna and system design, his research has remained at the forefront of technology.

Finally, I would like to express my gratitude to all persons who contributed to this book. In particular, I thank the authors for writing articles of deep technical content and high scientific quality and the members of the review board who helped to improve the quality of this book.

Munster, Germany

Frank Sabath Editor-in-Chief

Carl Baum, a Great Man and Eminent Scientist

My First Encounters with Carl

In the spring of 1986, I began my work as a theoretical physicist with the German Federal Armed Forces in Munster. It was my job to analyze and interpret the NEMP data that had been measured. Since I was coming from a very different area of specialization, I first needed to intensively learn about my new area of research (EMP-analysis and EMP-lab measurements).

At the end of the 1970s, the German Research Institute for NBC Protection in Munster built a large EMP-wave guide simulator with large support from American firms and under the guidance of Carl Baum. Thereafter, Carl—who was living in the USA—became a regular guest in Munster and, during his visits, gave many lectures about his EMP research results, all of which he also published in his Notes. Of particular interest to us at that time were his Interaction Notes and the Sensor and Simulation Notes. During his lectures, two things in particular struck me about Carl. First, Carl used handwritten, colorful overhead transparencies on which each of the colored letters represented a specific physical quantity. And, secondly, he expected a certain level of knowledge from his audience, which most of us did not have. During the question and answer period, he often referred to his notes with the demand to read them again, or he gave answers which were too complicated to understand. This, of course, was not very good teaching practice and did not please many people in his audience. In short, his lectures were appreciated by only a small number of people. This was a particular challenge for me, and I began to study Carl's Notes and decided to stay in contact with him. The next opportunity to do just that arrived with the EMP-measurement campaign for the jet fighter "Tornado" at the Kirtland Air Force Base in Albuquerque, New Mexico, USA, in 1987. I participated in these measurements for 3 weeks and met with Carl many times during this period either in his office or for dinner. Because a large part of our measurements on the "Tornado" were current measurements on cable bundles, I was particularly interested in talking to Carl about EMP-coupling problems and EMP interactions with cables and cable bundles. The focus was laid on the

topological coupling models for large systems and subsystems as well as on the BLT-equations and possible solutions of them. During these talks I found a willing listener in Carl when I suggested generalizing the classical transmission-line equations to nonhomogeneous cable layouts. In addition to our science talks, I realized that Carl loved to eat and drink well. His keen interest in history, especially the history of famous battles, was well known. Sometimes he would give his fantasy free reign and imagine how famous battles could have turned out if other tactical and strategic maneuvers had been used. Furthermore, he was a science fiction film fan. He would often have fun imagining that the final battle scenes in such films took place in his prestige object the TRESTLE—a huge EMP simulator made of wood—which he planned and designed.

After these initial meetings with Carl, I decided that I would like to spend a research year at the Air Force Base in Albuquerque with Carl as my advisor. Upon an agreement between the Federal Republic of Germany and the US Ministries of Defense, my petition for a year-long research opportunity was granted.

A Research Year Under the Guidance of Carl

In April 1989 I was finally ready to begin. After a 3-month intensive English class, taught by native speakers, I was able to take a sabbatical year in the USA, together with my family, in the scope of the German–American Exchange Project for Military Science. With Carl's help, I was able to secure a job in the Air Force Lab at Kirtland Air Force Base in Albuquerque, NM, in the same building as Carl's office. This would be a very impressive and successful year for me.

I was able to freely choose my research topic, and since my institute in Germany was especially interested in the coupling of electromagnetic fields to cable bundles, I chose a research topic in this area, namely, "Electromagnetic Coupling Processes to Nonuniform Transmission Lines." This topic was quite topical in the 1930s but more in the framework of matching impedances between generators and antennas. We wanted to now study the generalization of classical transmission theory to nonhomogeneous transmission-line systems. It was in our interest to represent all new results analytically, at least as far as possible. We built upon the BLT-equation, which Carl helped develop. We wanted to extend it to quasi-classical transmissionline systems (weak disturbances of the TEM modes). To do this we used mode decompositions methods, symmetry characteristics of line arrangements (e.g., cyclical symmetry), and quantum methodological solution methods. At the end of the year, we had published many Notes (Interaction Notes and the first Physical Note), which later were published in the IEEE Transactions on EMC. Every day I found in Carl a good discussion partner and advisor. He regularly walked up and down the lab hallways with a mug of coffee in his hand, looked into the various rooms to see what was going on, and, of course, was always ready to tell a story or anecdote.

Carl also made sure to introduce me to all the lab heads and important people in his department and brought me along to see all the EMP simulators on the Air Force Base. During the lunch breaks in the cafeteria, I also regularly ran into scientists from the Sandia lab.

In addition to our professional meetings, my family and I also met with him privately. Carl introduced us to Bill Prather, who helped us find a good preschool for our daughter and a high school for our son to attend and who was also a big help in our finding a small house to rent. We liked to host people in our new home, and Carl was almost always among our guests. He was a lover of good German beer and red wine. We also enjoyed dining at restaurants in Albuquerque with him. He loved to be the center of discussions, and due to his loud laugh and voice, he was easy to find in a crowd. Once we were invited to Carl's "bachelor pad," Carl had decided not to marry and have a family, but instead to dedicate his life to science and to his work with the Catholic Church. When we were in his house, it was clear that the house could have benefited from a woman's touch. In the middle of the living room were a large TV and a round wooden table, on which lay a large pile of handwritten notes, unfinished musical compositions, and red leather-bound books with musical compositions by Carl, which had already been published and performed in churches. On the kitchen table were large bottles of whiskey and gin, the contents of which surely lasted for a very long time. And although there were a few overhead transparencies written on in colored markers, Carl did most of this research in his office with paper and pencil.

At the end of my research year in 1990, our Interaction Note 477 entitled "Splitting of Degenerate Natural Frequencies in Coupled Two-Conductor Lines by Distance Variation" was awarded Best Basic Paper at the AMEREM 1990 in Albuquerque. This was a wonderful way to finish my collaborative year with Carl.

Carl and Europe

It can be said that Carl was an EM prophet. He traveled around the world in order to spread his message (his scientific knowledge) with all who were interested. At first he limited his travels to West-European countries but later traveled to Russia and Asia as well. His trips through Europe were meticulously planned. In most of the countries he traveled to he had a contact person (in Germany this was me), who recommended the various places he should visit on his trip and, upon his agreement, found a contact person in each town. The contact person was also responsible for organizing his lodging. Carl would then finalize his itinerary. His luggage always consisted of a large suitcase that had been repaired many times and refurbished with brown fabric as well as a large square briefcase filled to overflowing with his colorful lecture transparencies, which were carefully sorted into files according to the various lectures he would be giving. His visitors had the chance to request a specific lecture topic. The institutes and companies that Carl regularly visited in Germany included the Technical University of Hamburg, the *Bundeswehr* Research

Institute for Protective Technologies and NBC Protection in Munster, the Leibniz University of Hannover, the Otto-von-Guericke University in Magdeburg, the University of Kassel, and the company Diehl in Röthenbach/Pegnitz. As time went on, other universities and cities would join the list. Carl was very interested in keeping up with the newest scientific discoveries and added his thoughts and advice to the conversation. Often his ideas would lead to new discoveries or to new questions. His discussion partners could always find in Carl a reliable source for good explanations, constructive criticism, and new ideas. Thus, for most of the people he came into contact with, it was not so important whether they could understand his hard-to-decipher, colorful lecture materials. More important, especially for young doctoral students, were his helpful comments, explanations, and ideas. In this way, Carl quickly become known within the German electrical engineering academic world as an international expert in this area, and his publications, including his Notes, became increasingly popular. Carl was also warmly welcomed in my department at the Otto-von-Guericke University in Magdeburg. My colleagues took great pleasure in their discussions with him, and in sharing his company in the evening, sitting and talking over a glass of wine and a good meal. These were always very entertaining evenings.

Bestowal of the Honorary Doctoral Degree to Carl

In 2004, the 4th EUROEM conference was held in Magdeburg and was organized and sponsored by my department at the university there. The organizers and myself wanted to present Carl with a special honor and knew that the Honorary Doctoral Degree from the Department of Electrical Engineering and Information Technology at the Otto-von-Guericke University was a title that Carl did not yet have among his long list of academic awards and titles. I personally made the petition to start the review process for this award, basing my arguments on the extraordinary service that Carl had provided to our department over the years. The petition was accepted and a faculty awarding committee formed. After receiving and reviewing recommendations from, among others, Professors Tesche and Felsen, the faculty committee decided in favor of the award bestowal. The EUROEM conference presented itself as a good opportunity to award Carl with the honorary degree.

We began to plan the celebrations and award ceremony, keeping in mind that Carl had a special relationship to the Church and to history and that he had a penchant for old academic traditions. Therefore, we decided to award the Honorary Doctoral Degree in two steps. First, there would be an official ceremony in which the official degree would be bestowed to Carl by the Dean of the Faculty and in the presence of the university Rector. This would take place in the *Johanniskirche*, an old church in Magdeburg in which Martin Luther preached more than 450 years ago.



Carl with his certificate for honorary doctorate. From left to right: myself, Rector, Carl, and Dean

The second, traditional academic part of the ceremony would take place in the Historical Museum of Magdeburg: wearing historical academic gowns, the degree written in Latin would be awarded in this separate ceremony.



Carl's entry into the Historical Museum, escorted by a medieval guard

The opening ceremony of the EUROEM was held at the end of the first day of the conference in the *Johanniskirche*. The conference participants were seated as the organizers, sponsors, and Carl entered the church to the sounds of a trumpet fanfare played from the upmost gallery. The trumpet music was, in fact, composed by Carl. A second special feature, which was not lost on the audience, is that Carl was, as far as I know, wearing a suit for the very first time. Naturally, he did not bring the suit along with him in his luggage; we bought it for him the previous weekend. He wore the suit two more times during his visit and then proceeded to leave it behind in Germany when he returned to the USA.



Carl's short lecture (with his famous multicolored slides) after the appointment of an honorary doctorate

The mayor of the city of Magdeburg hosted the second ceremony in the Historical Museum. During this ceremony the academic officials, dressed in their academic robes, entered the hall together with Carl. They were led by a procession dressed as an historical, medieval guard. During this ceremony the participants learned a bit about the centuries-old history of the city. The honorary degree, this time written in Latin, was once again handed over by the Rector of the university.



Carl (in black and white cap and gown) shows his certificate for honorary doctorate, which is written in Latin and refers to his special academic achievements. To his *left* myself, to his *right* the Rector and the Dean, and behind him the guard

We learned later that of all the awards and honors bestowed upon Carl in his lifetime, the Honorary Doctoral Degree from the Otto-von-Guericke University in Magdeburg meant the most to him. Carl, we are very pleased to know this, we thank you with utmost respect, and we will never forget you.

Magdeburg, Germany

J. Nitsch

Several Years of Interaction with Carl E. Baum

Being Introduced to Carl

My first encounter with Carl was in 1991 at the Prague URSI General Assembly. I was preparing my Ph.D. on EM Topology (EMT), and, of course, he was the person to whom I had to present my results. I remember very well to have introduced the model of two wires coupled by an aperture and the conditions for applying "Good Shielding Approximation." What a surprise for Carl, a young French scientist working for 3 years on EM Topology, a subject that was mostly considered as a theory at this time! I remember very well Carl putting me, the young impressed student, around his large arms and telling me that I had to visit him quickly. Since this encounter, Carl has always followed and supported the evolution of my work. Especially, we visited each other many times, and Carl has always been curious of the progress of the applications I would make of his theoretical work on EMT and cables. We were both aware that I would never be a kind of theoretical guy as he was but, as an engineer, I was a fully convinced applicator of his scientific ideas.

Carl and His "Protégés"

From 1993 to 1996, my research group was quickly given the opportunity to apply for the first time the CRIPTE code and its EMT formalism, right at Carl's working place in Albuquerque, on a complex system, the EMPTAC, this test-bed aircraft on which several EMP tests had been made in the past. The objective was challenging: obtaining computed results of cable responses in a real aircraft and comparing them to measurements at the same time. It is true that the first campaign was a success because of the impact our results had on people. During our 3 weeks of experiments, Carl has always been very present. When he came visiting the test site, he was like a commander discussing measurement techniques with his troops and checking that the sensors were used properly. Hopefully, our guys were well aware of usual bad practices with field and current sensors.

The most impressive impact was when, 1 month later, we were urgently requested by Carl to present our fresh new results in a restricted Air Force meeting. Carl presented our work as what we had been waiting for 20 years! The situation was embarrassing.

The same situation happened in 1996 when Carl spoke after my plenary session lecture on EMT at the URSI General Assembly in Lille, promoting the "outstanding results" of this French Group." Carl had "protégés" all over the world, as he used to say. For sure Carl knew how to rely on the work of his friends and colleagues in order to promote and justify his own messages on EM. As far as EMT is concerned, he wrote the theory, and we put it in application, or at least modestly, partly!

Carl and Administration Work

After 1996, I had taken the decision to apply for a Sabbatical position sponsored by my French MoD. For this, I had to find a laboratory to welcome me during 1 year. I had the opportunity to visit two laboratories. My first laboratory visit was full of hope, but my contact told me I had to wait for some time before having the final decision. My second visit was at Carl's office. Carl was so delighted of my request that in a 1-h interview in which I explained briefly my project, my recommendation letter was written with Carl's pencil on a sheet of his yellow notebook, sent to the secretary for being typed, duly signed by Carl's hand, and finally back in my hands. Who said that Carl was not interested in administrative work? Actually, he could be very efficient with administration when he needed it to reach his objectives!

Carl's Working Environment

So I had the chance to spend 1 year at the Phillips Lab (now "Air Force Research Lab") in Albuquerque working in a small building close to Carl's building and leaving in a small house close to Carl's house. I could thereby share with him some more intimacy in work and private life. This period definitely sealed our friendship.

Carl's office is something that was worth visiting: a large room topologically divided in two elementary volumes separated by a wall made of cabinets in a row. In the first room, accessible by the entrance door, many chests of drawers for storing the paper versions of the Notes sorted out in perfect order in each drawer. The second topological volume, in the back of the room, was Carl's office place, with a comfortable armchair, a large desk, and a high-fidelity radio set always turned on MPR classical music channel. On his office table lay Carl's working tools—a pencil, a rubber, a pencil sharpener, a yellow-page notebook, and, very important, on his left-hand side, another table with several small boxes containing all his contact addresses and telephone numbers. Indeed, Carl's address book was huge, built in from the many trips he was making around the globe. Finally, to conclude this EM Topological decomposition, along the main wall, common to the Notes volume and Carl's Office volume, was Carl's library. At the end of my 1-year assignment in Albuquerque, I did not even think of going to the AFRL library anymore; Carl had everything in his library, even the English editions of our French EMC reference books! Besides, Carl did not like changes. When he moved to the University of New Mexico (UNM) as associate professor, he installed his new office identically to his office at AFRL. The visitors did not see any changes. The UNM associate professor still looked exactly as the Air Force well-known scientist.

Carl's house was also worth the visit, a very welcoming house, where you could always be offered one of the largest choices of whiskeys in Albuquerque. But, most of all, it was another working place as well for him, with a large and thick circular wood table full of musical composition sheets, presentation slides, and colored felttip pens spread all over the surface. If the pencil was his main instrument for formalizing his thinking, the overhead projector has been for a long time an important communication tool to give colors to his thinking!

Working with Carl

Working close to Carl was also an experience. Actually, you did not work with Carl, but you worked close to him. Indeed, the best way to interact with Carl was to submit him your work for review; never think of asking him to explain something you did not understand, especially if the question was about one of his papers! His first reaction would have been the surprise and disappointment that you could not understand it!

Once this lesson learned for good, my 1-year assignment became a very intense scientific period in which Carl reviewed my whole work. I had the chance to investigate and implement several topics that are still nowadays key features of the CRIPTE code. Ideas went fast since Carl picked up things very quickly. One example is as follows: I investigated several ways to optimize the labeling of waves on a network for the BLT sparse resolution, and, after some research that logically took me a significant time, I reached a solution that I realized to be Carl's first idea when we started discussing of this issue.

Carl the Leader and Organizer

Another aspect of Carl's impact on our scientific community was the fact that he did not have the soul of an organizer but he always had clever ideas of organizations. One of the most durable ones is of course the cycle of AMEREM and EUROEM conferences which keeps having a specificity that cannot be found in any other conference. Carl wanted me for EUROEM 2012, and finally, after calling me at home every day in May 2008, with the stubbornness of a stripped LP, he convinced me to propose ONERA's candidature, which of course was accepted in a row at the EUROEM conference in Lausanne.

Another aspect of Carl's durable memory is surprisingly a teaching activity. Carl invented the HPE course concept, a week in an isolated place with no other temptation than working on EM coupling and high-power electromagnetics, the whole thing in a good friendship atmosphere. In the HPE course, the word "seminar" takes a real sense. Students are like monks: even the waking up with a bell is part of the ritual. However, the originality of this course is not there; it is in the active role that Carl wanted students to play. Indeed, they were submitted problems by the faculty, had to solve them in groups, and present them to the class. This contributes to the cohesion of the group and builds real lasting friendship among students. The good thing is that this course still continues nowadays, thanks to Dave Giri who perpetuates the tradition.

Carl the Globe-Trotter

Carl was also a globe-trotter. Of course, he was not the kind of person carrying a bag pack, but everyone will remember him pulling his large wheeled suitcase in one hand and his smaller one, full of slides cautiously selected for the travel stops, well stored in dedicated brown folders, in his other hand. How many trips did he make? How many countries did not welcome him?

Actually, we say "travels," but we should speak of "tours." After my first encounter with Carl in Prague, ONERA had become one of the labs to stop by in the long trips he made all over the world. There was good French food and valuable work to accommodate his standards. Back from Albuquerque in 1999, I became Carl's French contact for the organization of his "Tours de France." Everything had to be prepared well in advance with several stops in several French cities. Those stops had to be followed by other stops in other European cities organized by some of my European colleagues who took over from me. Nowadays, I realize how much those tours were pilgrimages to bring his word of EM to the newcomers. Can you imagine being a student and having the chance to present your subject to Doctor Baum? How to forget such a moment? This brings me back to the beginning of our discussion, some 20 years ago, when I presented my aperture model to my friend, Carl.

Toulouse, France

J.-P. Parmantier

Contents

Part I Theory and Modeling

Quantifying Uncertainties Strength from Electromagnetic Stochastic Simulations B. Jannet, P. Bonnet, S. Lalléchère, and B. Pecqueux	3
Transmission-Line Super Theory as Antenna Theory for Linear Structures R. Rambousky, J. Nitsch, and S. Tkachenko	13
Electromagnetic Coupling to Transmission Lines with Symmetric Geometry Inside Rectangular Resonators	31
Analysis of Open TEM-Waveguide Structures	49
Antenna Radiation in the Presence of an Infinite Interface Y. Béniguel	59
A Standard for Characterizing Antenna Performance in the Time Domain Everett G. Farr	71
Time-Domain Distortion Characterization of ElectromagneticField Sensors Using Hermite–Gauss SubspacesShekoofeh Saboktakin and Behzad Kordi	79
Amplification of Signal Intensity in Imaging Through DiscreteRandom Media Through Signal Interference Gatingand the Use of Mutual Coherence FunctionElizabeth H. Bleszynski, Marek K. Bleszynski, and Thomas Jaroszewicz	89

Propagation of Impulse-Like Waveforms Through the Ionosphere Modeled by Cold Plasma	101
Electromagnetic Environment of Grounding Systems M. Lefouili, I. Hafsaoui, K. Kerroum, and K. El Khamlichi Drissi	113
A Global Circuit Tool for Modeling Lightning Indirect Effects on Aircraft	127
Fully Absorbing Conditions in the Study of AxiallySymmetrical UWB RadiatorsOlena Shafalyuk and Paul Smith	137
Part II Time Domain Computational Techniques	
The New Vector Fitting Approach to Multiple ConvexObstacles Modeling for UWB Propagation ChannelsP. Górniak and W. Bandurski	147
FDTD Analysis of Shielded High-Tc Microstrip Resonatorson Anisotropic SubstratesM.L. Tounsi, O. Madani, and M.C.E. Yagoub	159
Parametric Evaluation of Absorption Losses and Comparison of Numerical Results to Boeing 707 Aircraft Experimental HIRF Results J. Kitaygorsky, C. Amburgey, J.R. Elliott, R. Fisher, and R.A. Perala	169
Part III Antennas	
Directional Dependence of the Minimum-Phase Property of the TEM Horn Transfer Function	179
UWB Dual-Polarized Antenna for HPEM Sources J. Schmitz, M. Camp, M. Jung, G. Adamiuk, S. Scherr, and T. Zwick	189
Modal Analysis of Reflector Backed Hybrid Printed Bow Tie Antenna Dhiraj K. Singh, D.C. Pande, and A. Bhattacharya	199
A Novel Class of Reconfigurable Spherical Fermat Spiral Multi-port Antennas D. Caratelli, A. Yarovoy, and N. Haider	209

Highly Directive Epsilon Negative Metamaterial-Loaded		
Circular Patch Antenna for Triple-Band Performance	219	
M.H. Ababil, M.F. Saimoom, S.M.H. Chowdhury,		
M.R.C. Mahdy, and M.A. Matin		

Part IV Pulsed Power

A New Set of Electrodes for Coaxial, Quarter Wave, and Switched Oscillators	231
F. Vega, F. Rachidi, D.V. Giri, B. Daout, F. Roman, and N. Peña	
Performances of a Compact, High-Power WB Source with Circular Polarization P. Delmote, S. Pinguet, and F. Bieth	239
Design Considerations for a Switch and Lens System for Launching 100 ps, 100 kV Pulses	251
Part V UWB Interaction	
EMI Risk Management with the Threat Scenario, Effect, and Criticality Analysis	265
On the Use of Probabilistic Risk Analysis for Intentional Electromagnetic Interference E. Genender, S. Fisahn, H. Garbe, and S. Potthast	279
Susceptibility of Electrical Systems to UWB Disturbances Due to the Layout of Exit Cables	289
Breakdown Behavior of a Wireless Communication Network Under UWB Impact	299
The Technique for Evaluating the Immunity of Digital Devices to the Influence of Ultrawideband Electromagnetic Pulses	309
Reciprocity Theorem: Practical Application in EMC Measurements	319
Coupling of Hyperband Signals with an Underground Cable K. Sunitha, M. Joy Thomas, and D.V. Giri	331

Part VI SP Measurement

HPM Detector with Extended Detection Features Chr. Adami, Chr. Braun, P. Clemens, M. Jöster, HU. Schmidt, M. Suhrke, and HJ. Taenzer	345
High Dynamic Range, Wide Bandwidth ElectromagneticField Threat DetectorD.B. Jackson, T.R. Noe, and G.H. Baker III	355
Resistive Sensor for High-Power Microwave Pulse Measurement in Double-Ridged Waveguide	369
Characteristic HPEM Signals for the Detection of IEMI Threats A. Kreth, T. Peikert, B. Menssen, and H. Garbe	379
High-Frequency Impedance Measurement of Electronic Devices Using a De-embedding Technique	393
Automated and Adaptive RF Effects Testing	403
Pockels' Effect-Based Probe for UWB and HPEM Measurements G. Gaborit, P. Jarrige, Y. Gaeremynck, A. Warzecha, M. Bernier, JL. Lasserre, and L. Duvillaret	411
Part VII UWB Sensing	
Estimating Magnetic Polarizability Tensor of Buried Metallic Targets for Land Mine Clearance	425
UWB Short-Pulse Radar: Combining Trilateration and Back Projection for Through-the-Wall Radar Imaging	433
Toward Integrated μNetwork Analyzer M. Kmec, M. Helbig, R. Herrmann, P. Rauschenbach, J. Sachs, and K. Schilling	443
M-Sequence-Based Single-Chip UWB-Radar Sensor	453
UWB Antennas for CW Terahertz Imaging: Geometry Choice Criteria I. Türer, A.F. Dégardin, and A.J. Kreisler	463

UWB Antennas for CW Terahertz Imaging: Cross Talk Issues	473
A.J. Kreisler, I. Türer, X. Gaztelu, and A.F. Dégardin	
Evaluation of Imaging Algorithms for Prototype Microwave Tomography Systems	483
Index	493

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xxviii

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Part I Theory and Modeling

Quantifying Uncertainties Strength from Electromagnetic Stochastic Simulations

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Abstract In this chapter, we associate sensitivity analysis (SA) with the stochastic collocation (SC) method in the context of electromagnetic problems. The SC has shown its efficiency and precision on problems with a little number of random variables [Chauvière et al. SIAM J. Sci. Comput. 28: 751–775, 2006]. The SA, as a first step, compensates the major drawback of the SC in reducing the number of entries to only the predominant ones. The combined method appears to be precise and effective to deal with problems with numerous uncertainties.

Keywords Time reversal • Stochastic collocation • Sensitivity analysis • Morris method

1 Context

This work is achieved in the context of the time reversal (TR) process [1]. The TR was developed by Fink's team in acoustics in the early 1990s. Then the method has been adapted to electromagnetism. It is based on the reversibility of the wave equation $[\varphi(t) \text{ and } \varphi(-t) \text{ are solutions of the same Eq. (1)}]$:

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Fig. 1 Scheme of the two steps of the TR process

$$\frac{1}{c^2}\frac{\partial^2 \varphi}{\partial t^2} = \Delta \varphi, \tag{1}$$

where φ may represent the electric E-field or magnetic H-field and c is the propagation speed of the electromagnetic waves in the medium.

The process is traditionally split into two stages. In the first step, the source produces a signal that propagates in the medium and is recorded by receptors. In the second step, the recorded signals are temporally reversed and sent back in the medium from the places they were recorded. The waves will refocus on the initial source point (Fig. 1).

One of the main properties of TR is the spatial and temporal focus on the source point. This method is used in many applications, like tumor destruction, focused communication, or localization behind walls [2].

However, this method supposes that the medium does not change between the steps. In practice it is seldom to have these conditions.

There can be many variations that can affect the medium between the steps of the TR (modifications of positions or dimensions, variations of electric properties of equipments or medium, some changes due to the repeatability and precision of measurement devices, positions of source or receptor may have changed, etc.). These variations are considered as uncertainties to be dealt with. To see how they affect the TR focalization, there are many quality criteria (Fig. 2). We choose in this chapter to study the amplitude of the focused electric field.

The classical (and reference) approach for uncertainty determination is the Monte Carlo (MC) method [3]. It is simple and robust, but its convergence rate is slow (it requires a lot of realizations to access a trustable result).

We use here another stochastic method which is very cost-effective (from computational requirements) and yet precise: the stochastic collocation (SC) method.



Fig. 2 Different quality criteria to qualify the TR

2 The Stochastic Collocation Method

The SC is a non-intrusive technique which can give the statistical moments of the field with only a small number of realizations (details in [4]). The SC is based on several steps: the field E is projected on a Lagrangian polynomial basis:

$$E(a) = \sum_{i=0}^{n} E_i L_i(a).$$
 (2)

The Lagrangian polynomials have a very interesting property:

$$L_i(a) = \prod_{j=0}^n \frac{a-a_j}{a_i - a_j},\tag{3}$$

$$L_i(a_j) = \delta_{ij} \to E_i = E(a_i). \tag{4}$$

The collocation points p(x) and weights ω_i are found with a Gaussian quadrature:

$$\int_{D} p(x)E(x)dx \approx \sum_{i=0}^{n} \omega_{i}E(x_{i}),$$
(5)

$$\int_{D} p(x)L_{i}(x)\mathrm{d}x = \omega_{i}.$$
(6)

The statistical moments are calculated; let us see the formulas for mean and variance:

Fig. 3 The 2D configuration with a moving object, reflectors, and absorbers



$$\langle E(x)\rangle = \sum_{i=0}^{n} \omega_i E_i,\tag{7}$$

$$\operatorname{var}(E(x)) = \sum_{i=0}^{n} \omega_i E_i^2 - \langle E(x) \rangle^2.$$
(8)

The SC is applied on a TR example; a metallic object (50 cm \times 50 cm) is placed in a bidimensional (2D) domain. The position of this object changes along a linear one-dimension path between the two steps of the TR (Fig. 3).

The object position follows the formalism: $Y = Y_0 + u$, where Y_0 is the mean position of the object and u is a random variable (RV) following a uniform law on the range [-5 cm, 5 cm]. The electromagnetic problem is simulated from FDTD formalism [5]. The computational domain is 1 cm sampled. The maximum frequency of the Gaussian source is 3 GHz. Absorbing boundary conditions are chosen. At the edges, some PEC (perfectly electric conductor) reflectors are placed. We use one source and one receptor.

The first step of the method is to ensure that the SC converges. As said earlier, for each collocation point, the amplitude of focalization is studied. The means of SC for five, seven, and nine test points in the test object are represented in Fig. 4; the SC quickly converges (seven collocation points). A comparison is performed with MC, and it shows a good agreement between the converged values: the efficiency of SC (19 executions for SC, 3,000 realizations for MC).

Then we need to evaluate the impact of uncertainties on the TR. For this, some quantitative criteria are necessary.

First, we define the amplitude ratio (AR, in percent): $AR = 100 \times \mu/E_{max}^0$, where μ is the converged mean from the SC and E_{max}^0 is the focalization amplitude without change between the steps. The AR outlines the impact of the variations on the focalization amplitude and tells if the TR is preserved or not.



Fig. 4 Means of SC and MC for Fig. 3 2D case, SC convergence with seven points

Table 1 Number of required realizations for the SC with nine collocation points

RV number	1	2	3	4	5	6	7	8
Cost	21	161	1,221	9,665	79,221	665,441	5,686,821	49,208,705

Second, we use the coefficient of variation (CV, in percent), which indicates the variability of the case: $CV = 100 \times \sigma/\mu$, where σ is the converged standard deviation from geometrical uncertainties (Y movement) from the SC.

Here, obtained criteria are AR = 60 % and CV = 33.94 %. The CV is relatively high, and potentially a huge variability (influence) may be expected due to stochastic variations. The AR is quite low; the movement of the object has a strong effect during TR process. We have here an efficient method to evaluate the impact of uncertainties on the TR.

The SC is effective, but this efficiency falls with the increase of the number of RVs. Shown in Table 1 are the costs (number of realizations for nine collocation points) in function of the numbers of RV. This number is equal to $1 + (3^d - 1) + (5^d - 1) + (7^d - 1) + (9^d - 1)$, where *d* is the number of RV.

A solution could be to use another stochastic method (Unscented Transform [6], Stroud [7], etc.). A second approach consists in the use of a sensitivity analysis (SA) method at a first stage of the analysis, to reduce the number of variables and then use the SC on the remaining entries.

3 Sensitivity Analysis Methods

The aim of SA [8] in which we are interested is to reduce the model in setting the non-influential entries to their mean value. There are numerous methods with different advantages and drawbacks (screening techniques, experimental design,



Fig. 5 Morris's method results: three RVs appear influential

graphical methods, etc.). We use the Morris [9] method, because it is simple and gives the global influence of each RV for an acceptable cost. It is non-intrusive, so its association with the SC keeps this property. It belongs to the screening technique family. This process sorts the entries in three groups:

- Group 1: Negligible effects
- Group 2: Linear effects and no interaction with other variables
- Group 3: Nonlinear effects and/or interactions with the other variables

In the updated model of the problem, the entries of the two first groups will be set to their mean values; the SC will be computed on group 3.

The details of the Morris technique may be found in [10]. Nevertheless the principle can be summarized as explained in the following.

Let "*N*" be the number of RV. The domain is discretized, and we draw a set of entries (a starting point) and then we achieve *N* "one-at-a-time" (OAT) experiments (1 for each RV). For each OAT (each RV), we calculate an elementary effect. A number of sketches are drawn. This process allows calculation of the mean (μ^*) and standard deviation (σ) of effects from each RV.

The results (μ^* and σ) are plotted on a graph. They show the interactions (μ^*) as a function of the influences (σ). A certain number of repetitions of the previous process give a cloud of points for each RV. This method has been applied on an analytical formula of a transmission line [9] with 11 RVs (on geometry and electrical properties). Figure 5 reveals that only three RVs will remain influential after the SA step, considering a classical EMC issue.

4 Combining the Methods (SA + SC) on a Numerical TR Problem

In this section, we associate the Morris method with the SC considering a TR simulation (Fig. 6) with the Gorf3D [11] FDTD software of CEA-Gramat (France). We will show that as a preliminary step, the Morris method sorts the variables and puts out two or three entries. After, the SC on the reduced problem quantifies the impact of the variations on the TR.

The setup is an extension in 3D of the previous 2D case with two wire dipoles inside; absorbing conditions are used, and at the edges of the domain, several PEC reflectors are placed. The TR is applied between the wires (current is generated and received in the middle of the dipoles). The source signal is a Gaussian-modulated sinusoidal pulse, $f(t) = A \times \cos(2\pi F_0 t) \times \exp(-0.5((t - t_0)/\alpha)^2)$, with A = 1,000, $F_0 = 2$ GHz, $t_0 = 10^{-8}$ s, and $\alpha = 1.61 \times 10^{-10}$ s.

Eight RVs are selected to be studied: two resistances and the radius for each wire (ResS1, ResS2, RadS and ResR1, ResR2, RadR), a shift of the position of the receptor wire in two directions (Sx and Sy). The RVs follow uniform laws, and the variation ranges are presented in Table 2.



Fig. 6 The 3D case with two wires dipoles

Table 2 Variation ranges of the eight RVs of the 3D case with wires

	Sx	Sy	RadS	RadR	ResS1	ResS2	ResR1	ResR2
Average	0	0	2 mm	2 mm	50 Ω	50 Ω	50 Ω	50 Ω
Range (+/-)	0.5 cm	0.5 cm	0.9 mm	0.1 mm	10Ω	10 Ω	5Ω	5Ω
Percent	$6.25 \ \%^{a}$	$6.25 \ \%^{a}$	45 % ^b	5 % ^b	20 % ^b	20 % ^b	10 % ^b	10 % ^b

^aPercent of the size of the wire

^bRelative variation from mean value



Fig. 7 Morris's method results in Fig. 6 numerical case; three RVs stand out



Fig. 8 Convergence of SC (seven points needed) and MC (10,000 points)

The number of eight RVs (here) is too important to use directly the SC as showed previously in Table 1 (>49 million). So, the Morris method is applied. The result is shown in Fig. 7.

Only three variables can be considered as predominant. So the SC can be used on these entries. The convergence of the SC with these three RVs is studied (we use five, seven, and nine collocation points) and compared to the convergence value of MC including all the RVs as a reference method (Fig. 8).

The SC converges with seven collocation points and with a good agreement with global MC. Despite the relatively heavy computational cost of MC (10,000 realizations) and the fact that MC has not yet completely converged, this brings a reference result, and the SC data coincide well with it. Thus, withdrawing five negligible RVs did not alter dramatically the accuracy of the simulations.

On this numerical case ("academic" but close to TR realistic setup), the variability is important (CV = 70 %), and there is a significant impact on the TR focalization (AR = 63 %). The number of realizations needed is 270 (Morris) + 1,195 (SC) = 1,465, which represents a great saving in comparison to the direct SC (49,208,705) or the 10,000 points of the MC method. The method could have been much more efficient in the presence of only two RVs (270 + 153 = 423 realizations) or even one RV (270 + 19 = 289 realizations).

5 Conclusion

We have used a simple, precise, and effective method to deal with uncertainties, even for some problems with a high number of RVs. If the entries are easy to distinguish in the Morris step, the converged mean of the SC gives a good approximation of the real result. It is non-intrusive so it can be applied in various domains with different data (numerical, experimental, etc.). In our context, the validity domain of the TR can be efficiently determined. Further works are planned in order to use this process with heavier numerical test cases (for instance, relying on TR electromagnetic bazooka [12] experiments) to complete experimental knowledge.

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Transmission-Line Super Theory as Antenna Theory for Linear Structures

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Abstract A new generalized transmission-line theory is presented to treat multiconductor as well as antenna systems. Maxwell's equations are cast into the form of classical telegrapher's equations. Two quite different examples are calculated to illustrate the wide use of this theory.

Keywords Transmission-line theory • Antenna theory • Electromagnetic compatibility

1 Introduction

Antennas and transmission lines play an important role in electromagnetic compatibility (EMC). They are necessary for connecting electronic components and devices, but, at the same time, they are ideally suited for gathering unwanted electromagnetic energy from their environment. This coupled energy is often the reason for the failure of electronic devices. Since nowadays, the packing density of electronic components is increasing with a simultaneous increase of the used frequencies, it also seems necessary to extend and adapt the existing theories for estimating energy coupling at high frequencies to non-uniform linear structures. In this context, it can be definitely stated that classical transmission-line theory is no longer sufficient. This was the reason for the development of the presented new, generalized transmission-line theory for linear structures presented here which enables one to deal with cables and antennas alike.

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2 Fundamentals of Transmission-Line Super Theory and Its Application

2.1 Geometrical Description of a System of Non-uniform Conductors

Consider a total number of N non-uniform conductors which are geometrically described by three-dimensional curves $\mathbf{C}_i(l_i)$ (i = 1, 2, ..., N) and are parameterized by their natural parameter, the arc length l_i . All conductors have circular cross sections of the same small radius r_0 and an individual total length of L_i . The curves represent the centre lines of the conductors. Then one can represent the surface of the transmission lines using the so-called Frenet frame which is composed of three vectors: the tangential unit vector $\mathbf{T}_i(l_i)$, the normal unit vector $\mathbf{N}_i(l_i)$, and the binormal unit vector $\mathbf{B}_i(l_i) = \mathbf{T}_i(l_i) \times \mathbf{N}_i(l_i)$. These vectors are derived from the centre curves as

$$\mathbf{T}_{i}(l_{i}) = \frac{\partial \mathbf{C}_{i}(l_{i})}{\partial l_{i}},\tag{1a}$$

$$\mathbf{N}_{i}(l_{i}) = \frac{1}{\kappa_{i}(l_{i})} \frac{\partial \mathbf{T}_{i}(l_{i})}{\partial l_{i}}$$
(1b)

with the quantity $\kappa_i(l_i)$ being the curvature of the *i*th conductor. Then the surface points of the conductors are expressed by

$$\mathbf{C}_{i}^{s}(l_{i},\alpha) = \mathbf{C}_{i}(l_{i}) + \mathbf{N}_{i}(l_{i})r_{0}\cos\alpha + \mathbf{B}_{i}(l_{i})r_{0}\sin\alpha, \quad \alpha \in [0, 2\pi].$$
(2)

Since all of the conductors are circular tubes, the tangential vectors along the centre lines are also the tangential vectors to the surfaces, i.e.

$$\frac{\partial \mathbf{C}_{i}^{S}(l_{i})}{\partial l_{i}} = \frac{\partial \mathbf{C}_{i}(l_{i})}{\partial l_{i}} = \mathbf{T}_{i}(l_{i}).$$
(3)

The last two summands on the right-hand side of Eq. (2) vanish by differentiation with respect to l_i .

There is, however, one situation where Eq. (1b) cannot be applied to the normal unit vector: In the case of a straight wire or wires with linear curve parts, the curvature is zero and Eq. (1b) is not defined. Then one has to obtain the necessary normal vector in a different way. A simple construction of $N_i(l_i)$ is obtained by a 90° rotation of the tangent vector $T_i(l_i)$.

In the following examples, the circular wires will be conducted above a perfectly conducting plane. Therefore, one also needs the mirrored curves $\widetilde{\mathbf{C}}_i(l_i)$ and the mirrored tangential vectors $\widetilde{\mathbf{T}}_i(l_i)$, which are easily obtained from the

original quantities. Furthermore it is assumed that the transmission lines are lossless and can be treated in the thin-wire approximation [1, 2], i.e. only axial currents are considered, and they are concentrated on the axes (centre lines) of the wires. Only the boundary conditions [see Eq. (12)] have to be measured on the boundary of the conductors for one direction α , e.g. $\alpha = 0^{\circ}$. Under these restrictions, all physical equations are presented below.

In order to express the upcoming equations in a more compact matrix form, another formal step is necessary: the introduction of bijective mappings which transform all arc lengths l_i to only one parameter l as follows:

 $l_i = l_i(l)$ for $\forall l_i(0 \le l \le L)$ with $l_i(0) = 0$ and $l_i(L) = L_i$. (4)

A simple example for such mappings is

$$l_i = \frac{L_i}{L}l$$
 with $L = 1 \,\mathrm{m}$ and $l \in [0, 1]$, or $L = \frac{1}{N} \sum_{i=1}^{N} L_i$. (5)

Of course, if one uses this new single parameter, all equations have to be represented with respect to l like the derivative of the potential

$$\frac{\partial \varphi_i(l_i)}{\partial l_i} = \frac{\partial \varphi_i(l)}{\partial l} \left(\frac{dl_i(l)}{dl}\right)^{-1}.$$
(6)

Note that the mappings in Eqs. (4) or (5) are not unique. They are only introduced to simplify the mathematical representations and operations. Therefore, all physical quantities that are represented with respect to this common parameter have no direct physical meaning. After having obtained the final desired results, they have to be transformed back into the physical space where they represent observable quantities.

2.2 Derivation of the New Telegrapher's Equation (Antenna Equation) from the Mixed-Potential Integral Equation

The scattered electrical potential $\varphi(r(l_i))$ can be written with the aid of the scalar Green's function of free space

$$G(r) = \frac{\exp(-jkr)}{r} (j \text{ is the imaginary unit})$$
(7)

as

R. Rambousky et al.

$$\varphi(\mathbf{r}) = \frac{1}{4\pi\varepsilon_0} \sum_{i=1}^{N} \int_0^{L_i} q'_i \left(l'_i \right) \left(G\left(\left| \mathbf{r} - \mathbf{C}_i \left(l'_i \right) \right| \right) - G\left(\left| \mathbf{r} - \widetilde{\mathbf{C}}_i \left(l'_i \right) \right| \right) \right) dl'_i.$$
(8)

In Eq. (8) the charge per unit length is related to the mirrored charge per unit length by $\tilde{q}'_i(l_i) = -q'_i(l_i)$.

Using the continuity equation

$$q'_{i}(l_{i}) = -\frac{1}{j\omega} \frac{\partial i_{i}(l_{i})}{\partial l_{i}} (i_{i} \text{ denotes the axial current along wire } i), \qquad (9)$$

Eq. (8) can be rewritten as

$$\varphi(\mathbf{r}) = -\frac{1}{j4\pi\epsilon_0\omega} \sum_{i=1}^N \int_0^{L_i} \frac{\partial i_i(l_i)}{\partial l_i} \Big(G\Big(\Big|\mathbf{r} - \mathbf{C}_i\Big(l_i'\Big)\Big|\Big) - G\Big(\Big|\mathbf{r} - \widetilde{\mathbf{C}}_i\Big(l_i'\Big)\Big|\Big) \Big) dl_i'.$$
(10)

Similar to the scalar potential, the magnetic vector potential can be expressed as

$$\mathbf{A}(\mathbf{r}) = \frac{\mu_0}{4\pi} \sum_{i=1}^{N} \int_0^{L_i} i_i \left(l'_i \right) \left(\mathbf{T}_i \left(l'_i \right) G\left(\left| \mathbf{r} - \mathbf{C}_i \left(l'_i \right) \right| \right) - \widetilde{\mathbf{T}}_i \left(l'_i \right) G\left(\left| \mathbf{r} - \widetilde{\mathbf{C}} \left(l'_i \right) \right| \right) \right) dl'_i.$$
(11)

Now, the potential equations (10) and (11) allow for expressing the total electrical field (exciting and scattered) on the surface of the conductors in curvilinear coordinates as (boundary condition with $\alpha = 0^{\circ}$)

$$\left(\mathbf{E}^{\text{total}}(\mathbf{r} = \mathbf{C}_i(l_i) + \mathbf{N}_i(l_i)r_0\right)_{l_i} = 0 = \mathbf{T}_i(l_i)\mathbf{E}^{\text{total}}(l_i).$$
 (12)

In explicit form and after some rearrangement, Eq. (12) reads

$$\frac{\partial \varphi_{i}(l_{i})}{\partial l_{i}} + j\omega \frac{\mu_{0}}{4\pi} \sum_{j=1}^{N} \int_{0}^{L} i_{i} \left(l_{j}^{'} \right) \left[\mathbf{T}_{i}(l_{i}) \cdot \mathbf{T}_{j} \left(l_{j}^{'} \right) G_{ij} \left(l_{i}, l_{j}^{'}; k \right) - \mathbf{T}_{i}(l_{i}) \cdot \widetilde{\mathbf{T}}_{j} \left(l_{j}^{'} \right) \widetilde{G}_{ij} \left(l_{i}, l_{j}^{'}; k \right) \right] dl_{j}^{'}$$

$$= \mathbf{E}^{\text{exc}}(l_{i}) \cdot \mathbf{T}_{i}(l_{i}).$$
(13)

Here the abbreviations

$$G_{ij}(l_i, l'_j; k) := G(\left|\mathbf{C}_i(l_i) + \mathbf{N}_i(l_i)r_0 - \mathbf{C}_j(l'_j)\right|)$$

and

$$\widetilde{G}_{ij}\left(l_i, l'_j; k\right) := G\left(\left|\mathbf{C}_i(l_i) + \mathbf{N}_i(l_i)r_0 - \widetilde{\mathbf{C}}_j\left(l'_j\right)\right|\right)$$
(14)

are used.

Similarly, Eq. (10) is rewritten on the boundary of the conductor *i* as

. .

$$\varphi_i(l_i) + \frac{1}{j4\pi\varepsilon_0\omega} \sum_{j=1}^N \int_0^{L_j} \frac{\partial i_j(l_j')}{\partial l_j'} \Big[G_{ij}\Big(l_i, l_j'; k\Big) - \widetilde{G}_{ij}\Big(l_i, l_j'; k\Big) \Big] dl_j' = 0$$
(15)

Using at this stage the general parameter l as introduced in Eqs. (4), (5), and (6), Eqs. (13) and (15) can be represented in a compact matrix form

$$\frac{\partial \boldsymbol{\varphi}}{\partial l} + j\omega \frac{\mu_0}{4\pi} \int_0^L \mathbf{G}_L(l,l';k) \cdot \mathbf{i}(l') dl' = \mathbf{v}_{\text{exc}}(l)$$
(16)

$$\int_{0}^{L} \mathbf{G}_{C}\left(l, l'; k\right) \cdot \frac{\partial \mathbf{i}(l')}{\partial l} dl' + j\omega 4\pi \varepsilon_{0} \mathbf{\phi}(l) = \mathbf{0}.$$
(17)

The above matrices and vectors are defined as follows:

$$\mathbf{\phi}(l) := (\varphi_1(l), \varphi_2(l), \dots, \varphi_N(l))^{\mathrm{T}}, \mathbf{i}(l) := (i_1(l), i_2(l), \dots, i_N(l))^{\mathrm{T}}$$
(18)

and

$$\mathbf{v}_{\text{exc}}^{\prime}(l) := \left(\mathbf{E}^{\text{exc}}(\mathbf{C}_{1}(l)) \cdot \mathbf{T}_{1}(l), \mathbf{E}^{\text{exc}}(\mathbf{C}_{2}(l)) \cdot \mathbf{T}_{2}(l), \dots, \mathbf{E}^{\text{exc}}(\mathbf{C}_{N}(l) \cdot \mathbf{T}_{N}(l))\right)^{\text{T}},$$
(19)

$$\left(\mathbf{G}_{L}\left(l,l';k\right) \right)_{ij} := \mathbf{T}_{i}(l) \cdot \mathbf{T}_{j}\left(l'\right) G\left(\left| \mathbf{C}_{i}(l) + \mathbf{N}_{i}(l)r_{0} - \mathbf{C}_{j}\left(l'\right) \right| \right) - \mathbf{T}_{i}(l) \cdot \widetilde{\mathbf{T}}_{j}\left(l'\right) G\left(\left| \mathbf{C}_{i}(l) + \mathbf{N}_{i}(l)r_{0} - \widetilde{\mathbf{C}}_{j}\left(l'\right) \right| \right)$$

$$(20)$$

and

$$\left(\mathbf{G}_{C}\left(l,l';k\right)\right)_{ij} := G\left(\left|\mathbf{C}_{i}(l) + \mathbf{N}_{i}(l)r_{0} - \mathbf{C}_{j}\left(l'\right)\right|\right) - G\left(\left|\mathbf{C}_{i}(l) + \mathbf{N}_{i}(l)r_{0} - \widetilde{\mathbf{C}}_{j}\left(l'\right)\right|\right).$$
(21a)

Remember, the transformation of all equations into the representation with respect to the general parameter l has to be correctly applied to all operations, coordinate transformation and operator transformation; see, e.g. Eq. (6). Equations (16) and (17) are the so-called mixed-potential integral equation (MPIE), which represent Maxwell's equations in the Lorenz gauge for a non-uniform multiconductor system. Observe, however, that all quantities depend on the

("artificial") parameter l and therefore do not have a direct physical meaning. In addition, the potential φ is not uniquely defined due to its general dependency on the integration contour.

If one replaces the derivative of the potential in Eq. (16) using Eqs. (17) and (9), then Eq. (16) can be written in the form which will be used later:

$$\frac{1}{4\pi\varepsilon_0} \int_0^L \mathbf{G}_C(l,l';k) \mathbf{q}'(l') dl' + j\omega \frac{\mu_0}{4\pi} \int_0^L \mathbf{G}_L(l,l';k) \cdot \mathbf{i}(l') dl' = \mathbf{v}_{\text{exc}}(l).$$
(21b)

In the low-frequency limit $(k \rightarrow 0)$, when radiation is still absent, the currents and their derivatives can be pulled out of the integrals in Eq. (21b), because the phase-independent Green's functions have very strong peaks at the values l' = land therefore act almost like delta functions, resulting in

$$\left(\int_{0}^{l} \mathbf{G}_{L}(l,l';0) dl'\right) \cdot \mathbf{i}(l) =: \mathbf{G}_{L}^{\prime 0}(l) \cdot \mathbf{i}(l), \qquad (22a)$$

$$\left(\int_{0}^{L} \mathbf{G}_{C}(l, l'; 0) dl'\right) \cdot \frac{\partial}{\partial l} \mathbf{i}(l) =: \mathbf{G}_{C}^{\prime 0}(l) \cdot \frac{\partial}{\partial l} \mathbf{i}(l).$$
(22b)

Then the MPIE can be simplified and written in a more compact and familiar form

$$\begin{pmatrix} \frac{\partial}{\partial l} \boldsymbol{\varphi}(l) \\ \frac{\partial}{\partial l} \mathbf{i}(l) \end{pmatrix} + j\omega \begin{pmatrix} \mathbf{0} & \mathbf{L}'(l) \\ \mathbf{C}'(l) & \mathbf{0} \end{pmatrix} \cdot \begin{pmatrix} \boldsymbol{\varphi}(l) \\ \mathbf{i}(l) \end{pmatrix} = \begin{pmatrix} \mathbf{v}_{\text{exc}}(l) \\ \mathbf{0} \end{pmatrix}.$$
(23)

This equation is the classical transmission-line equation for multiconductors with the exception that the elements of the parameter matrix (inductances and capacitances) depend on the location. In Eq. (23) the usual definitions have been made:

$$\mathbf{L}'(l) = \frac{\mu_0}{4\pi} \mathbf{G'}_{L}^{0}(l) \tag{24a}$$

and

$$\mathbf{C}'(l) = 4\pi\varepsilon_0 \left(\mathbf{G}'_C^{(0)}(l)\right)^{-1}.$$
(24b)

Also the solution of Eq. (23) is well known [3], even if the block matrix is completely occupied and its elements are all complex valued:

Transmission-Line Super Theory as Antenna Theory for Linear Structures

$$\begin{pmatrix} \mathbf{\varphi}(l) \\ \mathbf{i}(l) \end{pmatrix} = \overline{\mathbf{M}}_{l_0}^l \Big\{ -j\omega \overline{\mathbf{P}}^{*(0)} \Big\} \begin{pmatrix} \mathbf{\varphi}(l_0) \\ \mathbf{i}(l_0) \end{pmatrix} + \int_{l_0}^l \overline{\mathbf{M}}_l^l \Big\{ -j\omega \overline{\mathbf{P}}^{*(0)} \Big\} \cdot \mathbf{v}_{\text{exc}} (l') dl'.$$
(25)

The block matrix $\overline{\mathbf{P}}^{*(0)}$ represents the block matrix in Eq. (23). The quantity $\overline{\mathbf{M}}_{l_0}^{l}$ is the so-called Matrizant or product integral [3]. In order to find an explicit solution for Eq. (25), one needs to know $\overline{\mathbf{P}}^{*(0)}$ and the exciting sources. The general procedure for how to solve Eq. (25) will be given below.

Note that the first-order differential equation system Eq. (23) can be written as a second-order wave equation for the current.

$$\left(\frac{\partial^2}{\partial l^2} + \mathbf{C}' \cdot \frac{\partial \mathbf{C}'^{-1}}{\partial l} \frac{\partial}{\partial l} + \omega^2 \mathbf{C}' \cdot \mathbf{L}'\right) \cdot \mathbf{i}(l) = -j\omega \mathbf{C}' \mathbf{v}_{\text{exc}}(l).$$
(26)

Remember, this equation is valid in the low-frequency approximation. New in this otherwise classical equation is the occurrence of the first derivative of the current vector due to losses caused by reflections along the lines. As expected, the solutions of Eq. (26) describe forward and backward running current waves. This result gives rise to the assumption that even in the exact case (all frequencies are allowed), the current fulfils a second-order wave equation of the general form

$$\left(\frac{\partial^2}{\partial l^2} + j\omega \mathbf{P}_{11}(l) \cdot \frac{\partial}{\partial l} + \omega^2 \mathbf{P}_{12}(l)\right) \cdot \mathbf{i}(l) = -j\omega \mathbf{q}_{\text{exc}}''(l).$$
(27)

With the aid of the continuity equation (9), this equation can be converted into a coupled first-order differential equation system:

$$\frac{\partial}{\partial l} \begin{bmatrix} \mathbf{q}'(l) \\ \mathbf{i}(l) \end{bmatrix} + j\omega \underbrace{\begin{bmatrix} \mathbf{P}_{11}(l) & \mathbf{P}_{12}(l) \\ \mathbf{1} & \mathbf{0} \end{bmatrix}}_{\overline{\mathbf{P}}(l)} \begin{bmatrix} \mathbf{q}'(l) \\ \mathbf{i}(l) \end{bmatrix} = \begin{bmatrix} \mathbf{q}''_{\text{exc}}(l) \\ \mathbf{0} \end{bmatrix}.$$
(28)

In Eq. (28) the charge per unit length, \mathbf{q}' , now appears as the physical quantity instead of the potential. But again, the solution of Eq. (28) is known:

$$\begin{pmatrix} \mathbf{q}'(l) \\ \mathbf{i}(l) \end{pmatrix} = \overline{\mathbf{M}}_{l_0}^l \{-j\omega\overline{\mathbf{P}}\} \begin{pmatrix} \mathbf{q}'(l_0) \\ \mathbf{i}(l_0) \end{pmatrix} + \int_{l_0}^l \overline{\mathbf{M}}_l^l \{-j\omega\overline{\mathbf{P}}\} \cdot \mathbf{q}_{\text{exc}}''(l) dl'.$$
(29)

The next subsection will show how the Matrizant and the parameter matrix are calculated by an iterative procedure.

3 Iterative Methods for the Solution of the Matrizant and Solution of the TLST Equations

In the next step towards the generalized telegrapher's equation, it will be proven that Eq. (28) indeed follows from the MPIE Eqs. (21b) and (9). For this purpose, Eq. (21b) is rearranged as

$$\begin{bmatrix} j\omega \mathbf{1} & \mathbf{1}\frac{\partial}{\partial l} \end{bmatrix}_{(N,2N)} \int_{0}^{L} \overline{\mathbf{G}}\left(l,l';k\right) \begin{bmatrix} \mathbf{q}'(l')\\ \mathbf{i}(l') \end{bmatrix} dl' = \mathbf{v}_{\text{exc}}'(l).$$
(30)

Here the block matrix is given by

$$\overline{\mathbf{G}} := \begin{bmatrix} \mathbf{0} & \frac{\mu_0}{4\pi} \mathbf{G}_L \\ 4\pi\varepsilon_0 \mathbf{G}_C & \mathbf{0} \end{bmatrix}.$$
(31)

Then Eq. (29) is inserted into Eq. (30) and after some minor conversions becomes

$$\mathbf{I}_{21}\frac{\partial \mathbf{q}'}{\partial l} + \mathbf{I}_{22}\frac{\partial \mathbf{i}}{\partial l} + \left(j\omega\mathbf{I}_{11} + \frac{\partial \mathbf{I}_{21}}{\partial l}\right)\mathbf{q}' + \left(j\omega\mathbf{I}_{12} + \frac{\partial \mathbf{I}_{22}}{\partial l}\right)\mathbf{i} + j\omega\mathbf{I}_{01} + \frac{\partial \mathbf{I}_{02}}{\partial l} = \mathbf{v}_{\text{exc}}^{'}.$$
(32)

Note that Eq. (30) [and therefore also Eq. (32)] implicitly contains the continuity equation. Here the components I_{ij} are elements of the block matrices

$$\overline{\mathbf{I}}(l) = \int_{0}^{L} \overline{\mathbf{G}}\left(l, l'\right) \overline{\mathbf{M}}_{l}^{l'} \left\{-j\omega\overline{\mathbf{P}}\right\} dl'$$
(33a)

$$\begin{bmatrix} \mathbf{I}_{10}(l) \\ \mathbf{I}_{20}(l) \end{bmatrix} = \int_{0}^{L} \overline{\mathbf{G}}\left(l, l'\right) \int_{l}^{l'} \overline{\mathbf{M}}_{l''}^{l'} \left\{-j\omega\overline{\mathbf{P}}\right\} \begin{bmatrix} \mathbf{q}_{s}^{''}\left(l''\right) \\ \mathbf{0} \end{bmatrix} dl'' dl'.$$
(33b)

Collecting now the summands in Eq. (32) in terms of block vectors and block matrices, one indeed obtains Eq. (28) and thereby confirming the assumption that the current obeys a wave equation with the following parameters and renormalized source term:

$$\overline{\mathbf{P}} = \begin{bmatrix} \mathbf{I}_{21} & \mathbf{I}_{22} \\ \mathbf{0} & \mathbf{1} \end{bmatrix}^{-1} \begin{bmatrix} \mathbf{I}_{11} + \frac{1}{j\omega} \frac{\partial \mathbf{I}_{21}}{\partial l} & \mathbf{I}_{12} + \frac{1}{j\omega} \frac{\partial \mathbf{I}_{22}}{\partial l} \\ \mathbf{1} & \mathbf{0} \end{bmatrix}$$
(34a)

Transmission-Line Super Theory as Antenna Theory for Linear Structures

$$\mathbf{q}_{\text{exc}}^{''} = \mathbf{I}_{21}^{-1} \left(\mathbf{v}_{\text{exc}}^{'} - j\omega \mathbf{I}_{10} - \frac{\partial \mathbf{I}_{20}}{\partial l} \right).$$
(34b)

In other words, Maxwell's equations applied to linear thin conducting structures have been cast into the form of the transmission-line equations which are known from classical transmission-line theory (TLT). No restrictions are made on frequencies or heights of the lines above ground. Thus, in particular, radiation is included. There is, however, an essential difference compared to classical TLT: The parameter matrix is fully occupied, and its elements are complex valued. This can be better seen if the Eq. (28) is transformed into the potential-current representation, which is done with the aid of the Eqs. (9) and (17) as well as with the relations

$$\boldsymbol{\varphi} = \mathbf{I}_{21} \mathbf{q}' + \mathbf{I}_{22} \mathbf{i} + \mathbf{I}_{20} \tag{35a}$$

or

$$\begin{bmatrix} \boldsymbol{\varphi} \\ \mathbf{i} \end{bmatrix} = \begin{bmatrix} \mathbf{I}_{21} & \mathbf{I}_{22} \\ \mathbf{0} & \mathbf{1} \end{bmatrix} \begin{bmatrix} \mathbf{q}' \\ \mathbf{i} \end{bmatrix} + \begin{bmatrix} \mathbf{I}_{20} \\ \mathbf{0} \end{bmatrix}$$
(35b)

and results in

$$\frac{\partial}{\partial l} \begin{bmatrix} \boldsymbol{\varphi}(l) \\ \mathbf{i}(l) \end{bmatrix} + j\omega \overline{\mathbf{P}}^*(l) \begin{bmatrix} \boldsymbol{\varphi}(l) \\ \mathbf{i}(l) \end{bmatrix} = \begin{bmatrix} \boldsymbol{\varphi}'_s(l) \\ \mathbf{i}'_s(l) \end{bmatrix}$$
(36a)

with

$$\overline{\mathbf{P}}^{*} = \begin{bmatrix} \mathbf{I}_{11}\mathbf{I}_{21}^{-1} & \mathbf{I}_{12} - \mathbf{I}_{11}\mathbf{I}_{21}^{-1} \\ \mathbf{I}_{21}^{-1} & -\mathbf{I}_{21}^{-1}\mathbf{I}_{22} \end{bmatrix}$$
(36b)

and

$$\begin{bmatrix} \mathbf{\phi}'_{s} \\ \mathbf{i}'_{s} \end{bmatrix} = \begin{bmatrix} \mathbf{v}'_{\text{exc.}} + j\omega (\mathbf{I}_{11}\mathbf{I}_{21}^{-1}\mathbf{I}_{20} - \mathbf{I}_{10}) \\ j\omega \mathbf{I}_{21}^{-1}\mathbf{I}_{20} \end{bmatrix}.$$
 (37)

With the knowledge of the block matrix $\overline{\mathbf{I}}$, one can determine the parameter matrices $\overline{\mathbf{P}}$ and $\overline{\mathbf{P}}^*$ in the charge-current representation (28) and in the potential-current representation Eq. (36a), respectively. To calculate the parameter matrix, it is preferable to use the charge-current representation, and in a last step, the current is obtained from Eq. (36a) with the chosen boundary conditions (see below).

To solve Eqs. (28), (34a), and (34b), one needs to know the parameter matrix $\overline{\mathbf{P}}$ and the Matrizant $\overline{\mathbf{M}}_{l}^{l} \{-j\omega \overline{\mathbf{P}}\}$, both operators enter the key Eqs. (33a) and (33b) for $\overline{\mathbf{I}}$ and the block vector $[\mathbf{I}_{10} \quad \mathbf{I}_{20}]^{T}$. However, both depend on each other in an implicit equation. Thus, one may start an iteration procedure to calculate $\overline{\mathbf{I}}$,

beginning with $\overline{\mathbf{M}}_{l}^{l} \{-j\omega \overline{\mathbf{P}}\} = \overline{\mathbf{1}}$ and $\begin{bmatrix} \mathbf{I}_{10}^{(0)} & \mathbf{I}_{20}^{(0)} \end{bmatrix}^{T} = \begin{bmatrix} \mathbf{0} & \mathbf{0} \end{bmatrix}^{T}$ in a general series expansion for $\overline{\mathbf{I}}$

$$\overline{\mathbf{I}} = \overline{\mathbf{I}}_0 + j\omega\overline{\mathbf{I}}_1 + \dots \tag{38}$$

In the low-frequency limit ($\omega \rightarrow 0$) together with the above assumptions, this results in

$$\overline{\mathbf{I}}_{0}(l) \equiv \overline{\mathbf{I}}^{(0)}(l) = \int_{0}^{L} \overline{\mathbf{G}}\left(l, l'; k\right) \big|_{k \to 0} dl'.$$
(39)

Now, the calculations are performed following the chain [4]:

$$\overline{\mathbf{I}}^{(0)}(l) \underset{(34)}{\longrightarrow} \overline{\mathbf{P}}^{(0)}(l) \underset{[3]}{\longrightarrow} \overline{\mathbf{M}}_{l}^{l} \left\{ -j\omega \overline{\mathbf{P}}^{(0)} \right\} \underset{(33)}{\longrightarrow} \overline{\mathbf{I}}^{(1)}(l) \underset{(34)}{\longrightarrow} \overline{\mathbf{P}}^{(1)} \underset{[3]}{\longrightarrow} \overline{\mathbf{M}}_{l}^{l} \left\{ -j\omega \overline{\mathbf{P}}^{(1)} \right\} \rightarrow \dots$$

$$(40)$$

This iterative procedure is terminated when convergence is achieved. In practical applications [4, 5], convergence was already reached after the first iteration with $\overline{\mathbf{I}}^{(1)}$. Then from $\overline{\mathbf{I}}^{(1)}$ one finds $\overline{\mathbf{P}}^{*(1)}$ (Eq. (36b)) and can solve Eq. (36a) to get the final result for current and potential.

4 Radiated Power

The average power radiated by a lossless multiconductor system above conducting ground, which is excited by lumped sources, was derived in [6]. This power is obtained by the difference of the power that is fed into the system at the beginning of the conductors and the power that arrives at the end of the conductors:

$$P_{\rm rad} = -\frac{1}{2} \operatorname{Re} \sum_{i=1}^{N} \int_{0-\Delta}^{L_i-\Delta} \frac{d}{dl_i} (\varphi_i(l_i) i_i^*(l_i)) dl_i \text{ with } \Delta \to 0.$$
(41)

When transforming Eq. (41) into the matrix notation, it becomes via Eq. (36a)

$$P_{\rm rad} = j\frac{\omega}{4} \int_0^L \left[\boldsymbol{\varphi}^+ (\mathbf{P}_{22}^* - \mathbf{P}_{11}^{*+}) \mathbf{i} + \mathbf{i}^+ (\mathbf{P}_{11}^* - \mathbf{P}_{22}^{*+}) \boldsymbol{\varphi} + \mathbf{i}^+ (\mathbf{P}_{12}^* - \mathbf{P}_{12}^{*+}) \mathbf{i} + \boldsymbol{\varphi}^+ (\mathbf{P}_{21}^* - \mathbf{P}_{21}^{*+}) \boldsymbol{\varphi} \right] dl.$$
(42)

The quantities which carry the upper index ⁺ denote the transposed complex conjugate vectors and matrices. The formula Eq. (42) simplified to one conductor reads

$$P_{\rm rad} = -\frac{\omega}{2} \int_0^L \left[\mathrm{Im}(P_{12}^*) |i|^2 + \mathrm{Im}(P_{21}^*) |\varphi|^2 + \mathrm{Im}(\varphi(P_{11}^* - P_{22}^{*(*)}) i^*) \right].$$
(43)

The element $P_{22}^{*(*)}$ is the complex conjugate of P_{22}^{*} . It can be recognized from Eq. (42) that all block matrices \mathbf{P}_{ij}^{*} contribute to the radiated power with both their real and their imaginary parts.

5 **Application Examples**

Two completely different application examples are investigated: a simple transmission line above ground with vertical risers and a linear antenna with different angles relative to the ground plane. In the two examples, the line and the antenna are both fed by a voltage source of 1 V at their near ends and have a load impedance of 50 Ω . Figures 1 and 2 show their configuration and their geometry. The transmission line is terminated with its (classical) characteristic impedance of 359 Ω . The antenna is open ended.

Figures 3 and 4 display the real parts of the inductances per unit length along the wire/antenna. In Fig. 3 it can be seen that this parameter shows a symmetrical behaviour relative to the central point of the line. For low frequencies $\lambda \gg h$ around this central point, the inductance per unit length equals the value of classical transmission-line theory $L_0 = (\frac{\mu_0}{2\pi}) \ln(\frac{2h}{r_0})$. Comparing a point from the central





Fig. 3 The real part of the inductance per unit length along the TL at 1 GHz



Fig. 4 The real part of the inductance per unit length along the antenna

part of the wire where the wire is present at both sides, with the bend points, it can be seen that in the latter the line is mainly present at only one side. Consequently, the inductance decreases to its minimum values. In Fig. 4 the real part of the inductance per unit length shows a very similar behaviour for $\alpha = 0$ as in Fig. 3,



Fig. 5 Imaginary part of the capacitance per unit length along the TL at 1 GHz



Fig. 6 Imaginary part of the capacitance per unit length along the antenna

except at the right end, due to the missing riser. However, with growing angle size, the inductance increases and starts to oscillate with the eigenvalue $\lambda/2$ of the current, caused by increasing radiation losses. The biggest inductance occurs at 90°.

Unlike in the previous case, Figs. 5 and 6 now represent the graphs of those quantities which do not have an analogue in classical transmission-line theory:



Fig. 7 Current along the TL

namely, the imaginary parts of the complex capacitance. Although their values are relatively small, they nevertheless contribute to the radiation losses of the otherwise lossless line. Note that at the bend points of the wire and antenna, the imaginary parts of the capacitances take their maxima, indicating that at these points, an essential part of radiation is created. Also oscillations along the lines with $\lambda/2$ eigenvalue can be observed. For the antenna, these oscillations increase with the angle α .

As soon as the parameter block matrix is known, the super transmission-line Eq. (36a) can be solved by applying known methods [5] to obtain the current for the given boundary conditions. In the case of the transmission-line configuration, the wire is terminated at the far end by its classical (constant) characteristic TL impedance. Note, however, that in the present case, the expression $(P_{12}^*/P_{21}^*)^{0.5}$ is a function of the local coordinate and frequency. Therefore, the question arises, "At which location shall this value be chosen?" In [7] it was shown that the maximum of the above square root along the line (which equals the constant classical characteristic impedance value) led to the least oscillations of the current along the wire. A constant value of the current like in classical TL theory cannot be achieved due to reflections at all non-uniformities. These reflections, of course, cause standing waves of the wire and antenna. In Figs. 7 and 8, the currents ring with frequencies of 70 MHz and 150 MHz, respectively, according to their $\lambda/2$ eigenvalues which correspond to their total length. As can be seen in Fig. 8, the angle α has only a minor influence on the very pronounced current amplitude oscillations along the antenna.

Figures 9 and 10 display the radiated power from the TL and from the antenna at different angles. It can be observed that the radiated power grows with frequency.



Fig. 8 Current along the antenna



Fig. 9 Radiated power of the TL



Fig. 10 Radiated power of the antenna at different angles

Due to the resonance phenomena, the power oscillates with the eigenfrequency of the wire (70 MHz) and the antenna (150 MHz), respectively. Since the transmission line is matched with its "characteristic impedance", the power oscillations are smaller than in the open-ended antenna. The influence of the angle of the antenna on the radiated power becomes more noticeable at lower frequencies ($f \le 900$ MHz) than at higher frequencies. But above 45°, the maximum power only changes marginally for frequencies ≥ 1 GHz.

6 Conclusion

It was shown that the TLST can also be used as an antenna theory for linear structures. The derived Eqs. (36a), (36b), and (37) are Maxwell's equations for a multi-wire system. When the potential and current for such a system are known, one can use these quantities to calculate the total electric field and the magnetic field [5], and, therefore, the Poynting vector as well.

The radiated power results from the parameter block matrices, the currents and the potentials. In its structure, this new theory resembles the classical transmissionline theory very much. There are, however, essential differences: The transmissionline parameters are now complex and are frequency and location dependent. They depend on the chosen gauge and coordinate system (see [4]) and, therefore, lose their physical meaning, unlike in the classical theory. The above examples show the wide range of possible applications of the new theory.

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Electromagnetic Coupling to Transmission Lines with Symmetric Geometry Inside Rectangular Resonators

S. Tkachenko, J. Nitsch, and R. Rambousky

Abstract In this chapter, the analysis of transmission lines inside rectangular resonators is extended from one conductor with two sources/loads at the ends to many loads along the conductor and to the interaction between two conductors. To do this, the many loads are described as passive small (δ -) sources, whose corresponding electrical fields superpose with fields from other sources. The currents in the individual loads are found by imposing the boundary condition for the total electrical field on the surface of the conductor and applying a Fourier series expansion for the current. The calculation of the interaction of two lines in the resonator is based on the known description for one conductor and the two-dimensional Green's function, which occurs in the analytical expressions for the currents. This Green's function is used for both conductors separately. Then, with the aid of the superposition principle, the individual currents for the lines are derived. These computational procedures are illustrated by two examples.

Keywords Transmission line • Cavity resonator • Analytical solutions • Thin wires

1 Introduction

Cables and interconnections play an important role in EMC. Not only do they transport information and energy in electromagnetic and electronic systems, but they also frequently serve as collectors for unwanted electromagnetic (em) energy. This unwanted energy then is transferred to the ports of receivers and may enter

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sensitive parts of electronics causing failure or even damage of the devices. In general, electrical engineering equipment is tested according to EMC standards, which usually happens in appropriate laboratories. Therefore, these devices already contain some basic protection against electromagnetic interference. Finally, the tested individual devices can be integrated in larger systems that may be located in shielded rooms. Hence, the final environment of the tested device usually differs in several respects from the original test environment, and consequently, the possible em interference can also differ from the test conditions. How strongly the range of threats can vary over the test spectrum is studied in this chapter on two examples: a loaded wire, which is symmetrically located inside a rectangular resonator and excited by distributed and/or lumped sources, as well as two parallely conducted wires in the resonator. This is done both analytically and numerically. First, the already published and necessary basic equations are summarized, followed by novel results for the currents of a multi-loaded wire and for the interaction between two conductors inside a resonator. These analytical results are confirmed and supplemented by numerical calculations using numerical software packages. The results are presented and discussed, especially in view of the differences that would result for the analog configurations of the wires above ground without any shielding. The chapter concludes with final remarks and prospects for future work.

2 Exact Analytical Solution of EFIE for Thin-Wire Structures Keeping the Symmetry of the Resonator

In [1, 2] the method of small antennas (MSA) was described and applied to small thin scatterers (small electric and magnetic dipoles) inside rectangular resonators. In principle, this method is also applicable to long thin wires of arbitrary geometrical configuration inside resonators [3]. However, the singular and regular parts of the Green's function are only represented in an approximate form. Therefore, it is of interest to perform a comparison with an exact solution. Such a comparison is the subject of this section.

As known from theoretical physics, exact analytical solutions of a problem can be found in a simpler way, if one deals with symmetrical configurations. Such a configuration is also assumed for the problem "Long conductor inside a resonator" (see Fig. 1).

The represented wire is conducted parallely to a principal axis of the resonator and connects two opposite walls. The wire is short-circuited at both ends. Based on the geometry shown in Fig. 1, first the necessary component G_{zz}^E

$$G_{zz}^{E}\left(\vec{r},\vec{r}'\right) = \frac{\eta_{0}c}{jk}\left(k^{2} + \frac{\partial^{2}}{\partial z^{2}}\right)G_{zz}^{A}\left(\vec{r},\vec{r}'\right)$$
(1)



of the dyadic of the Green's function for the electrical field is derived with the aid of the mirror principle [4]. Here G_{zz}^A is the *zz*-component of the of the Green's function for the vector potential \vec{A} of the resonator. The quantity c denotes the speed of light, and $\eta_0 = \sqrt{\mu_0/\varepsilon_0}$. As a result one obtains [5, 6]:

$$G_{zz}^{E}(\vec{r},\vec{r}_{1}) = \frac{4\eta_{0}}{jkV} \sum_{\substack{n_{1},n_{2} = 1 \\ n_{3} = 0}}^{\infty} \frac{\varepsilon_{n_{3},0}(k^{2} - k_{vz}^{2})\sin(k_{vx}x)\sin(k_{vx}x_{1})\sin(k_{vy}y)\sin(k_{vy}y_{1})\cos(k_{vz}z)\cos(k_{vz}z_{1})}{k_{v}^{2} - k^{2} + j\delta}$$
(2)

with the following abbreviations:

$$\vec{k}_{v} = (k_{vx}, k_{vy}, k_{vz}), \ k_{vx} = \frac{\pi}{a}n_{1}, \ k_{vy} = \frac{\pi}{b}n_{2}, \ \text{and} \ k_{vz} = \frac{\pi n_{3}}{h}, \ v := |n_{1}, n_{2}, n_{3}\rangle$$

$$n_{i} = \text{number of reflections in } i\text{-direction and } \delta \to +0, \ \varepsilon_{n,m} = \begin{cases} 1, & m = n \\ 2, & m \neq n \end{cases}.$$
(3)

Using the above equations one can easily establish the relation between the scattered electrical field and the current I(z) induced on the conductor:

$$E_{z}^{\rm sc}\left(\vec{r}\right) = \int_{0}^{h} G_{zz}^{E}\left(\vec{r},\vec{r}'\right) I\left(z'\right) \mathrm{d}z'. \tag{4}$$

Here it was assumed that inside the resonator a field $\vec{E}^{0}(\vec{r})$ has excited the system. Due to the chosen symmetry of the line relative to the walls of the resonator, it is sufficient to use only the *zz*-component of the Green's function \hat{G}^{E} for calculating the scattered current. From Eq. (4) and the boundary conditions for the total electrical field (exciting and scattered), the Fourier components of the current along the line are obtained [5, 7]:

$$I_{n_3} = -\frac{E_{z,n_3}^0(x_0 + r_0, y_0)}{\frac{4\eta_0}{jkab} \sum_{n_1, n_2=1}^{\infty} \frac{\left(k^2 - k_{vz}^2\right) \sin\left(k_{vx}(x_0 + r_0)\right) \sin\left(k_{vx}x_0\right) \sin^2\left(k_{vy}y_0\right)}{k_v^2 - k^2 + j\delta}.$$
 (5)

Remember that the conductor was assumed to be short-circuited at both ends with the corresponding walls, and due to the symmetry also only the E_z^0 -component of the field $\vec{E}^0(\vec{r})$ excites the line. The total current is then obtained from the Fourier series:

$$I(z) = \sum_{n_3=0}^{\infty} I_{n_3} \cos(n_3 \pi z/h)$$
(6)

with the correct boundary conditions $dI/dz|_{z=0} = dI/dz|_{z=h} = 0$.

The more sophisticated part of the analysis now occurs in the calculation of the Fourier components, in particular of the denominator in Eq. (5). To speed up numerical calculations and to simplify the limit to classical transmission-line theory of a cable inside a resonator, first, the double sum in the denominator is rewritten in a more compact form. For this purpose, the function S:

$$S = S\left(k, \vec{\rho}, \vec{\rho}'\right) := \frac{4}{ab} \sum_{n_1, n_2=1}^{\infty} \frac{\sin(k_{\nu x}x)\sin(k_{\nu x}x')\sin(k_{\nu y}y)\sin(k_{\nu y}y')}{k_{\nu \rho}^2 + k_{\nu z}^2 - k^2 + j\delta}$$
(7)

with the abbreviations $\vec{\rho} = (x, y)$, $\vec{k}_{v\rho} = (k_{vx}, k_{vy})$ is defined.

Using this equation, Eq. (5) can be rewritten as

$$I_{n_3} = \frac{jkE_{z,n_3}^0(x_0, y_0)}{\eta_0(k_{v_z}^2 - k^2)S(x = x_0 + r_0, y = y_0, x' = x_0, y' = y_0)}.$$
(8)

A longer calculation results in the following expression for the function S [5, 7]:

Electromagnetic Coupling to Transmission Lines with Symmetric Geometry...

$$S = \frac{1}{2\pi} \sum_{n_1, n_2 = -\infty}^{\infty} (-1)^{n_1 + n_2} \begin{cases} K_0(\gamma_{vz} | \vec{\rho} - \vec{\rho}(n_1, n_2) |), & k_{vz}^2 - k^2 > 0\\ -\frac{j\pi}{2} H_0^{(2)}(\tilde{k}_{vz} | \vec{\rho} - \vec{\rho}(n_1, n_2) |), & k_{vz}^2 - k^2 < 0 \end{cases}$$
(9)

with $\vec{\rho}(n_1, n_2) = (X(n_1), Y(n_2)), X(n_1) = x_1 \cdot (-1)^{n_1} + ((1 - (-1)^{n_1})/2 + n_1) \cdot a$, and

$$Y(n_2) = y_1 \cdot (-1)^{n_2} + ((1 - (-1)^{n_2})/2 + n_2) \cdot b_2$$

Here the functions K_0 and $H_0^{(2)}$ denote cylinder functions and Hankel functions of the second kind, respectively.

Note that the function *S* also has other representations (see [5] and [7]).

Inserting Eq. (9) into Eq. (8) finally leads to the desired result for the current along the line. So far, the cable was assumed to be excited by a distributed field inside the resonator. In the case that in addition to a distributed field concentrated sources also feed the line, then corresponding currents arise. This situation is the subject of the next subsection.

3 One Wire with Many Loads

Consider a single line loaded with several impedances $Z_1, Z_2, ..., Z_M$ at the positions $z_1, z_2, ..., z_M$ along the line (see Fig. 2).

The line is excited by a distributed field of the form (due to the chosen symmetry) [5, 7]:

$$E_z^0\left(\vec{r}\right) = \sum_{n_3=0}^{\infty} E_{z,n_3}^0\left(\vec{\rho}\right) \cdot \cos\left(k_{vz}z\right).$$
(10)

The loads can be regarded and represented as concentrated "sources" and expressed through

$$E_{z,cs}^{0} = \sum_{m=1}^{M} U_m \delta(z - z_m), \text{ and } U_m = -Z_m I(z_m).$$
 (11)

Then, for the total electrical field which excites the conductor, one obtains

$$E_{z,\text{tot}}^{0} = E_{z}^{0}(z) + E_{z,\text{cs}}^{0}.$$
 (12)

Inserting Eqs. (10) and (11) into Eq. (12) and imposing the boundary condition on the total field yield the connection between the Fourier coefficients of the fields:



$$E_{z,\text{tot},n_{3}}^{0} = E_{z,n_{3}}^{0} \left(\overrightarrow{\rho}_{0} \right) + E_{z,\text{cs},n_{3}}^{0}$$
$$= E_{z,n_{3}}^{0} \left(\overrightarrow{\rho}_{0} \right) - \frac{\varepsilon_{n_{3},0}}{h} \sum_{m=1}^{M} Z_{m} I_{m}(z_{m}) \cos\left(k_{z}^{\nu} z_{m}\right).$$
(13)

With the known ansatz for the total current along the line as [5, 7]

$$I(z) = \sum_{n_3}^{\infty} I_{n_3} \cos\left(\frac{\pi z}{h} n_3\right) \quad \text{and} \quad I_{n_3} = \frac{jkE_{z,\text{tot},n_3}^0}{\eta_0 \left(k_{\nu z}^2 - k_0^2\right)},\tag{14}$$

one obtains

$$I(z) = \frac{jk}{\eta_0} \sum_{n_3=0}^{\infty} \frac{E_{zn_3}^0 \cos(k_{vz}z)}{(k_{vz}^2 - k_0^2)S} - \frac{jk}{\eta_0 h} \sum_{m=1}^M Z_m I(z_m) \sum_{n_3=0}^{\infty} \frac{\varepsilon_{n_3,0}}{(k_{vz}^2 - k_0^2)S} \cos(k_{vz}z_m) \cos(k_{vz}z)$$
(15)

or the shorter version

$$I(z) = I_{\rm f}(z) - \sum_{m=1}^{M} Z_m I(z_m) Y_m(z)$$
(16)

with the definitions:

Electromagnetic Coupling to Transmission Lines with Symmetric Geometry...

$$I_{\rm f}(z) = \frac{jk}{\eta_0} \sum_{n_3=0}^{\infty} \frac{E_{zn_3}^0 \cos\left(k_{\nu z} z\right)}{\left(k_{\nu z}^2 - k_0^2\right) S}$$

and

$$Y_m(z) = Y(z, z_m) := \frac{jk}{\eta_0 h} \sum_{n_3=0}^{\infty} \frac{\varepsilon_{n_3,0}}{\left(k_{\nu z}^2 - k_0^2\right) S} \cos\left(k_{\nu z} z_m\right) \cos\left(k_{\nu z} z\right).$$
(17)

Then, substituting $z = z_m$ in Eq. (16), one gets a linear equation system for estimation of the unknown values of the current $I(z_m)$:

$$\sum_{m_1=1}^{M} \left[\delta_{m_1,m} + Z_{m_1} Y(z_{m_1}, z_m) \right] I(z_{m_1}) = I_{\mathbf{f}}(z_m).$$
(18)

This equation has to be solved with respect to the current values at the positions of the impedances. Then, all of the necessary quantities are known and the voltages U_m can be estimated at the locations: z_m ; $U_m = -Z_m I(z_m)$.

4 Interaction of Two Wires Inside a Rectangular Resonator

In this section, the electromagnetic coupling between two parallely conducted cables inside a rectangular resonator is investigated (for geometry, see Fig. 3).

The following analysis is based on the above results, for one conductor in the resonator. Similar equations are valid, but now for the individual conductors. The coupling of the conductors results from their contributions to the total scattered field $E_z^{sc}(\vec{r})$:

$$E_{z}^{\rm sc}(\vec{r}) = \int_{0}^{h} G_{zz}^{E}(\vec{\rho}, z, \vec{\rho}_{01}, z') I_{1}(z') dz' + \int_{0}^{h} G_{zz}^{E}(\vec{\rho}, z, \vec{\rho}_{02}, z') I_{2}(z') dz'.$$
(19)

Fourier series expansions are again made for the external excitation and the induced currents:

$$E_z^0\left(\vec{r}\right) = \sum_{n_3=0}^{\infty} E_{zn_3}^0\left(\vec{\rho}\right) \cos\left(n_3\frac{\pi z}{h}\right),\tag{20a}$$





$$I_1(z) = \sum_{n_3=0}^{\infty} I_{1n_3}\left(\vec{\rho}\right) \cos\left(n_3 \frac{\pi z}{h}\right)$$
(20b)

and

$$I_2(z) = \sum_{n_3=0}^{\infty} I_{2n_3}\left(\vec{\rho}\right) \cos\left(n_3 \frac{\pi z}{h}\right).$$
(20c)

The known boundary conditions for the total field are valid on the surface of the respective conductors:

$$E_z^{\rm sc}(x_{01} + r_0, y_{01}, z) + E_z^0(x_{01} + r_0, y_{01}, z) = 0$$
(21a)

and

$$E_z^{\rm sc}(x_{02}+r_0,y_{02},z)+E_z^0(x_{02}+r_0,y_{02},z)=0.$$
 (21b)

Substituting now successively the above equations for conductors 1 and 2 in the boundary conditions for the total field, one obtains after a lengthy calculation two equations for the two unknown Fourier expansion coefficients I_{1n_3} and I_{2n_3} :

$$S_{11}I_{1n_3} + S_{12}I_{2n_3} = \frac{-jk}{\eta_0 (k^2 - k_{\nu z}^2)} E^0_{zn_3} \left(\vec{\rho}_{01}\right), \tag{22a}$$

$$S_{21}I_{1n_3} + S_{22}I_{2n_3} = \frac{-jk}{\eta_0 \left(k^2 - k_{\nu z}^2\right)} E^0_{zn_3} \left(\vec{\rho}_{02}\right).$$
(22b)

These equations have to be resolved with respect to I_{1n_3} and I_{2n_3} . The indexed quantities S_{ij} are defined as follows:

$$S_{11} := S\left(k, \vec{\rho}_{01} + \vec{e}_x r_0, \vec{\rho}_{01}, n_3\right), \tag{23a}$$

$$S_{22} := S\left(k, \vec{\rho}_{02} + \vec{e}_x r_0, \vec{\rho}_{02}, n_3\right), \tag{23b}$$

$$S_{12} = S_{21} := S\left(k, \vec{\rho}_{01}, \vec{\rho}_{02}, n_3\right),$$
 (23c)

$$S\left(k,\vec{\rho},\vec{\rho}',n_{3}\right) = \frac{4}{ab} \sum_{n_{1},n_{2}=1}^{\infty} \frac{\sin\left(k_{\nu x}x\right)\sin\left(k_{\nu x}x'\right)\sin\left(k_{\nu y}y\right)\sin\left(k_{\nu y}y'\right)}{k_{\nu}^{2} - k^{2} + j\delta}.$$
 (24)

The function *S* is the two-dimensional scalar Green's function in the rectangular resonator. The function S_{12} describes the coupling between the two lines and contributes to the change of the resonance spectrum of the resonator, compared to that which arises for the same resonator but only equipped with one conductor.

Note that for the calculations in this subsection, only a distributed source was assumed to excite the lines. If in addition lumped sources are also taken into account, the corresponding electrical fields have to be added to the distributed one to give the total exciting field. Of course, one may only consider lumped sources for the excitation. For example, this is the case if one lumped source at the near end of one conductor excites the lines that are otherwise terminated with impedances.

5 Numerical Examples

First, a prominent example [8] is given for a multi-loaded line as shown in Fig. 2. The wire is loaded by a total of five equidistant impedances with the values $Z_1 = Z_2 = Z_3 = Z_4 = 5Z_c$ and terminated at the far end with $Z_5 = Z_c \approx 333 \Omega$. It is fed by an ideal source with $U_0 = 1$ V. Below, results for the currents through the individual impedances are presented, which are obtained by three different methods: first with the aid of the above derived analytical formulas (but here for a lumped source! [5, 7]), second by the numerical package named PROTHEUS (MLFMM method [9]), and third (for comparison) by the classical transmissionline theory for a lossless conductor above perfectly conducting ground. Before the results are discussed, they are displayed in order. In Figs. 4, 5, 6, 7, 8, and 9 the ordinate axes indicate the magnitude of the current at the z-position of the respective load. In this sense |I(z(1), f)| denotes the magnitude of the current spectrum at the z-position of Z_1 .

The chosen conductor configuration is a periodic arrangement. As one knows from the lines in classical transmission-line theory, they act as filters with allowed and forbidden spectral zones. In the considered frequency domain, this is not



Fig. 4 Magnitude of the current spectrum in load Z_1



Fig. 5 Magnitude of the current spectrum in Z_2

the case for a corresponding line in the resonator. Here different interaction mechanisms occur: interaction between the loads and the lumped source via the line, interaction between the loads themselves ("near"-field and "far"-field interaction), and interaction between resonator walls and loads due to reflections from the walls. The decrease of the signal amplitude by almost three orders of magnitude is due to strong reflections between the loads. Based on these kinds of interactions, one can roughly explain the shape of the curves for the current (Figs. 4, 5, 6, 7, and 8).



Fig. 6 Magnitude of the current spectrum in Z_3



Fig. 7 Magnitude of the current spectrum in Z_4

The first important observation throughout all figures is that the analytical and numerical results agree quite well. The TL approximation in Fig. 4 clearly shows the resonance frequencies between two adjacent line impedances of the conductor. But for the current spectra of the line in the resonator, a different picture emerges. Around those just mentioned resonance frequencies, many sharper resonance frequencies of the resonator occur. One recognizes, however, that the bandwidths of the eigenfrequencies of the resonator appear wider than those of a line, which is


Fig. 8 Magnitude of the current spectrum in the terminating matched load Z_5 . Linear scale (transfer coefficient)



Fig. 9 Magnitude of the current spectrum in the terminating matched load Z_5 . Logarithmic scale (transfer coefficient)

terminated with single matched impedance [5, 7]. This can be explained by the interactions between the loads themselves and the loads with resonator walls causing more new eigenfrequencies which overlap with those of the resonator. This superposition of frequencies leads to the apparent broadening of the resonator frequencies. This phenomenon is especially pronounced in Fig. 5, but it can also be seen in Figs 6 and 7. Figures 8 and 9 display the current spectra in the matched load in different scales. As can be seen in the logarithmic representation, the current spectrum in the TL approximation clearly shows the expected zone structure (allowed and forbidden zones), whereas in the resonator this structure is completely absent. Moreover, the amplitudes of both spectra (TL approximation vs. exact analytical formula) differ by more than one order of magnitude.

The second example deals with two parallel conductors inside a rectangular resonator (see Fig. 3). The parameters of the resonator are the same as in Fig. 1. The two conductors are positioned as follows: conductor one has the coordinates $x_{01} = a/2$ and $y_{01} = b/2$ and conductor two the coordinates $x_{02} = a/4$ and $y_{02} = b/2$. Conductor one was fed by an ideal voltage source with $U_0 = 1$ V. The terminations of the conductors have been chosen differently: $Z_{12} = Z_{21} = Z_{22}$ = 50 Ω ; matched; short circuited (0.1 Ω in the numerical calculations). For the matched case the two conductors have different impedance values: $Z_{12} = Z_{C1}$ = 333.74 Ω and $Z_{21} = Z_{22} = Z_{C2} = 298.19 \ \Omega$. Due to the limited space in this chapter, only figures for the case with short-circuited terminations are presented. However, the discussion includes all of the mentioned cases. From the results, one can draw the following conclusions. For all terminations, line resonances are clearly distinguishable from those of the resonator. They are manifolds of $\lambda/2$ resonance frequencies (190 MHz, 380 MHz, etc.). The more the terminations approach the matched values, the smaller and broader the line resonance peaks become. Also the sharp resonance peaks which stem from the resonator decrease with those of the lines. When the lines are short-circuited, their peaks also become narrower and higher. In this case there is maximal reflection at their ends, and therefore more energy remains inside the resonator.

In the subsequent graphs the current magnitudes in the short-circuited terminals are presented in logarithmic scale. Since the density of resonance frequencies increases above 1 GHz, the logarithmic scale is preferred for better distinction between them. Again analytical results obtained with the aid of Eqs. (22a) and (22b) were compared with numerical ones (PROTHEUS), and the agreement is good. The positions of all eigenresonances coincide; however, some of the resonance amplitudes are slightly different. This is due to the narrower discretization of the frequencies in the analytical calculation (1 MHz) with the numerical calculation PROTHEUS (2 MHz) and the terminations with exact 0 Ω instead with 0.1 Ω in the numerical program PROTHEUS (Figs. 10, 11, 12, and 13).

The first eigenresonances of the resonator occur above 600 MHz and their density increases with growing frequencies. For resonances that are removed from cavity resonances, the magnitude of the current in the second (passive) wire is smaller than that of the first one, while for the resonances that are near the cavity resonances, both wires can have the same order of magnitude. In order to study the



Fig. 10 Magnitude of the current at the left terminal of the first wire



Fig. 11 Magnitude of the current at the right terminal of the first wire

influence of the second wire on the first one, the currents at the ends of the first wire have been calculated both in the presence of the second wire and without it. Also a TL approximation [5, 7] was made with only the first wire. The results are depicted in Figs. 14 and 15.



Fig. 12 Magnitude of the current at the left terminal of the second wire



Fig. 13 Magnitude of the current at the right terminal of the second wire

From Figs. 14 and 15, one can clearly see that for resonances outside the area of the cavity resonances, the influence of the second wire on the first is quite small and that the current of the first line is well approximated by a TL approximation. Deviating from this, one can see that in the region of cavity resonances, there are



Fig. 14 Current at the left terminal of the first wire, with and without the second one



Fig. 15 Current at the right terminal of the first wire with and without the second one

clear differences: The resonance frequencies in the current spectrum of the line which are caused by the interaction with the cavity are shifted to higher frequencies if the second conductor is present, a phenomenon which was discovered in [7] for a similar configuration. With the addition of further closely spaced, nonparallely conducted wires, one would expect a much richer current spectrum on the wires, both in the area of line resonances (e.g., resonance splitting [10, 11]) and in the area of resonances.

6 Conclusion

In this chapter transmission lines inside a rectangular resonator were analyzed. This was done for a multi-loaded wire and for two interacting wires. The analytical results obtained for the current spectra (at the impedances of the conductors) were compared with numerical ones, and good agreement was found. In the first case it was observed that the zone structure (allowed and forbidden zones) was absent in the spectra in the domain of resonator eigenresonances, whereas below this frequency region, such a structure appears. In the case of two wires in the resonator, one could observe a shifting of resonator eigenfrequencies to higher values. Thus, in future studies one can expect to find that the inclusion of multiple wires (in particular nonuniform multiple cables) will increase the eigenfrequency density in the higher frequency domain. Likewise, it is assumed that in the case of nonlinear terminations of the conductors, resonator eigenfrequencies in the lower frequency domain will occur. The investigations of these assumptions will become the subject of forthcoming studies.

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Analysis of Open TEM-Waveguide Structures

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Abstract This work belongs to a research project on the analysis and characterization of higher order modes occurring in open TEM-waveguide structures. An open TEM waveguide, derived from a conventional GTEM cell by removing the sidewalls, is investigated. The intrinsic resonances of the electromagnetic field occurring in the test volume of the waveguide are analyzed in frequency domain by computer simulation and measurement. This resonance behavior is compared to that of more simplified wire models, describing just the planar septum of the original TEM waveguide. The influence of the number of wires used in the wire model is investigated with respect to the resonant behavior. The use of wire structures is a prerequisite for application of transmission-line super theory (TLST) for further analysis.

Keywords TEM waveguide • Transmission-line super theory • MoM • TEM mode • Higher order modes

1 Introduction

TEM waveguides are widely used in EMC testing as a convenient alternative to Open Area Test Sites (OATS), Fully Anechoic Rooms (FAR), and Semi-Anechoic Chambers (SAC). Because of the generation of higher order TE and TM modes, closed TEM waveguides are generally restricted in frequency range. Therefore in

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qualification testing of equipment in fast transient environments, as, for example, NEMP or UWB environments, open TEM-waveguide structures are widely used. While higher order modes in TEM cells were sufficiently described in an analytic form in the past [1], there is currently no satisfying understanding and (semi-) analytic description of the resonance effects in open TEM waveguides.

2 Resonances in the Closed GTEM Cell

The used closed GTEM cell (EMCO 5320) is about 1.3 m long and has a width of about 0.65 m at the far end. To investigate the resonant behavior of the cell, the absorber material, usually used for suppressing higher order modes, was removed. Also the septum was short circuited at the far end. These measures lead to an artificial enhancement of the Q-factor of the cell which is usually unwanted for the use of TEM waveguides in EMC testing. Measurements of the magnetic field components were done with a half-loop ground sensor at a reference point on the conducting ground plane in the test volume of the cell. The reference point was chosen away from any symmetry lines. Details can be seen in Fig. 1. The position of the used *H*-field sensor is depicted as a small circle and named "H-Sensor."

TEM-wave propagation implies that there is no *E*- or *H*-field component in propagation direction. In the used TEM-waveguide geometry, the propagation direction changes with the height in the test volume due to the spherical nature of the propagating wave between ground plane and septum. Only in the vicinity of the ground plane, the propagation direction is identical to the *z*-axis of the used Cartesian coordinate system. Therefore, the measurement of the *z*-component of the magnetic field strength, H_z , at the ground plane of the GTEM cell is a sensitive indicator for resonances showing the deviation from TEM-mode propagation. These resonances can be seen in Fig. 2 where the measured peaks are compared with analytic results (colored vertical lines) using the Wentzel–Kramers–Brillouin (WKB) method described in [1, 2]. This is a well-known method in theoretical physics and in this case based only on the (theoretical) length of the cell, the gradient of the horizontal ground plane, and the cutoff frequencies of the higher order TM and TE modes.



Fig. 1 Construction data of the EMCO 5320 GTEM cell used for closed and open versions of a TEM waveguide



Fig. 2 Resonances of H_z in the closed GTEM cell for the reference point near the ground plane

3 Turning the GTEM Cell into an Open TEM Waveguide

The closed GTEM cell was turned into an open TEM waveguide by mechanically detaching the sidewalls. Except of the sidewalls the GTEM cell and the so made open TEM waveguide have exact the same geometry. Therefore, the difference in the resonant behavior can directly be contributed to the change from a closed to an open system. A TDR measurement showed that there is only a minor enhancement of the characteristic impedance by 4 Ω . The original EMCO 5320 GTEM cell has a characteristic impedance of 50 Ω . A picture of the open TEM waveguide is shown in Fig. 3.

As expected the resonances of the z-component of the magnetic field strength, H_z , are smaller by orders of magnitude compared to the closed version [2], as can be seen in the left graph of Fig. 4. Of course this is contributed essentially to the fact that the Q-factor declines significantly by removing the sidewalls of a conducting closed structure. But as we zoom in (see Fig. 4, right graph), we can still recognize a resonant structure in H_z which is, at least above 600 MHz, clearly above noise level and fits quite well to measurement.

Both the closed and the open TEM waveguide were modeled using the Method of Moments (MoM)/Multilayer Fast Multipole Method (MLFMM) code PROTHEUS [3], developed by EADS/Cassidian in cooperation with German Armed Forces (Bundeswehr). The left graph of Fig. 4 shows the PROTHEUS simulation of H_z for the closed and open model. It is interesting to compare the simulation result of the closed GTEM cell to the corresponding measurement



Fig. 3 Open TEM waveguide (EMCO 5320 GTEM cell without sidewalls)



Fig. 4 Left: H_z resonances of the open TEM waveguide compared to the closed GTEM cell (MLFMM computer simulation). Right: Measured H_z resonances of the open TEM waveguide compared to MLFMM computer simulation

displayed in Fig. 2. One recognizes that the measured resonances of the H_{10} mode cannot be seen in the simulation. The reason is that the simulation model is, in contrast to the real cell, perfectly symmetric and therefore the H_{10} mode is not excited in the simulation but in the real cell, where absolutely perfect symmetry can technically not be achieved.

4 Simplification by Using Wire Models

4.1 Reason for Simplification

A plan is already in place to investigate open TEM-waveguide structures using the transmission-line super theory (TLST) [4]. TLST is a full wave description of the Maxwell's equations cast into the form of the classical telegrapher equations.



Fig. 5 Simplification of the open TEM waveguide (*top left*) by substituting the septum by wire structures and omitting the top plate

Therefore, higher order modes and radiation effects are intrinsically included in the TLST results. At least at high frequencies, radiation effects play a significant role concerning open TEM waveguides. For practical reasons, concerning the numerical effort, for TLST, it is necessary to use the "thin-wire approximation" [5]. Therefore, instead of planar conductors just wires have to be used. Wires instead of the original planar conductors would be possible only if the characteristic behavior of the electromagnetic field in the test volume did not change significantly. To show this fact is the goal of the presented remaining work.

4.2 Defining Different Wire Models

Inspired by the geometry of several wire-based NEMP simulators [6], the idea was just to model the septum of the TEM waveguide by wires and omit the top plate totally. This implies to transfer the problem of an open GTEM cell to the problem of a conical transmission line. For the electromagnetic fields of the TEM mode of conical transmission lines, closed form expressions are known since the 1970s [7, 8]. The most complex wire model used in this work represents the septum by nine wires, hence called the 9-wire model. A further simplification represents the septum by three wires (3-wire model), and the most extensive simplification of the TEM waveguide is the 1-wire model, which is just the projection of the septum on the vertical symmetry plane. Figure 5 shows the model reduction from the original TEM waveguide to the different mentioned wire models. The graphical



Fig. 6 Transversal magnetic field strength on the *xy*-plane through the reference point in the test volume for the planar TEM-waveguide structure and the wire models

presentation of the models was done with the Concept-II software [9] developed at TU Hamburg-Harburg.

4.3 Comparing the Transversal Magnetic Fields

For further investigations the x-component of the magnetic field strength, H_x , was chosen. For a pure TEM wave propagating in z-direction, E_y and H_x are the only components of the electromagnetic field. Therefore, these field components lie on transversal planes of the propagation direction. Figure 6 shows the simulated (Concept-II) magnetic field strength for a frequency of 1 GHz on the transversal xy-plane through the reference point for the planar TEM waveguide and the wire models. The fact that the traveling wave has a spherical wave front is neglected here. The colors in the different plots represent the same range of magnetic field strength values. It can be seen that the magnetic field of the planar TEM waveguide and the 9-wire model are almost the same, at least for the region of the test volume below the septum. Also for the 3-wire model, the deviation in the test volume is not



Fig. 7 Frequency dependency of the transversal magnetic field component, H_x , for the planar TEM waveguide and the wire models

significant. Of course, for the 1-wire model, which is the most extensive simplification, there is a clear reduction of homogeneity in the potential test volume.

4.4 Frequency Dependency of the Transversal Magnetic Field

The frequency dependency of the transversal component of the magnetic field strength, H_x , was simulated for the different models for the frequency range of 100 MHz–1 GHz in frequency steps of 1 MHz. The simulations were performed with the PROTHEUS software and the results are shown in Fig. 7.

In the regarded frequency range, the transversal magnetic field of the open TEM waveguide shows a characteristic course. There are distinct minima at the frequencies of about 175, 550, and 910 MHz. Between every two of these minima, three resonance peaks appear. The spacing of these peaks ($\lambda/2$) corresponds to an average value of the total length of the conducting structure. The spacing of the distinct minima, however, corresponds to the electrical length of the path on the back plane from the septum to the ground plane. This characteristic structure of H_x remains for the simplifications represented by the wire models even though the frequencies of the peaks and minima are slightly shifted. It can be observed that the resonances between the minima are more distinct, the fewer wires are used in the model.



Fig. 8 Measurement setup of the 3-wire model

It could be shown that additional wires representing the top plate of the open TEM waveguide do not improve the convergence of the resonances. Additional Concept-II simulations of the open TEM waveguide with the planar septum showed that there are no significant cross currents on the septum. This simulation also showed that due to the skin effect, the current density in the left and right outer region of the planar septum is higher than in the inner region.

4.5 Comparison to Measurement

The wire models were also built as an experimental setup shown in Fig. 8 for the 3-wire model. Silver wires with a diameter of 0.5 mm were used. They were mounted on a 2 m \times 1 m aluminum ground plane, and a magnetic field ground sensor (half loop) was integrated in the ground plane at the reference position. For the measurement, a vector network analyzer (VNA) was used. The VNA output was connected directly to the wire structure, the magnetic sensor to the VNA input.

Figure 9 shows the results of H_x for the PROTHEUS simulation, the measurement, and TLST numerical analysis for the 1- and the 3-wire model. The correspondence between the three methods is rather good. The numerical implementation of TLST for the one-wire case is thoroughly described in [10].

5 Conclusion

Having a direct comparison between a closed and an open, but otherwise identical TEM waveguide using the EMCO 5320 GTEM cell, the difference in the resonant behavior of the transversal magnetic field could be directly investigated. As



Fig. 9 Comparison of the transversal magnetic field component, H_x , for simulation, measurement, and TLST analysis for the 1-wire model (*left*) and the 3-wire model (*right*)

expected, the resonance effects in the open TEM waveguide are smaller by order of magnitudes, but there is still a characteristic resonance structure identifiable. Compared to the closed cell, there are fewer but broader resonance peaks resulting not at least from the much lower Q-value of the open structure.

It could be also shown that the planar open TEM waveguide can be simplified by using wire structures just for the septum and the back plane. This implies that the characteristic electromagnetic behavior of an open GTEM cell could be transferred to that of a conical transmission line. The characteristics of the resonances are not changed significantly by this simplification. The main characteristics of the resonances are distinct minima due to standing waves on the back plane between the end of the septum and the ground plane. Between each two of these minima, there are three (planar TEM waveguide, 9-wire model) or four (3- and 1-wire model) peaks observable. For the 1-wire model, these peaks are most significant and their spacing corresponds to the overall length of the wire. The more wires are used, the more the peaks vanish. This behavior is contributed to the coupling between the different wires and a kind of averaging effect because the wires have different lengths on the septum.

Besides the numerical simulations, the wire models were also built as an experimental setup and measurements of the transversal magnetic field were performed. It could be shown that the correspondence between numerical simulation and measurement is rather good. These results show that it is appropriate to do further investigations with wire models. The basic results can then be transferred to planar TEM-waveguide structures. First results with the TLST are very satisfying as the good correspondence of the transversal magnetic field between TLST analysis and classical numerical results and also measurement shows.

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Antenna Radiation in the Presence of an Infinite Interface

Y. Béniguel

Abstract This chapter addresses the problem of antenna analysis for radiating structures located at the vicinity of one or several infinite interfaces between different media. In the first stage, it is shown how the field integral equations shall be modified in order to take the interface into account and how the Sommerfeld integral calculation can be implemented in an efficient way. The propagation problem of the radiated field over an irregular interface is then addressed. The results obtained are presented for a typical bi-conical HF antenna.

Keywords Mixed potential integral equation • Sommerfeld integrals • Parabolic equation

1 Introduction

The electromagnetic field integral equations technique is one of the most appropriate for antenna analysis. It allows considering surfaces and wire elements, including dielectric parts as lenses or substrates. Using the classical method of moments, the analysis technique only considers interfaces between media. These interfaces are meshed into surface or wire elements which constitute the problem unknowns. The case of infinite interfaces can also be considered modifying the Green's problem function in order to avoid the meshing of these interfaces. However, it includes in this case the calculation of the Sommerfeld integrals.

The VLF and HF antennas are examples of such antennas. They are placed on the ground. They usually include a ground plane constituted by radial wires placed on the interface. The electrical ground characteristics strongly influence the antenna radiating pattern and the VSWR. Another important example is the case of the

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microstrip antennas. Analyzing such antennas using the method of moments and solving the field integral equations implies considering both electric and magnetic currents on all interfaces. On such antennas the current is located close to the microstrip elements. It is consequently possible to consider the substrate interfaces as of infinite extent.

This chapter addresses the problem of infinite interfaces and presents a recent technique for an efficient calculation of the Sommerfeld integrals and its implementation in a numerical code (http://www.ieea.fr/en/softwares/icare-mom.html). The propagation over an irregular terrain is evaluated subsequently, solving the parabolic equation. The near field radiated by the antenna which includes both a sky wave and a ground wave contribution is used as an input to this problem.

2 The Mixed Potential Integral Equation

It has been shown [1] that there are three different ways to write the Lorentz gauge relationship in order to meet the boundary conditions on the interface. Although equivalent from a theoretical point of view, formulation C is the most convenient as regards its numerical implementation. This formulation has been used. With respect to the Electric Field Integral Equation (EFIE), the new equation, named Mixed Potential Integral Equation (MPIE), includes additional terms to take the Green's function modification into account.

In the case of a metallic structure, the MPIE equation is

$$E^{\rm s} = -j\omega A - \nabla \Phi \tag{1}$$

with for the vector potential

$$A = \int \overline{G}^{A}(r, r') J_{s}(r') ds'.$$
⁽²⁾

In order to meet the boundary conditions on the interface, the Green function shall be a tensor and can be written as

$$G^{A} = G^{A}_{vv} + \widehat{z} \, \widehat{u} \, G^{A}_{zu} + \widehat{z} \, \widehat{z} \, G^{A}_{zz}, \tag{3}$$

where $\widehat{z} \ \widehat{u}$ means the vertical (z) contribution of a horizontal [u(x or y)] dipole.

The scalar potential is correspondingly

$$\boldsymbol{\Phi} = \frac{j\omega}{k_i^2} \int \left[\nabla \cdot \overline{G}^A(\boldsymbol{r}, \boldsymbol{r}') \right] \cdot J_{s}(\boldsymbol{r}') \mathrm{d}\boldsymbol{s}' \tag{4}$$

and the Lorentz gauge relationship becomes in that case

Antenna Radiation in the Presence of an Infinite Interface

$$\frac{j\omega}{k_{i}^{2}}\nabla\cdot\overline{G}^{A}\left(r,r^{'}\right) = -\nabla^{'}K_{\varPhi}\left(r,r^{'}\right)$$
(5)

The resulting modified scalar potential equation is finally

$$\boldsymbol{\Phi} = \int_{S} K_{\boldsymbol{\Phi}}(\boldsymbol{r}, \boldsymbol{r}') \nabla' \cdot J_{s}(\boldsymbol{r}') \mathrm{d}\boldsymbol{s}' - \int_{C} K_{\boldsymbol{\Phi}}(\boldsymbol{r}, \boldsymbol{r}') J_{s}(\boldsymbol{r}') \cdot \widehat{\boldsymbol{n}} \, \mathrm{d}\boldsymbol{\ell}'.$$
(6)

This calculation is greatly simplified in the case where the structure does not have vertical elements. There is no difficulty to consider additional terms for magnetic currents in case of dielectric surfaces. They have not been included here for the sake of clarity.

3 Dipoles Interaction in the Presence of an Infinite Interface

The calculation of the different terms introduced in the MPIE formulation is greatly facilitated when performed in the spectral domain. In the space domain, the cylindrical coordinates are used with $\overline{\rho} = x \,\widehat{x} + y \,\widehat{y}$. The corresponding variable in the spectral domain is $\overline{k}_{\rho} = k_x \,\widehat{x} + k_y \,\widehat{y}$. The field propagation is calculated using the analogy of this problem to the well-known transmission line formalism using voltage and current sources.

The equivalent transmission line equations can be written as

$$\frac{\mathrm{d}V^p}{\mathrm{d}z} = -jk_{zi}Z_i^p I^p + v^p \quad \text{and} \quad \frac{\mathrm{d}I^p}{\mathrm{d}\,z} = -\frac{jk_{zi}}{Z_i^p}V^p + i^p,\tag{7}$$

where v^p and i^p are the source terms which can be related to the incident field.

In this expression, the superscript p stands for e in case of TM fields and for h in case of TE fields. Z is the characteristic impedance, and the longitudinal propagation constant is such that

$$k_{zi} = \sqrt{k_i^2 - k_\rho^2},\tag{8}$$

where k_i is the wave number inside the medium numbered *i*.

In the case of two media separated by one infinite interface, the reflection coefficients for TE and TM polarized fields are given by

$$R^{\mathrm{TE}} = \frac{k_{z1} - k_z}{k_{z1} + k_z}$$

and

$$R^{\rm TM} = \frac{n^2 k_z - k_{z1}}{n^2 k_z + k_{z1}}.$$
(9)

The problem can easily be extended to an arbitrary number of layers [2]. The reflection coefficients become in that case a combination of the reflection coefficients at each interface, involving the thickness of each layer. The numerical implementation is straightforward.

Solutions of the transmission line equations have been derived by Michalski [1, 3], as the set of following equations:

$$j\omega\mu_0 G_{zz} = \eta_0^2 I_v^e, \quad j\omega\mu_0 G_{vv} = V_i^h, \quad G_{zu} = \frac{1}{jk_\rho} \left(I_i^h - I_i^e \right),$$
(10)

and

$$K_{\boldsymbol{\phi}} = \frac{j\omega\varepsilon_0}{k_{\rho}^2} \left(V_i^e - V_i^h \right). \tag{11}$$

4 Calculation of the Sommerfeld Integrals

The MPIE integral terms can be written in the general form as

$$F(\rho) = \frac{1}{2\pi} \int f(k_{\rho}) J_n(k_{\rho}\rho) k_{\rho} \mathrm{d}k_{\rho}$$
(12)

with n equal 0 or 1. A special case of these integrals corresponds to the following, known as the Weyl identity:

$$G = \frac{\exp(-jkR)}{R} = \int_0^\infty \frac{\exp(-jk_z|z|)}{jk_z} J_0(k_\rho \rho) k_\rho \mathrm{d}k_\rho.$$
(13)

In the classical derivation of the problem as introduced by Sommerfeld [4], double derivatives with respect to the radial and vertical distances are introduced. This creates a difficulty when trying to implement this in a classical EFIE code with rooftop basis functions (both on wires and on surfaces). In the spectral domain, this corresponds to an increase in the Bessel function order and can be addressed more easily.

Another benefit to perform the calculation in the spectral domain is the fact that it allows isolating the different contributions. They are then subtracting from the integrand in order to get regular functions more suited to a numerical evaluation.

62

The quasi-dynamic contribution corresponds to the value of $f(k_{\rho})$ when $k_{\rho} \rightarrow \infty$. In particular the asymptotic values of the TE and TM reflection coefficients values which appear in $G_{\nu\nu}$ and G_{zz} are

$$R^{\rm TE} = \frac{k_{z1} - k_z}{k_{z1} + k_z} = 0$$

and

$$R^{\rm TM} = \frac{n^2 k_z - k_{z1}}{n^2 k_z + k_{z1}} = \frac{n^2 - 1}{n^2 + 1} = q \tag{14}$$

and combinations of these two for the other terms.

The poles contributions are extracted by pairs [5, 6]. In the case of two layers, this only applies to the terms involving the R^{TM} coefficient. Each MPIE integral term is finally decomposed into three parts, an asymptotic value, a pole contribution, and a regular part:

$$f_{1}(k_{\rho}) = f(k_{\rho}) - \lim_{k_{\rho} \to \infty} f(k_{\rho}) - \frac{2k_{\rho\rho}R(\rho_{\rho})}{k_{\rho}^{2} - k_{\rho\rho}^{2}}$$
(15)

with $R(\rho_p)$ the residue of function $f(k_p)$ at the pole value in the k_p complex plane.

The remaining function to be integrated after subtraction of the quasi-dynamic part and of the residues term is regular and well suited for numerical evaluation. The contribution of the quasi-dynamic contribution (QD) and of the poles is reintroduced subsequently as an analytical expression:

$$F(\rho) = \frac{1}{2\pi} \int f_1(k_\rho) J_0(k_\rho \rho) k_\rho dk_\rho + \text{QD } G_0 + 2j \sum \text{Residues}$$
(16)

with G_0 the free space contribution.

Two techniques may be used for numerical evaluation of the remaining integrals, either the phase stationary technique together with the steepest descent contour or the complex image technique. These two are of interest, and they allow at the minimum to cross-check the results obtained. For the first technique the steepest descent contour is different for each MPIE integral term. It should consider the occurrence of the poles and of the branch points.

The second technique known as complex image technique became more popular in the recent years. It consists essentially in a change of the integration plane, the k_z complex plane replaces the k_ρ complex plane, and in the choice of an integration contour which ensures very fast convergence and applies to all MPIE integral terms.

The integration contour in the k_z plane is a straight line which exhibits consequently a linear relationship between the integration variable and the longitudinal wave number component. The different integrals of the remaining regular terms decrease very rapidly along this contour. They are approximated by a sum of exponential terms using the matrix pencil approximation [7]. Each one of these integrals is written as

$$f_1(k_{\rho}) = \sum_{1}^{N} a_i \exp(-b_i k_z),$$
(17)

where N is a small number, usually below 5.

The use of the Weyl identity is made possible provided that new distances be considered, defined as

$$R_i^2 = \left(\rho^2 + \left(z + z' - jb_i\right)^2\right)^2.$$
 (18)

As b_i is a complex number, the result may be seen, using the Weyl identity, as a sum of complex images.

This way to proceed is illustrated on Fig. 1 below. The variation of the Green function terms along the contour shows a very rapid decrease. The respective contributions of the three terms depending on the radial distance between one source point and one observation point are shown on the right panel. The weight of the different contributions strongly depends on the ground parameters.

The extraction of the quasi-dynamic contribution and that of the poles contributions are still amenable in the case of an arbitrary number of layers with dipoles (either surface elements or wires) located in the different media. The different integral terms involve a combination of the different wave numbers in the different media when considering the field radiated by a current element located in medium 1 on a current element located in medium 2.

5 **Propagation over an Irregular Terrain**

The propagation over an irregular terrain is calculated using the parabolic equation:

$$\frac{\partial^2 u}{\partial z^2} + 2jk\frac{\partial u}{\partial x} + k^2(n^2 - 1)u = 0$$
⁽¹⁹⁾

with x the coordinate along the terrain profile and z the coordinate along the vertical to the mean tangent plane.



Fig. 1 The regular terms behavior along the contour in the *kz* complex plane (*left panel*) and their respective contribution to the Sommerfeld integrals (*right panel*)

For HF propagation the atmospheric index contribution is not significant. Knowing the field on a vertical line at one x value along the propagation axis, the corresponding values at the following x value are obtained using the following equation:

$$u(x,z) = \exp\left(j\frac{k}{2}(n^2 - 1)\Delta x\right) \text{FFT}^{-1}\left\{\exp\left(-j\frac{k_z^2\Delta x}{2k}\right) \text{FFT}[u(x_0,z)]\right\}$$
(20)

Inputs of this multiple split step technique are the field values at the antenna location. These are the outputs of the antenna problem. They include the two contributions: the ground wave and the sky wave. In addition a boundary condition is applied on the ground (Leontovich condition), and an absorbing condition is applied at the upper part of the calculation domain.

The calculation is performed on a 2D window. It takes a few tenths of seconds for distances up to 100 km from the antenna location and allows considering any field polarization and terrain profile. Variable electrical characteristics can be affected to the terrain along the propagation direction.

6 Results Obtained

6.1 The Antenna Problem

The case considered is a typical bi-conical antenna of 7 m high placed on the ground (dielectric constant 15 and conductivity 0.01 S/m). The calculation provides the VSWR, the currents, and the pattern (near field, far field, and ground wave).

The ground plane plays a major role on the antenna matching, allowing to decrease the VSWR to acceptable values. The sky wave far-field pattern is zero in the ground plane which is usually of a small extent as compared to the wave-length. The near-field pattern includes the contribution of the ground wave which exhibits a peak value in the interface plane (Figs. 2 and 3).

6.2 Ground Wave Propagated Field

Figure 4 shows the incident field on a vertical to the ground plane at a few hundred meters from the antenna location. The ground wave is the major contribution, but the sky wave amplitude is also significant. This field profile constitutes the input



Fig. 2 The current distribution on a bi-conical antenna (a) and its input impedance and VSWR over the HF bandwidth (b)

data for the parabolic equation. The right panel figure shows the field propagation losses in a vertical domain 90 km away from the antenna and 400 m in height consequently at grazing angles.

7 Conclusion

The analysis of antennas located at the vicinity of an infinite interface has been presented. It has been shown how the field integral equations shall be modified in order to take infinite interfaces into account. The technique has been presented for two media but can easily be extended to an arbitrary number of layers. Such a multiple layers structure occurs in particular in the case of microstrip antennas.



Fig. 3 Antenna far field and near field

As derived by Michalski [1], the Green's problem function is a tensor. Each term of this tensor includes a combination of the TE and TM reflection coefficients. There is some benefit to split these integrals into three separate contributions corresponding to the poles contribution, to a "quasi-dynamic" contribution, and to a regular term which shall be calculated numerically. Two techniques may be used to evaluate the regular term: the steepest descent contour technique and the complex image technique. This last offers a number of advantages as regards, in particular, its implementation. The simultaneous development of these two techniques allows to check the accuracy of the results.

The developments presented have been verified by checking the results with respect to measurements on existing antennas [8]. The case of a bi-conical HF antenna has been shown. Although not presented here, the validations made on the VSWR have shown an excellent agreement. The ground wave propagation is coupled to the antenna problem. This allows an accurate evaluation of the radiated near field (sky wave and ground wave) used as an input for the propagation problem.



Fig. 4 Field profile on a vertical close to antenna location (*left panel*) and propagation losses (*right panel*)

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A Standard for Characterizing Antenna Performance in the Time Domain

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Abstract We derive here a simple function describing antenna performance in the time domain. This function describes antenna performance in both transmission and reception and in both the time and frequency domains. The resulting equations are as simple as possible. From this function one can simply derive such conventional frequency domain quantities as gain, realized gain, and antenna factor. It is hoped that this function will be adopted as an IEEE standard for time domain antenna performance.

Keywords Time domain antenna response • Gain • Antenna impulse response

1 Introduction

There are already standards for characterizing antenna performance in the frequency domain, but no such standard exists in the time domain. This becomes a challenge if, for example, one wishes to buy or sell an antenna with a specified performance in the time domain. In the frequency domain, one normally uses antenna gain [1], but this offers little information about the antenna's time domain performance.

A number of earlier papers have addressed this issue [2-5]. However, there remains no standard method of describing antenna performance in the time domain. We demonstrate here a method of simplifying the equations as much as possible, leading to a standard waveform describing antenna performance. This chapter is a condensed form of the work appearing in [6, 7].

In order to characterize an antenna in the time domain, a function should have five characteristics:

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Fig. 1 Characterizing an impulse radiating antenna with four different waveforms for receive (*top*) and transmit (*bottom*) and at two different pulse widths (*left* and *right*). Note that t_{FWHM} is the full-width half max of the incident field and t_{rise} is the rise time of a source voltage

- 1. The function should fully describe antenna performance with equations that are as simple as possible.
- 2. The function should describe antenna performance in both transmission and reception.
- 3. The function should describe antenna performance in both the frequency and time domains.
- 4. The function should be simply related to frequency domain standards, such as gain, realized gain, and antenna factor.
- 5. The function should be simply related to quantities that are measurable in the laboratory, typically with an oscilloscope.

By characterizing an antenna's performance with its impulse response, we simplify our understanding of antenna physics—especially in the time domain. For example, consider the time domain response of an impulse radiating antenna (IRA), as sketched in Fig. 1. It is common to characterize its performance differently in transmission and reception and for different rise times or pulse widths. On the top is the received voltage when the IRA is excited by an impulse-like electric field, with two different pulse widths. On the bottom is the radiated field when the IRA is driven by an impulse-like voltage, with two different rise times. The bottom waveforms are proportional to the derivatives of the corresponding top waveforms. In this formulation, four waveforms are required to fully describe antenna performance. However, a single waveform, $h_N(t)$, can be formulated to contain all the information in these four waveforms. That is the purpose of this paper.

2 The Proposed Function: $h_{\rm N}(t)$

Consider the configuration shown in Fig. 2. A single antenna is used in either transmit or receive configurations. In transmission, it is driven by a voltage source with a 50 Ω load. In reception, the antenna has a 50 Ω load.

We have found that the antenna equations exhibit a striking simplicity and symmetry if they are expressed not in terms of electric fields and voltages but in terms of the square root of power or power density. Thus, instead of voltages, we use voltages divided by the square root of the load or source impedance, and instead of electric fields, we use electric fields divided by the square root of the intrinsic impedance of free space. In this format, we have the following equations for transmission and reception on boresight, with dominant polarization:

$$\frac{E_{\rm rad}(t)}{\sqrt{377\,\Omega}} = \frac{1}{2\pi cr} h_{\rm N}(t) \circ \frac{dV_{\rm src}(t')/dt}{\sqrt{50\,\Omega}},$$

$$\frac{V_{\rm rec}(t)}{\sqrt{50\,\Omega}} = h_{\rm N}(t) \circ \frac{E_{\rm inc}(t)}{\sqrt{377\,\Omega}},$$

$$V_{\rm s}(t) = 2V_{\rm src}(t),$$

$$t' = t - r/c,$$
(1)

where $V_{\rm rec}(t)$ is the received voltage into a 50 Ω load or oscilloscope and $V_{\rm src}(t')$ is the source voltage in retarded time as measured into a 50 Ω load or oscilloscope. Furthermore, $E_{\rm inc}(t)$ is the incident electric field, $E_{\rm rad}(t)$ is the radiated electric field, r is the distance away from the antenna, c is the speed of light in free space, and " \circ " is the convolution operator. Note also that $h_{\rm N}(t)$ has units of meters per second in the time domain and meters in the frequency domain.

The above expressions have been simplified in three ways. They refer only to dominant polarization, they are valid only on boresight, and attenuation from source to receiver has been ignored. All three effects can be easily restored. However, these effects do not affect the derivation, and we find it easier to manipulate simpler versions of the equations.



Fig. 2 An antenna shown in transmit and receive configurations

If one has the response of the antenna in a 50 Ω system, this is sufficient to find the response in a system with arbitrary source or load impedance. We concentrate on the 50 Ω system because that is the most common impedance for test equipment. We note further that common definition of S_{11} includes an assumption of 50 Ω feed at the input port, so our assumption here is consistent with that idea.

The above equations may be combined into a single equation that incorporates both transmit and receive antennas, as seen on an antenna range. In this case, the equation is

$$\frac{V_{\rm rec}(t)}{\sqrt{50\Omega}} = \frac{1}{2\pi cr} h_{\rm N,RX}(t) \circ h_{\rm N,TX}(t) \circ \frac{\mathrm{d}V_{\rm src}(t')/\mathrm{d}t}{\sqrt{50\Omega}}.$$
(2)

. ..

This is the equation that we actually use on our time domain antenna range. A natural way to calibrate the antenna range is to use identical transmitting and receive antennas. In this case, the only unknown is $h_N(t)$, which is found by deconvolution. Once $h_N(t)$ is found, one can substitute the device under test for one of the identical antennas. Note also that $h_N(t)$ may be generalized to both polarizations and to arbitrary angles with the more general expression, $\vec{h}_N(\theta, \phi, t)$.

3 Derivation of $h_{\rm N}(t)$ Using Reciprocity

The proof of Eq. (1) requires more space than what is available here. However, we can provide an overview of the proof and a list of its assumptions. The complete proof is available in [6].

We describe the antenna response in terms of open circuit, short circuit, and 50 Ω load conditions for both transmission and reception. These six cases give us six functions describing antenna performance. The three transmitting functions are related to each other by the source impedance and the input impedance of the antenna. Similarly, the three receive functions are related to each other by the load impedance and antenna input impedance.

The challenge lies in relating the transmit antenna functions to the receive antenna functions. To do so, we treat a two-antenna system as a simple two-port circuit, as shown in Fig. 3. We then assume that one of the antennas is a small electric dipole, whose transmitting and receiving characteristics are already well known. We then apply an open-circuit voltage first to Port 1, and then to Port 2, and calculate the received short-circuit currents. Since reciprocity holds, the two currents must be equal, and this provides enough information to relate the transmitting and receiving antenna characteristics to each other, thereby proving Eq. (1).



Fig. 3 A two-antenna system shown as a circuit

4 Other Equations

Once we have the $h_N(t)$ for an antenna, it is straightforward to derive its gain, G; realized gain, G_r ; and antenna factor, AF. These are expressed as

$$\widetilde{G}_{r}(\omega) = \frac{4\pi}{\lambda^{2}} |\widetilde{h}_{N}(\omega)|^{2},$$

$$\widetilde{G}(\omega) = \widetilde{G}_{r}(\omega) / \left[1 - \left| \widetilde{S}_{11}(\omega) \right|^{2} \right],$$

$$AF = \frac{\widetilde{E}_{inc}(\omega)}{\widetilde{V}_{rec}(\omega)} = \sqrt{\frac{377}{50}} \frac{1}{\left| \widetilde{h}_{N}(\omega) \right|} = \frac{9.73}{\lambda \sqrt{G_{r}}},$$
(3)

where S_{11} is the reflection coefficient looking into the antenna from a 50 Ω line.

We can now generalize the antenna equations by adding back in a dependence on angle, polarization, and attenuation. Thus, we have

$$\frac{\dot{E}_{\rm rad}(\theta,\phi,t)}{\sqrt{377\Omega}} = \frac{{\rm e}^{-\alpha r}}{2\pi rc} \vec{h}_{\rm N}(\theta,\phi,t) \circ \frac{{\rm d}V_{\rm src}(t')/{\rm d}t}{\sqrt{50\Omega}}, \quad t'=t-r/c,$$

$$\frac{V_{\rm rec}(t)}{\sqrt{50\Omega}} = \vec{h}_{\rm N}(\theta,\phi,t) \circ \frac{\vec{E}_{\rm inc}(\theta,\phi,t)}{\sqrt{377\Omega}}.$$
(4)

We can express the received voltage and radiated field for those cases in which the source or load impedance is not 50 Ω . The received voltage is [7]

$$\frac{\widetilde{V}_{\text{rec}}}{\sqrt{Z_{\text{L}}}} = \sqrt{\frac{\widetilde{Z}_{\text{L}}}{50\Omega}} \frac{\widetilde{Z}_{\text{in}} + 50\Omega}{\widetilde{Z}_{\text{in}} + \widetilde{Z}_{\text{L}}} \widetilde{h}_{\text{N}} \frac{\widetilde{E}_{\text{inc}}}{\sqrt{377\Omega}},$$
(5)

where Z_L is the load impedance. For the special cases of received voltage into an open circuit or short circuit, this simplifies to

$$\widetilde{V}_{oc} = \widetilde{h}_{V}\widetilde{E}_{inc}, \quad \widetilde{h}_{V} = \frac{\widetilde{Z}_{in} + 50\Omega}{\sqrt{50\Omega \times 377\Omega}}\widetilde{h}_{N},$$
$$\widetilde{I}_{sc} = \widetilde{h}_{I}\widetilde{E}_{inc}, \quad \widetilde{h}_{I} = \frac{1}{\widetilde{Z}_{in}}\frac{\widetilde{Z}_{in} + 50\Omega}{\sqrt{50\Omega \times 377\Omega}}\widetilde{h}_{N}.$$
(6)

The general equation for radiation from a source with arbitrary impedance is

$$\frac{\widetilde{E}_{\rm rad}}{\sqrt{377\,\Omega}} = \frac{j\omega}{4\pi c} \frac{{\rm e}^{-jkr}}{r} \sqrt{\frac{\widetilde{Z}_{\rm S}}{50\,\Omega}} \frac{\widetilde{Z}_{\rm in} + 50\,\Omega}{\widetilde{Z}_{\rm in} + \widetilde{Z}_{\rm S}} \widetilde{h}_{\rm N} \frac{\widetilde{V}_{\rm S}}{\sqrt{\widetilde{Z}_{\rm S}}},\tag{7}$$

where Z_S is the source impedance. For the special cases of field radiated from an open-circuit voltage or short-circuit current, this simplifies to

$$\widetilde{E}_{\rm rad} = \frac{e^{-jkr}}{r} \widetilde{F}_{\rm V} \widetilde{V}, \quad \widetilde{F}_{\rm V} = \frac{j\omega}{4\pi c} \sqrt{\frac{377\Omega}{50\Omega}} \quad \frac{\widetilde{Z}_{\rm in} + 50\Omega}{\widetilde{Z}_{\rm in}} \widetilde{h}_{\rm N},$$

$$\widetilde{E}_{\rm rad} = \frac{e^{-jkr}}{r} \widetilde{F}_{\rm I} \widetilde{I}, \qquad \widetilde{F}_{\rm I} = \frac{j\omega}{4\pi c} \sqrt{\frac{377\Omega}{50\Omega}} (\widetilde{Z}_{\rm in} + 50\Omega) \widetilde{h}_{\rm N}.$$
(8)

So we see that $\tilde{h}_{N}(t)$ fully characterizes antenna performance for arbitrary source and load conditions.

5 A Name for $h_N(t)$

A function as important as $h_N(t)$ deserves a descriptive name, and our preferred name is "antenna impulse response." This might seem surprising since $h_N(t)$ is proportional to the responses to a step voltage in transmission and to an impulse field in reception. While we understand why this might cause confusion, it reflects the way that antennas actually behave. An antenna in transmission responds not to the source voltage but to its time derivative.

A second choice would be to refer to $h_N(t)$ as the "characteristic response" of the antenna. This has the considerable advantage that it will not be controversial. However, it conceals the fundamental nature of the function being described.

6 Example: The IRA-3Q

As an example, we provide data on the IRA-3Q, which is the current version of the 46-cm IRA offered by Farr Fields. A photo of the antenna is shown in Fig. 4, left, and its measured $h_N(t)$ is shown in Fig. 4 right. The shape of this function is a classic prepulse followed by an impulse.

7 Discussion

Let us consider now the limits on the application of the impulse response, $h_N(t)$. As formulated here, it applies to any antenna whose input port is a TEM source or load. It works most naturally with a 50- Ω source or load, but it handles other impedances as well. It only applies to an antenna radiating into free space. This includes many of the commonly used antennas, but certainly not all. Antennas with waveguide feeds are not treated. Furthermore, antennas radiating into a medium other than free space are not covered. These are all cases that can be handled with some enhancements to the theory, which will appear in a future paper.

Let us also consider whether the impulse response fully describes antenna performance. Since one can generate gain or realized gain from impulse response, it describes antenna performance at least as well as classical descriptions. So, for example, impulse response can be manipulated to find beam width at a given frequency, by first converting impulse response to gain. Thus, impulse response retains all the classical information of antenna response, while adding important time domain information.



Fig. 4 The Farr Fields model IRA-3Q, *left*, and its $h_N(t)$, *right*
8 Conclusions

We have defined a new function, the antenna impulse response, which describes antenna performance as simply as possible in the time or frequency domains. The function characterizes many of the most commonly used antennas in both transmission and reception and is simply related to gain, realized gain, and antenna factor. This will be a useful way of communicating antenna response, especially in the time domain. Extensions to this theory that cover more cases will appear in a future paper.

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Time-Domain Distortion Characterization of Electromagnetic Field Sensors Using Hermite–Gauss Subspaces

Shekoofeh Saboktakin and Behzad Kordi

Abstract Time-domain techniques have shown promising applications in the characterization of wideband electromagnetic field sensors. Distortion analysis in particular has only been studied in the time domain as it has no proper equivalent counterpart in the frequency domain. The conventional time-domain distortion characteristic is known as the sensor fidelity. Fidelity measures the similarity between the incident electromagnetic field and the sensor response by calculating a cross-correlation integral. The cross-correlation integral is dependent on the field waveform and therefore is not suitable for comparison purposes. A more general approach using Hermite-Gauss functions is discussed in this work. The presented approach gives a visual illustration of distortion characteristics and is not dependent on the incident electric field waveform. In this chapter, the application of the proposed method is demonstrated by evaluating the distortion characteristics of an Asymptotic Conical Dipole (ACD) that is compared to that of a monopole antenna with the same sensitivity. Time-domain simulation is performed by solving an Electric Field Integral Equation (EFIE) using the Method of Moment. Simulated voltages are converted into 2D graphs as well as scalar numbers for comparison purposes.

Keywords Electromagnetic field sensor • Distortion analysis • Time-domain analysis • Hermite–Gauss functions

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79

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1 Introduction

Electromagnetic field sensors are passive devices that can measure electric or magnetic field variation with time [1]. The incident electromagnetic field is usually a transient pulse. The sensors convert this pulse into an electric voltage or current at their terminals [2]. Electromagnetic field sensors are basically designed as antennas operating at the receiving mode. As a result, antenna measurement and characterization techniques can be applied to them. Antenna characterization conventionally deals with frequency-domain definitions such as gain, directivity, and radiation pattern. However, for the electromagnetic field sensors, time-domain analysis has more capabilities in characterization of the sensors, as in the frequency-domain methods transient features correspond to wide bandwidths and require transformations between the time and the frequency domains.

Baum and Farr have approached antenna time-domain characterization by using appropriate norms for the time-domain signals and applying conceptually frequency-domain definitions [3]. Shlivinski et al. used antenna time-domain effective height to address the time-domain gain definition [4]. Unlike antenna gain, which is inherently a frequency-domain concept, there is no clearly defined frequency-domain characteristic to analyze the distortion introduced by the antenna in the received waveform. Frequency-domain characteristics such as reflection loss and input impedance are logically related to the distortion analysis, but in addition to the reflection loss, antenna phase center should be a fixed point for the whole frequency range of interest. In the time domain, on the contrary, distortion level is translated to measure the similarity between the received waveform and the expected one. To define a distortion characteristic, one has to determine a mathematical means of comparing the similarity between the two waveforms and also develop a standard to interpret the comparison result. The so-called fidelity measures as the statistical similarity between the received and the actual waveforms [5]. A more general approach based on Hermite-Gauss polynomial orthogonal subspaces has been used recently to obtain the distortion characterization for an antenna for a subspace of waveforms [6].

In this chapter, Hermite–Gauss subspaces method is applied to compare the distortion characteristics of a monopole antenna and an Asymptotic Conical Dipole (ACD) [7]. ACDs are *D*-dot sensors, i.e., ideally, the output voltage is proportional to the incident electric field. Our simulation results show that ACDs are almost twice as sensitive to the incident electric field as the monopoles with the same height. In this work, a 5-cm-high ACD and a 14-cm-high monopole antenna are modeled as wire structures. The simulation is carried out in the time domain by solving an Electric Field Integral Equation (EFIE) using the Method of Moments (MoM) [8]. Hermite–Gauss subspaces method is applied to the time-domain simulated voltages, and a transformation matrix is calculated as a mean of distortion analysis. The transformation matrix can be interpreted visually as a two-dimensional graph. It is also possible to use a mathematical norm to obtain a scalar number as a measure of the distortion. In this chapter, in addition to the

graphical means provided by the transformation matrix, a spectral interpretation of the results is discussed to analyze the distortion characteristics in different bandwidths.

2 The Fidelity

For an antenna, fidelity is a time-domain characteristic which measures the similarity between antenna output waveform and the waveform ideally expected [5]. If e(t) represents the incident electric field and the received voltage is denoted by v(t), fidelity is defined as the cross-correlation of the received voltage and a linearly transformed version of the electric field, $L\{e(t)\}$, as

$$F = \left| \int v(t) \cdot L\{e(t)\} dt \right|.$$
(1)

The operator $L\{\cdot\}$ is determined by the ideal performance of the antenna. For example, for a *D*-dot sensor, this operator is a differentiator. If the waveforms are normalized fidelity varies between zero and one, with higher values corresponding to a higher similarity between the waveforms. Fidelity, as defined in (1), is dependent on the input waveform [6]. When comparing two or more sensors with each other, one has to evaluate (1) for all the possible sets of input waveforms [6].

3 Application of Subspaces Method in Distortion Analysis

An electric field sensor, as a linear system, converts the incident electric field, e(t), to an output voltage, v(t). From a mathematical point of view, the sensor operates as a linear transformation, $T\{\cdot\}$, which maps signals in an input vector space, into signals in an output vector space. As the incident electric field and the received waveform are signals with finite energies, the input and output vector spaces actually belong to the set of square-integrable functions, $L^2(\mathbb{R})$. $L^2(\mathbb{R})$ is an infinite dimensional vector space, which means it has infinite number of basis functions. However, a proper finite set of basis functions $\{\varphi_1, \varphi_2, \ldots, \varphi_N\}$, can be found that spans the input and output vector spaces up to the desired bandwidth. The linear transformation $T\{\cdot\}$, representing the sensor, has a corresponding transformation matrix Γ that maps the vector representation of the input signal to the vector representation of the output signal. In an *N*-dimensional space, Γ is an $N \times N$ matrix whose elements, $[\gamma_{ij}]$, are given by [9]

$$\gamma_{ij} = \frac{\int T[\varphi_i(t)] \cdot \varphi_j(t) dt}{\|\varphi_i(t)\|^2}.$$
(2)

In other words, the *ij*th element of matrix Γ is proportional to the inner product of $T\{\varphi_i(t)\}$ and $\varphi_j(t)$. If the basis functions are orthonormal, the proportionality coefficient is unity.

The transformation matrix represents how the sensor maps every single basis function onto the same vector space. As every signal in the input vector space can be constructed in terms of the basis functions, the transformation matrix shows how sensor responds to an arbitrary incident electric field waveform [6]. While Γ can give a measure of how the actual sensor performs, an ideal reference matrix, \mathbf{R}_{ref} , can be considered as the transformation matrix for the ideal sensor. The distance between the two matrices can be calculated as the norm of the difference between the transformation matrix and the reference one as

$$\eta_N = \|\mathbf{\Gamma} - \mathbf{R}_{\text{ref}}\|,\tag{3}$$

where the norm of an $N \times N$ matrix such as $\mathbf{A} = [a_{ij}]$ is calculated using the two norms given by

$$\|\mathbf{A}\| = \sqrt{\frac{\sum_{i} \sum_{j} a_{ij}^{2}}{N}}.$$
(4)

4 Hermite–Gauss Functions

Hermite-Gauss functions are defined in relation to Hermite polynomials as

$$\varphi_n\left(\frac{t}{\sigma}\right) = \frac{1}{\sqrt{n!2^n\sqrt{\pi}}} e^{\frac{-(t/\sigma)^2}{2}} H_n\left(\frac{t}{\sigma}\right),\tag{5}$$

where σ is a scaling factor which shrinks or expands the function to include the desired time interval. $H_n(t)$ is the *n*th-order Hermite polynomial. Hermite polynomials are classic orthogonal polynomials which arise in probability and quantum physics [10].

Figure 1 shows the time-domain waveforms for different orders of Hermite–Gauss functions given by (5). It is seen that these functions have finite time duration that increases with the order of the polynomials. Some of the useful properties of Hermite–Gauss functions are listed below:



- Hermite–Gauss functions form an orthonormal set that spans the space of all functions with finite energy, $L^2(\mathbb{R})$, and therefore can be used as basis functions for the subspace method.
- Hermite–Gauss functions are eigenfunctions of the Fourier transform which means they remain unchanged under the Fourier transform. As the time- and frequency-domain waveforms are the same for the unity σ , both the time duration and the frequency bandwidth are controlled by σ . Figure 2 shows how the 1 % bandwidth varies with *n* for $\sigma = 1$.

As shown in Fig. 2, as the order of Hermite–Gauss functions increases, they cover both a wider time window and frequency range. The value of σ should be selected such that the desired bandwidth is covered. Also, we need to use enough number of basis functions for a satisfying resolution.

5 Simulation Results

A time-domain simulation is performed to compare the distortion characteristics of a 14-cm-high monopole antenna with a 5-cm-high Asymptotic Conical Dipole (ACD). A height of 14 cm is chosen for the monopole so that its received voltage in the low-frequency spectrum is approximately the same as the ACD's received voltage. As it will be shown, at lower frequencies, both sensors operate as differentiators. In the simulation, an Electric Field Integral Equation (EFIE) is solved on a wire structure using the Method of Moments (MoM) [8]. Both antennas are terminated in 50- Ω loads. Figure 3 shows the wire structure model used to simulate the ACD. The electric field is polarized along the antenna axis. As shown in Fig. 3, to calculate [γ_{ij}], the *ij*th element of the transformation matrix Γ , the *i*th basis function is used as the incident electric field waveform. γ_{ij} is calculated using

$$\gamma_{ij} = \int \varphi_j(t) \cdot v(t) \big|_{E(t) = \phi_i(t)} dt.$$
(6)

6 Discussions and Conclusion

It can be shown that the transformation matrix for an ideal differentiator has nonzero elements only immediately above and below the diagonal elements, whereas the transformation matrix for an identity system is a diagonal matrix [11]. Figure 4 shows the transformation matrices obtained for the ACD and the monopole antenna. Known as *D*-dot sensors, the ACD and the short monopole should ideally deliver a time derivative of the incident electric field at the sensor terminals, or the electric field waveform should be preserved in the integral of the received voltages. Figure 5 shows the transformation matrices for the same sensors when calculated for the integral of the received voltages. As shown in Figs. 4 and 5, none of the sensors perform consistently for the whole frequency range. However, one can identify two frequency ranges: a low-frequency range that is up to 500 MHz and a high-frequency range which covers from 500 MHz and above.



Fig. 3 ACD wire model used for the simulation



Fig. 4 Transformation matrix $\Gamma = [\gamma_{ij}]$ calculated for (a) ACD and (b) monopole antenna



Fig. 5 Transformation matrix $\Gamma = [\gamma_{ij}]$ calculated for the integrated voltages for (a) ACD and (b) monopole antenna

6.1 Distortion Analysis at the High-Frequency Range

Figure 4a shows that for incident electric fields with orders higher than 30, corresponding to a bandwidth of 500 MHz and above, the ACD tends to map every single basis function mostly on itself as the transformation matrix is almost diagonal at the higher frequencies. The monopole transformation matrix, as shown in Fig. 4b, has strong diagonal components only up to 700 MHz and dispersion is clearly visible around the diagonal at the higher frequencies. This basically means that for those frequencies, the ACD performs more similar to an identity system than the monopole. To gain a numerical measure of distortion, a proper reference matrix must be selected. The reference transformation matrix in this case would be the identity matrix which has nonzero elements only on the diagonal. One can calculate the difference between the higher frequency part of the transformation matrices and a corresponding identity matrix. In this case, the upper 70 rows and columns of each matrix are taken as the high-frequency sub-matrices.

 η_N , given by (3), is calculated for the two sensors with the assumption of \mathbf{R}_{ref} being the 70 × 70 identity matrix, I_{70} . The corresponding values of η_N for the ACD and the monopole are 1.98 and 3.88, respectively. The numerical merit clearly confirms the visual evaluation of Figs. 4 and 5.

6.2 Distortion Analysis at the Low-Frequency Range

At the lower frequencies, both sensors are expected to deliver a voltage proportional to the derivative of the incident field at their terminals. In Fig. 4, a similar behavior is observed in the first 30 rows and columns corresponding to frequencies up to 500 MHz. In the lower 30×30 sub-matrices, only the elements on the top and at the bottom of the diagonal elements have nonzero values. To interpret the results, one can either compare the lower sub-matrices with an ideal differentiator transformation matrix. It can be shown that for an ideal differentiator, the transformation matrix has nonzero elements only on the immediately above and below the diagonal elements. Alternatively, we can take the integral of the simulated voltages and calculate the transformation matrix. For an ideal identity system, the transformation matrix should have nonzero diagonal elements only. Figure 5 shows a diagonal sub-matrix up to 500 MHz for both the ACD and the monopole; therefore, the sensors are performing as differentiators when the incident field is constructed mainly by the first 30 basis functions. To compare the derivative behaviors for the sensors, η_N is calculated for the lower 30 \times 30 sub-matrices of Fig. 5 when compared to I_{30} . The value of η_{30} is 1.00 for the ACD and 1.53 for the monopole. Visually, in Fig. 5, this is observed as the ACD's diagonal elements have all the same intensity up to 500 MHz, while for the monopole the intensity increases with the frequency.

In summary, the subspaces method gives visual and numerical measures of the sensors distortion characteristics for a subspace of waveforms rather than for a single arbitrary waveform. Using the linear property of the sensors, the method was applied to compare the distortion characteristics of an ACD and a monopole antenna having the same sensitivities. Hermite–Gauss functions are used as the basis set. As the spectral content of these functions is increased with increasing the order, a more general distortion analysis was conducted by analyzing the results in two frequency ranges. The proposed method is advantageous over time-domain fidelity in the sense that it is not dependent on the incident electric field waveform, whereas fidelity has to be calculated for every possible incident electric field waveform to have a comprehensive conclusion.

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Amplification of Signal Intensity in Imaging Through Discrete Random Media Through Signal Interference Gating and the Use of Mutual Coherence Function

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Abstract An approach to range-based imaging providing an enhanced "image bearing" component of signal intensity, and applicable to wide-band signals propagating through dilute, random, discrete-scatterer media is considered. The method is based on measuring the two-frequency mutual coherence function (MCF) of the signal and on a processing technique which suppresses diffusive contributions to the MCF and ensures a range resolution inversely proportional to the signal bandwidth.

Keywords Wide-band pulse • Discrete random media • Imaging • Mutual coherence function • Enhancement of image bearing component • Suppression of diffusion

1 Introduction

Developing a methodology which would allow efficient signal transmission, high resolution active imaging, and/or target detection through optically obscuring, dilute, discrete-scatterer media such as clouds, fog, dust, and other aerosols is of significant interest in commercial as well as military applications. Various approaches have been proposed to mitigate the detrimental effects of the media on the signal propagation:

Approaches Based on Coherent Detection. Considerable research has been done in the domain of reducing attenuation of the *amplitude* of a propagating wave averaged over the random medium configurations. This quantity is being measured in radar and ladar *coherent detection* type experiments.

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In particular, the effects of enhanced medium transparency in the propagation of wide-band pulses through dispersive homogeneous media (dielectrics, plasma) have been extensively analyzed in the context of precursor formation (see, e.g., [1-3], and references therein). Most of the research in the area was based on modeling the medium as a homogeneous dispersive dielectric material characterized by an effective permittivity or refractive index. This analysis has been recently extended to wave propagation in random discrete-scatterer media [4]. In this case, dispersion in the medium is due not only to the frequency-dependent (dispersive) losses in the medium constituents but also to the random scattering processes themselves; it may, therefore, exhibit properties different from those of usual bulk dielectrics and give rise to different behavior of the propagating pulses.

A further step in the direction of developing imaging methods of using wideband signals has been taken in [5] where the transmitted signal consists a coherent train of short wide-band pulses emitted at chirped (linearly varying) time intervals. As the energy of a single pulse, after traveling a large distance in the medium, is almost entirely concentrated in the precursor-type structures associated with the leading and trailing edge of the pulse, the increase of the signal total energy is achieved by using trains with a large number of pulses. The (down-)range resolution is controlled by the chirp bandwidth characterizing the distribution of individual pulses in the train (that bandwidth may be sufficiently large to achieve a resolution typical of millimeter-wave radar), while the high cross-range resolution can be attained in analogy to the usual synthetic-aperture imaging.

Although the developments described above take into account the discrete random nature of a medium, from the point of view of *coherent* wave propagation a discrete-scatterer medium can still be described by an effective complex permittivity, as if it were a homogeneous dielectric. In other words, coherent propagation does not "see" a difference between absorption (a genuine irreversible energy dissipation) and scattering-induced attenuation in which energy is not lost, but only randomized. The distinct roles of absorption and scattering mechanisms are taken advantage of in the approach described in this contribution.

The Proposed Approach: Mutual Coherence Function (MCF) Measurement and Interference Gating. In order to recover at least a part of the scattered energy and to utilize it for imaging purposes, two developments are needed: (1) imaging must be based not on the coherent detection of the mean field, but rather on measuring the field intensity or, more generally, the field MCF [6, 7] since it is this quantity that contains the full amount of the scattered energy and (2) it is necessary to devise methods for suppressing effects of the field diffusion since scattering, although conserves energy, leads to its "randomization." Reduction of diffusion effects by means of various "gating" methods has been a subject of much research, mostly in the context of dense media and medical imaging (e.g., [8-11]), although some estimates and simulations in the context of imaging through the atmosphere were also reported [12–15]. However, those approaches (the most common being time gating) are not well suited to long-range imaging, especially when the ranges to the targets and thus the arrival times of signals are not known. In view of these difficulties, we developed an alternative "interference gating" technique based on suppressing scattering mechanisms responsible for diffusion. We describe this approach in the following.

2 Description of the Approach

Problem Statement: Wave Propagation in Discrete-Scatterer Medium. We are considering statistically homogeneous media consisting of sparsely distributed particles of size *a* and of average separation $R \gg a$. We denote by $k \equiv k(\omega) = \omega/c$, the free-space wave number and by $\lambda \equiv \lambda(\omega) = 2\pi/k(\omega)$, the corresponding wavelength. Attenuation of the mean-field in the medium is characterized by the mean free path $\ell(\omega) = R^3/\sigma_t(\omega)$, given in terms of the medium number density R^{-3} and the total cross section for scattering on a single medium constituent, $\sigma_t(\omega)$. We also require the wavelength λ to be small relative to scatterer separation *R* and assume non-resonant scattering, hence $\sigma_t(\omega)$ bounded by the geometrical cross section $\sim a^2$. All these conditions can be summarized as $a \ll R \ll \ell(\omega)$, $a \le \lambda(\omega) \ll R$.

For simplicity, we assume fields to be scalar and denote them, in time and frequency domains, by $u(t, \mathbf{r})$ and $\tilde{u}(\omega, \mathbf{r})$. In this case, the frequency domain field intensity $|\tilde{u}(\omega, \mathbf{r})|^2$ can be identified with field energy density and the vector Im $[\tilde{u}^*(\omega, \mathbf{r}) \nabla \tilde{u}(\omega, \mathbf{r})]$ with the energy flux.

Ensemble averages, taken over realizations of a random medium, are denoted by $\langle \cdots \rangle$. In particular, the average (mean) field value and the general *two-frequency* MCF, known also as the "mutual intensity," are defined as $\tilde{M}_1(\omega; \mathbf{r}) := \langle \tilde{u}(\omega, \mathbf{r}) \rangle$ and $\tilde{M}_2(\omega_1, \omega_2; \mathbf{r}_1, \mathbf{r}_2) := \langle \tilde{u}(\omega_1, \mathbf{r}_1) \tilde{u}^*(\omega_2, \mathbf{r}_2) \rangle$, with analogous expressions for M_1 and M_2 in time domain. The usual field intensity is then the MCF \tilde{M}_2 evaluated at coinciding spatiotemporal points, $\tilde{I}(\omega; \mathbf{r}) := \langle |\tilde{u}(\omega, \mathbf{r})|^2 \rangle$ and $I(t; \mathbf{r}) := \langle |u(t, \mathbf{r})|^2 \rangle$. Experimentally, $\langle u \rangle$ and $\langle |u|^2 \rangle$ are measured in *coherent (heterodyne)* detection (see, e.g., [16]) and in *incoherent* detection (e.g., direct measurement of the field intensity by means of a photodiode).

Following the established terminology [17], we represent the average intensity as the sum of "coherent" and "incoherent" components, defined by

$$\tilde{I}(\omega; \mathbf{r}) = \langle |\tilde{u}(\omega, \mathbf{r})|^2 \rangle$$

$$= \underbrace{|\langle \tilde{u}(\omega, \mathbf{r}) \rangle|^2}_{\text{coherent}} + \underbrace{\langle |\tilde{u}(\omega, \mathbf{r})|^2 \rangle - |\langle \tilde{u}(\omega, \mathbf{r}) \rangle|^2}_{\text{incoherent}} =: \tilde{I}_{c}(\omega; \mathbf{r}) + \tilde{I}_{i}(\omega; \mathbf{r}), \quad (1)$$

with an analogous decomposition for the MCF \tilde{M}_2 .

For a deterministic (e.g., homogeneous) medium, $\langle |\tilde{u}|^2 \rangle = |\langle \tilde{u} \rangle|^2$, and the incoherent contributions are identically zero. However, in *disordered discrete-scatterer media*, $\langle |\tilde{u}|^2 \rangle \ge |\langle \tilde{u} \rangle|^2$, and, in many instances, $\langle |\tilde{u}|^2 \rangle \gg |\langle \tilde{u} \rangle|^2$ (or, $\tilde{I} \approx \tilde{I}_i \gg \tilde{I}_c$). The latter situation arises when the energy carried by the field

is relatively weakly absorbed (dissipated) and mostly scattered by the medium constituents, hence "transferred" from the coherent to the incoherent component of the intensity or the MCF.

Known Approaches to Extracting the "Image Bearing" Component of the Field Intensity and MCF. The above features of the signal propagation suggest that it may be worthwhile to try to recover, for imaging purposes, at least part of the energy contained in the incoherent intensity or MCF. The well-known difficulty, however, is that, although random scattering processes conserve energy, they cause its randomization. Such diffusive scattering processes give rise to a number of difficulties in imaging. In particular, range measurement based on the arrival time of scattered short pulses is hampered by the appearance of long tails in the received signals, caused by diffusion-type processes of non-forward scattering.

The problem of extracting an "image-bearing" part of the field intensity amounts, therefore, to devising a method of data processing which would suppress effects of diffusive processes and retain its coherent and forward-propagating parts.

A number of "gating" methods for suppressing diffusion effects are known. The most commonly used one is *time gating*, in which one selects only the earliest time interval of the received pulse, under the assumption that this part of the pulse intensity is least affected by the diffusive processes. This approach is effective when the time of arrival of the pulse is known a priori or when the peak corresponding to the pulse is observable and not masked by tails of the previous pulses. However, it may not be well suited to imaging of multiple reflecting objects (targets) located at unknown ranges. Some other methods, such as *angular (field of view)* and *spatial* gating eliminate signals arriving at the detector at large angles or outside a limited spatial area. These procedures rely on forward collimation of scattering processes, but for long propagation paths may not be selective enough in rejecting diffusive contributions and removing long temporal tails of the received signals.

The "Interference Gating" Method. The purpose of the proposed interference gating technique is similar to that of other gating procedures: suppressing the diffusive component of the total field intensity I (or the MCF) and retaining its image-bearing "forward-propagation" component.¹ However, this goal is being achieved in a quite different way: while the other methods select some propagation processes by imposing conditions on the properties (arrival time, angle, and location) of the fields at the *receiver*, the interference gating modifies properties of the *entire propagation history*. This possibility is due to the realization that the field intensity of a *pulse*—as opposed to a *monochromatic signal*—may propagate either diffusively or directionally, i.e., as a wave, depending on what frequencies in its spectrum are selected or suppressed.

¹ In the following we are discussing mostly field intensities, which are adequate quantities when imaging is based on *short* pulses. The analysis can be extended to long-chirped signals, in which case it becomes more complex and has to be based on the MCFs.

The key observation is that the time-domain intensity (the Fourier transform of (1)) is related to the two-frequency MCF by

$$I(t; \mathbf{r}) = \int \frac{\mathrm{d}\omega_1}{2\pi} \frac{\mathrm{d}\omega_2}{2\pi} \,\mathrm{e}^{-\mathrm{i}(\omega_1 - \omega_2)t} \tilde{M}_2(\omega_1, \omega_2; \mathbf{r}, \mathbf{r}), \tag{2}$$

hence a short pulse corresponds to a wide range in the relative frequency $\omega := \omega_1 - \omega_2$ of the two propagating fields. Now, in a sparse particulate medium composed of particles small compared to the wavelength, evolution of the incoherent (and dominant) component of the MCF \tilde{M}_2 can be well characterized by a sequence of interactions with scatterers located at some positions² r_1, r_2, \ldots , with propagation between the scatterers described by products of the Green functions and their conjugates,

$$\frac{1}{(4\pi R_{ij})^2} e^{-i[K(\omega_1) - K^*(\omega_2)]R_{ij}} \approx \frac{1}{(4\pi R_{ij})^2} e^{-\operatorname{Im}[K(\omega_1) + K(\omega_2)]R_{ij}} e^{i(\omega_1 - \omega_2)R_{ij}/c}, \quad (3)$$

where $R_{ij} = |\mathbf{r}_i - \mathbf{r}_j| \gg \lambda(\omega_1) \sim \lambda(\omega_2)$ and $K(\omega)$ is the effective wave number characterizing *coherent* propagation of the wave.³ In a sparse medium Re $K(\omega) \approx \omega/c$, while its imaginary part is responsible for the wave attenuation and related to the mean free path by Im $K(\omega) = 1/(2 \ell(\omega))$; hence, a typical propagation distance is $R_{ij} \approx 2 (\ell^{-1}(\omega_1) + \ell^{-1}(\omega_2))^{-1}$.

The intensity (2) contains contributions of both small and large relative frequencies $|\omega_1 - \omega_2|$. In the first case, if

$$|\omega_1 - \omega_2| \lesssim 2 c \left(\frac{1}{\ell(\omega_1)} + \frac{1}{\ell(\omega_2)}\right),\tag{4}$$

the phase factors in (3) nearly cancel, their product is non-oscillatory, and hence the propagation has a diffusive, random-walk-type, character. On the other hand, if the relative frequency is large, rapid oscillations of the exponential factor in (3) give rise to wave-type propagation along more-or-less straight lines. In media composed of scatterers large compared to the wavelength, an additional collimation arises from the individual scattering processes.

The above observations suggest that the diffusive component of the MCF can be significantly reduced by removing (filtering out), in the integral (2), the region of nearly equal frequencies defined by (4),

$$I_{\boldsymbol{\Phi}}(t;\boldsymbol{r}) = \int \frac{d\omega_1}{2\pi} \frac{d\omega_2}{2\pi} e^{-i(\omega_1 - \omega_2)t} \tilde{M}_2(\omega_1, \omega_2; \boldsymbol{r}, \boldsymbol{r}) \tilde{\boldsymbol{\Phi}}(\omega_1 - \omega_2),$$
(5)

² These positions are integrated over in computing the MCF.

³This qualitative description can be formalized in terms of the Dyson and Bethe–Salpeter equations for random media.

with $\Phi(\omega)$ being the high-pass filter. We refer to this procedure as "interference gating," since the remaining, large- ω contributions involve significant wave-type interference of the two frequencies.

It is also evident that for the interference gating to be possible without removing the entire signal spectrum, the initial pulse bandwidth must be sufficiently large

$$B \gtrsim \frac{c}{\ell(\omega)} \tag{6}$$

for the relevant frequencies ω in the pulse spectrum. The above constraint has, in fact, a simple physical meaning: the spatial extent of the pulse, or, equivalently, the attainable range resolution $\mathcal{R} \sim c/B$, has to be smaller than the mean free path—a reasonable condition in propagation through relatively dilute media.

An Imaging Scheme Based on Intensity or MCF Detection with Interference Gating. In range imaging utilizing short pulses, we envisage a scheme of incoherent detection of intensity $I(t; \mathbf{r})$, followed by interference gating, i.e., essentially, application of high-pass filtering of $I(t; \mathbf{r})$. This procedure should suppress the diffusive long tails of the received pulses and provide an adequate range resolution, inversely proportional to the pulse bandwidth.

As mentioned before, this scheme can be extended to imaging based on longchirped pulses. In this case it is not sufficient to process the usual field intensity; it is necessary to measure the two-time intensity

$$I_2(t+\tau/2, t-\tau/2; \mathbf{r}) := M_2(t+\tau/2, t-\tau/2; \mathbf{r}, \mathbf{r}),$$
(7)

which may be an experimental challenge in itself. The two-time intensity should be then processed by means of matched filtering in the relative time τ and, as before, by interference gating in the mid-point time *t*. It should be also possible to further extend the approach to synthetic-aperture imaging scenarios.

3 Examples and Possible Applications

The essence of the proposed gating method is recovery, for the purposes of imaging, of a part of the energy contained in the incoherent component of the propagating field intensity (or its MCF). It is evident that only the *scattered* field can be recovered as a useful signal, while the *absorbed* and dissipated field energy is irretrievably lost. Therefore, the most favorable circumstances for applications of the proposed approach are situations in which the wave attenuation is primarily caused by *scattering* and not by absorption. Possible applications may include the following:

1. propagation through clouds and fog in the infrared and optical regimes (where the water droplets are low-absorbing);

2. propagation through aerosols such as dust at lower frequencies (for dusts consisting of low-absorption dielectric particles at low frequencies).

As a preliminary illustration, we describe below a one-dimensional model of wave propagation involving Rayleigh scattering. It is followed by a discussion of three-dimensional propagation in the Mie scattering region.

A One-Dimensional Problem in Rayleigh-Scattering Regime. We consider a one-dimensional propagation problem described by a single spatial coordinate z, in which the scatterers are sparsely distributed *lossy* thin dielectric sheets of thicknesses much less than the wavelength $\lambda(\omega)$. A relatively simple computation, based on the Dyson equation, results in the coherent-propagation effective wave number of the form

$$K(\omega) = \frac{\omega}{\nu} + i \frac{\gamma(\omega)}{2}, \qquad (8)$$

where v is the wave speed, $\gamma(\omega) = 1/\ell(\omega) = n\sigma_t(\omega) = n(\sigma_s(\omega) + \sigma_a(\omega))$ is the attenuation coefficient expressible in terms of the number density n and the total scattering cross-section σ_t , or the absorption σ_a and the scattering σ_s cross-sections, and ℓ , defined as the inverse of the attenuation coefficient, is the mean free path.

Consider a point field source S(t) localized at z = 0. The two-field MCF becomes

$$\tilde{M}_2(\Omega + \frac{1}{2}\omega, \Omega - \frac{1}{2}\omega; z) = \tilde{\Gamma}_2(\Omega + \frac{1}{2}\omega, \Omega - \frac{1}{2}\omega; z)\tilde{S}(\omega_1)\tilde{S}^*(\omega_2), \qquad (9)$$

For a small relative frequency, $|\omega| := |\omega_1 - \omega_2| \ll \Omega := \frac{1}{2} (\omega_1 + \omega_2)$, the resulting expression for the Green function $\tilde{\Gamma}_2$ simplifies to

$$\tilde{\Gamma}_{2}(\Omega + \frac{1}{2}\omega, \Omega - \frac{1}{2}\omega; z) \approx \frac{c^{2}}{4\Omega^{2}} \left(\frac{\omega + \mathrm{i}\,c\,\gamma(\Omega)}{\omega + \mathrm{i}\,c\,\gamma_{a}(\Omega)}\right)^{1/2} \mathrm{e}^{\mathrm{i}\,\mu_{\Omega}(\omega)\,|z|},\tag{10}$$

i.e., the MCF propagation is controlled by the "effective wave number"

$$\mu_{\Omega}(\omega) \approx \left\{ \left[\omega/c + \mathrm{i}\,\gamma(\Omega) \right] \left[\omega/c + \mathrm{i}\,\gamma_{\mathrm{a}}(\Omega) \right] \right\}^{1/2},\tag{11}$$

where $\gamma_{a}(\omega) = n\sigma_{a}(\omega)$ is the absorptive component of the attenuation $\gamma(\omega)$.

If $\gamma_a/\gamma \ll 1$, i.e., if albedo defined as $\beta = \gamma_s/\gamma$ with $\gamma_s = n\sigma_s$ is close to 1, the intensity attenuation Im $\mu_{\Omega}(\omega)$ becomes small for small ω , which signals the onset of diffusive behavior; it manifests itself in the long temporal tails of the evolved pulses.

As an effectiveness assessment of the interference gating technique in suppressing tails of the pulses, we show in Fig. 1 the behavior of the total intensity I(t; z), its coherent contribution $I_c(t; z)$ and the interference-gated intensity $I_{\Phi}(t; z)$. Calculations were performed for an oscillatory gaussian pulse $S(t) = e^{i\Omega_0 t} e^{-\omega_s^2 t^2/2}$ of carrier frequency $\Omega_0/2\pi = 4$ GHz and width $1/\omega_s = \ell/4c$, and the gaussian filter $\tilde{\Phi}(\omega) = 1 - e^{-\omega^2/2\omega_{\Phi}^2}$ with $\omega_{\Phi} = c/\ell$. The medium was composed of scatterers 20 µm in size, the medium visibility was assumed 30 m $(n = 10^7 \text{ m}^{-3})$, the scatterers permittivity was $\varepsilon = 2.6 + i0.1$ and 2. 7 + i0. 2 which resulted in the



Fig. 1 Behavior of the coherent, total ("original"), and interference-gated intensities for single pulses ((a) and (b)) and for a sequence of two pulses ((c) and (d)) for the propagation distance $z = 6 \ell$ and for the two albedo values $\beta = 0.869$ and $\beta = 0.787$

mean free path of $\ell = 162$ and 142 m and the albedo of $\beta = 0.869$ and 0.787 respectively, the propagation distance was $z = 6 \ell$.

Figure 1a, b shows that the evolved total intensities, plotted as functions of the rescaled dimensionless variable ct/ℓ , exhibit very long diffusive tails, while the coherent contributions are short, but several orders of magnitude lower. Interference gating significantly suppresses the tails of the pulses, while only moderately reducing their peak values (by less than a factor of 3). The widths of the peaks become comparable to those of the coherent intensity (the structure in the peaks is due to the change of sign in the gated intensity).⁴

The significance of interference gating in achieving an adequate range resolution is illustrated in Fig. 1c, d, where we plot the evolved signal due to a sequence of two transmitted signals displaced by $\Delta(ct/\ell) = 10$ or, in time, by $\Delta t = 10 \ \ell/c$. In contrast to the original signal, the peaks in the processed signal are easily localizable and distinguishable, and they are at most only twice as wide as the peaks in the coherent MCF component. Hence, the time (and thus range) resolution attainable

⁴ The lack of positivity of the *processed* intensity does not violate any physical principles; neither does its lack of causality, although the latter could be achieved by using a causal filter.

with interference gating of the MCF would be within a factor of 2 of the resolution obtained with coherent detection, while the magnitude of the signal is at least two orders of magnitude higher than for the coherent intensity.

A Three-Dimensional Problem in Mie-Scattering Regime. Results qualitatively similar to the above can also be obtained in the case of wave propagation in three-dimensional random-scatterer medium composed of particles larger than the wavelength, e.g., in propagation of an optical or infrared laser beam through an atmospheric cloud or fog. In this case we can utilize the analysis of [18] and [19] based on a paraxial approximation, where the beam propagates at small angles relative to a fixed z axis. The behavior of the evolved MCF in this case differs from the previous one-dimensional problem mostly in that, in contrast to (11), its dependence on the z coordinate is not purely exponential; hence, the effective intensity attenuation depends not only on ω , but also on z. As a result, the interference gating (filtering) parameters may have to be adjusted as functions of the estimated propagation distance. Nevertheless, for intermediate propagation distances (comparable to the medium mean free path), the main effect of diffusion is the appearance of long flat tails in the evolved signals. In such a situation the proposed filtering method has effects similar to those in the previous examples.

4 Summary

The principal features of the proposed approach to pulse diffusion suppression by the interference gating method can be summarized as follows:

- The method is applicable to dilute random media under the conditions specified by (6), which are not difficult to satisfy in typical imaging scenarios.
- The method is expected to strongly suppress the diffusion-induced tails in the time dependence of pulses and provide gated signals much stronger than, and about as wide as, the coherent intensity component. In the analyzed one-dimensional example problem, for propagation distances $z \gtrsim 10 \ell$, the interference-gated signal was several orders of magnitude stronger and not more than twice as wide as the coherent intensity distribution.
- As a consequence, in range imaging based on the return time of the scattered pulsed signal, the proposed approach should be able to clearly separate peaks corresponding to different targets with nearly optimal resolution achievable for the given signal bandwidth. In contrast, as shown in the examples, in the absence of interference gating, those peaks could be barely discernible due to long diffusive tails of the pulses.
- The proposed method does not require directivity in wave scattering on the medium constituents and is effective even for scatterer sizes much smaller than the wavelength, i.e., for essentially isotropic scattering.
- It is applicable to wide-band pulses and dispersive media, hence to problems involving precursor-type phenomena.

- It can be applied to both short pulses and to long-chirped signals and implemented, together with chirped pulse compression, as a set of filters applied to the input two-frequency or two-time MCF.
- For short pulses interference gating reduces to simple high-pass filtering of the measured time-dependent ordinary field intensity.
- If desired, the proposed method can be used in conjunction with other gating techniques.

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99

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Propagation of Impulse-Like Waveforms Through the Ionosphere Modeled by Cold Plasma

D.V. Giri and S.L. Dvorak

Abstract In this chapter, we have studied the propagation of short, impulse-like pulses propagating through the ionosphere. The ionosphere is modeled by simple, cold plasma. The impulse response of such a plasma model is known to consist of two terms. The first term is the impulse itself and the second term contains a Bessel function of first order. This means that the impulse propagates as an impulse followed by a long, oscillatory tail. The numerical example studied here is that of the prototype impulse radiating antenna (IRA). Closed-form expressions are developed for the prototype IRA waveform propagation through the cold-plasma model of the ionosphere. The results are cross-checked with numerical evaluation via a convolution process that uses the known impulse response.

Keywords Short pulse • Ionosphere • Propagation • Cold plasma • Plasma frequency • Collision frequency • Dispersion

1 Introduction

The emerging short-pulse technologies will continue to find applications in several forms of radar, target identification, wideband jammers, UXO, and perhaps even orbital debris detection. The pulse durations are of the order of 100 ps and the bandwidths range from 10's of MHz to several GHz. All of the current and future

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UWB applications will involve the interaction or short-pulse electromagnetic waves with earth, water, metallic structures, dielectric surfaces, and earth's atmosphere. These objects could be the subjects of radar interrogation or may be in the path of radar's interrogation. In this chapter, we are focusing on the interaction of approximate impulses with the ionosphere. Each frequency component of the approximate impulse is imparted certain magnitude and phase change. The superposition of such effects over broad frequency ranges and large distances can significantly alter the incident electromagnetic pulse. The study of approximateimpulse propagation through the earth's atmosphere is of recent interest, e.g., [1–4]. When the ionosphere is modeled as cold plasma with certain, fixed plasma cutoff frequency $\omega_{\rm p}$, then the dispersion exhibited is similar to that in a singlemoded and homogeneously filled waveguide. The impulse response for this problem can be written in closed form in terms of the impulse function which propagates and is followed by a Bessel function term [2–4 among others]. Dvorak and Dudley [2] have also extended the analysis to a double exponential pulse propagating through cold plasma.

Given the above background, the present goal is to study the propagation of an impulse-like waveform generated by a reflector IRA in general [5, 6] and the prototype IRA in particular [7]. To a first-order approximation, we may consider the pre-pulse and impulse terms of the radiated field in the time and frequency domains [8] while ignoring the post-pulse term resulting from diffracted signals from the reflector rim and launcher plates.

2 Ionosphere as a Cold Plasma

The ionospheric effect on the information carried by radio waves is legendary and a recent review may be found in [9]. The ionosphere is an electromagnetically complex medium with diurnal, seasonal, and long-term variations driven primarily by both solar activity and local disturbances and plasma instabilities. Relevant information is available and we have summarized it in Figs. 1 and 2.

To a first order, the earth's atmosphere is horizontally stratified, distinguished through the temperature (Fig. 1). Various physical phenomena such as radiative transport in the infrared, convection cooling near the earth's surface, ultraviolet (UV) absorption by ozone layers in the troposphere, etc., determine this temperature profile. It may also be noted that the ionosphere is the charged component of the earth's atmosphere ranging from about 60 to 1,000 km. There is no significant ionization below 60 km. The electron density and the electron collision frequency as a function of the altitude are shown in Fig. 2. The maximum value of the electron collision frequency is seen to be about 100 kHz at an altitude of 60 km. For impulse-like waveforms, whose frequency extends from HF to S-band, the propagation through the ionosphere is essentially collision less and consequently lossless. Messier [1] has developed an expression for these losses due to electronic collisions, given by



Fig. 2 The electron density N and collision frequency v (Hz) profiles of the ionosphere [9]

$$A\left(\frac{\mathrm{dB}}{\mathrm{m}}\right) \cong 7.28 \times 10^{-6} \left[\frac{Nv}{f^2 + v^2}\right],\tag{1}$$

where N = electron density (m⁻³), v = electron collision frequency (Hz), and f = frequency of the propagating wave (Hz).

Since $f \gg v$, over large propagation paths, the effects of electronic collisions leading to Ohmic losses have some effect for frequencies below 100 MHz. In this chapter, we have ignored the effects of collisions since the short-pulse transmission under consideration has spectral magnitudes predominantly in the 100 MHz to 2 GHz band.

Furthermore, when the short-pulse waveform hits the D-layer of the ionosphere at an altitude of about 60 km, part of the incident wave is reflected at this air–ionosphere interface. It is well known that the ionosphere acts like a highpass filter where the frequencies below the plasma frequency ω_p are reflected and frequencies above ω_p are transmitted. In other words, one could model the ionosphere as a cold plasma with a plasma frequency of ω_p given by

$$\omega_{\rm p} = \sqrt{\frac{Ne^2}{m\varepsilon_0}} \cong 56.37\sqrt{N},\tag{2}$$

where $e \equiv$ electronic charge $\simeq 1.602 \times 10^{-19}$ C, $m \equiv$ electronic mass $\simeq 9.108$ \times 10⁻³¹ kg, and $\epsilon_0 \equiv$ permittivity of free space \cong 8.84 pF/m. Since the maximum value of N is 10^{12} electrons/m³, the maximum value of the plasma frequency, $f_{\rm p}$, is seen to be about 9 MHz. Once again, the frequencies of interest in the propagation of impulse-like waveforms are significantly above the plasma frequency $f_{\rm p}$ and certainly above the collision frequency v. In addition to the effects discussed above, the propagation of the short-pulse waveforms is also affected by the earth's magnetic field and refractive bending at non-normal incidence. The effect of the earth's magnetic field is to render the ionosphere plasma medium into a bi-refringent, nonreciprocal medium [10]. The refractive bending is predominant at frequencies near $\omega_{\rm p}$ and we have neglected the geomagnetic field effects. Simply stated the approximations made here are as follows (a) the ionosphere is modeled by a cold-plasma or a slab model with a plasma frequency of $\omega_{\rm p}$, independent of time and bounded by two values for changes with altitude (a slab model), (b) reflection coefficient at the air-ionosphere surface is 1 below ω_p and 0 above $\omega_{\rm p}$, and (c) geomagnetic field and refractive bending effects are ignored. Under these assumptions, the geometry of the problem now becomes simple and is shown in Fig. 3. θ is seen to be the polar angle of launch of the impulse-like waveform transmitter. z_0 is the vertical height of the uncharged, air medium which is roughly 60 km. The propagation distance is $z_0 \sec(\theta)$ in air and $z \sec(\theta)$ in the ionosphere. The z = 0 plane is the air-ionosphere interface. When the electromagnetic wave radiated from the transmitter on ground is incident at this interface, only the frequencies $> \omega_p$ are transmitted into the charged medium. With reference to the electron-density profile shown in Fig. 2 we consider $N = 1.12 \times 10^{12}$ /m³, which



Fig. 3 Geometry of the problem

corresponds to $f_p \cong 9.54$ MHz. This is similar to the dispersive wave propagation in a waveguide and the plasma frequency is analogous to the cutoff frequency of the dominant mode of propagation. Both the waveguide and the ionosphere behave like a high-pass filter.

3 Brief Review of the Impulse Response of a Cold-Plasma Medium

Under the assumption of a homogeneous, cold-plasma model, the ξ -polarized electric field, propagating in the direction of ξ , must satisfy the harmonic equation

$$\left(\frac{d^2}{d\xi^2} + k_{\xi}^2\right)\overline{E}_{\xi}(\xi,\omega) = 0,$$
(3)

where $k_{\xi} = \sqrt{k_0^2 - k_p^2}$, $k_0 = \omega \sqrt{\mu_0 \varepsilon_0}$ and $k_p = \omega_p \sqrt{\mu_0 \varepsilon_0}$.

The general solution for the above differential equation is

$$\widetilde{E}_{\xi}(\xi,\omega) = \widetilde{E}_{\xi}(0,\omega) \left[e^{-jk_{\xi}\xi} + \Gamma e^{jk_{\xi}\xi} \right].$$
(4)

Numerous researchers have shown that the impulse response of a waveguide or cold plasma can be expressed in dosed form in terms of the Bessel function of the first kind (e.g., [2–4]). For an impulse function input $E_{\xi}(0,t) = \tilde{E}_0 \delta(t)$, under the assumption of no reflection ($\Gamma = 0$), which is valid for $\omega > \omega_p$, Eq. (4) reduces to

$$E_{\xi}(\xi,t) = \frac{\widetilde{E}_0}{2\pi} \int_{-\infty}^{\infty} \exp\left\{j\left[\omega t - \frac{\xi}{c}\sqrt{\omega^2 - \omega_p^2}\right]\right\} d\omega$$
(5)

$$=\widetilde{E}_{0}\left[\delta\left(t-\frac{\xi}{c}\right)-u\left(t-\frac{\xi}{c}\right)\frac{\omega_{\mathrm{p}}\xi J_{1}\left(\sqrt{t^{2}-\left(\xi/c\right)^{2}}\right)}{c\sqrt{t^{2}-\left(\xi/c\right)^{2}}}\right]V/m.$$
(6)

This result is well established and is available in the literature [2–4]. However, some comments about the impulse response are in order. We observe that after propagating through a distance ξ in the cold plasma, the incident impulse $\tilde{E}_0\delta(t)$ reappears, but is followed by an oscillatory behavior in time, which is represented by the first-order Bessel function term. The radiated waveform from the IRA is an approximate impulse $\delta_a(t)$ and not an exact impulse. One may expect for approximate-impulse propagation that the initial peak would propagate but be compressed in duration due to the interaction term with the Bessel function portion. It would then be followed by an oscillatory behavior, approaching the plasma resonant frequency.

In the next section, we look at the incident fields radiated from a reflector IRA (e.g., the prototype IRA [7, 8]) and striking the air–ionosphere interface. That is, these are fields radiated from the prototype IRA that have propagated a distance of $z_0 \sec(\theta)$, where $z_0 = 60$ km and θ is the polar launch angle.

4 Propagation of IRA Fields Through Cold Plasma

Since the propagation distance in air is $z_0 \sec(\theta)$ (see Fig. 3), the electric field at the air–ionosphere interface from a reflector IRA may be written as [8]:

106

Propagation of Impulse-Like Waveforms Through the Ionosphere Modeled...

$$\widetilde{E}_{\xi}(\xi,t) = \frac{V_0}{z_0 \sec\left(\theta\right)} \frac{D}{4\pi c f_g} \left[\frac{\partial V(t'-T)}{\partial t} - \frac{1}{T} \{V(t') - V(t'-T)\} \right].$$
(7)

Note that this field is at z = 0, and

$$t' = t - \frac{z_0 \sec{(\theta)}}{c}, \quad T = \frac{2F}{C}$$

$$V(t) = V_0 (e^{-\beta t} - e^{-\alpha t}) u(t) \equiv \text{voltage waveform}$$

$$F \text{ and } D \text{ are length and diameter of the reflector}$$

$$c \equiv \text{speed of light, } f_g \equiv Z_{\text{feed}}/Z_0 \equiv \text{geometrical factor}$$

$$Z_{\text{feed}} \equiv \text{TEM feed impedance, } Z_0 \equiv \text{impedance of free space}$$
(8)

Substituting V(t) from Eq. (8) into Eq. (7), we find the electric field at the air–ionosphere interface (z = 0) in the time and frequency domains to be

$$\widetilde{E}_{\xi}(z=0,t) = \frac{V_0}{z_0 \sec{(\theta)}} \frac{D}{4\pi c f_g} \Big[\Big\{ \alpha e^{-\alpha(t'-T)} - \beta e^{-\beta(t'-T)} \Big\} u(t'-T) \\ -\frac{1}{T} \Big\{ \Big(e^{-\beta t'} - e^{-\alpha t'} \Big) u(t') - \Big[e^{-\beta(t'-T)} - e^{-\alpha(t'-T)} \Big] u(t'-T) \Big\} \Big],$$
(9)

$$\widetilde{E}_{\xi}(0,\omega) = \frac{\widetilde{V}(\omega)}{z_0 \sec\left(\theta\right)} \frac{D}{4\pi c f_g} e^{-j\omega t_0} \left[j\omega e^{-j\omega T} - \frac{1}{T} \left\{ 1 - e^{-j\omega T} \right\} \right], \quad (10)$$

$$t_0 = z_0 \sec(\theta) / \operatorname{can} d\widetilde{V}(\omega) = V_0 \left[\frac{1}{j\omega + \beta} - \frac{1}{j\omega - \alpha} \right].$$
(11)

Equations (10) and (11) completely specify the electric field incident at the air-ionosphere interface in the time and frequency domains. For the present, we continue the formulation of propagating the above-described incident field in the ionosphere. The propagation constant in the ionosphere is known, and ignoring the collisions, the spectral domain field after propagating a distance of $z \sec(\theta)$ in the ionosphere is given by

$$\widetilde{E}_{\xi}(z,\omega) = \frac{\widetilde{V}(\omega)}{(z+z_0)\sec\left(\theta\right)} e^{-\frac{j\omega}{c}[z_0\sec\left(\theta\right)]} e^{\frac{j\omega}{c}|z_0\sec\left(\theta\right)|} \sqrt{1 - \left(\frac{\omega_p^2}{\omega^2}\right)}.$$
(12)

The above frequency-domain field needs to be Fourier inverted to obtain the time-domain electric field. However, noting the form of $\widetilde{V}(\omega)$, we define

$$e(u,t) = \frac{1}{2\pi j} \int_{-\infty}^{\infty} \frac{e^{j\omega t} e^{-j\frac{z}{c}} \sec\left(\theta\right) \sqrt{\omega^2 - \omega_{\rm p}^2}}{(\omega - ju)} d\omega.$$
(13)

107

Then the Fourier inversion of Eq. (12) leads to

$$E_{\xi}(z,t) = \frac{V_0}{(z+z_0)\sec(\theta)} \frac{D}{8\pi f_{\rm g} f_{\rm d}} F(t),$$
(14)

where $f_{\rm d} = \frac{F}{D}$ and F(t) is given by

$$F(t) = [\{(1 - \beta T)e(\beta, t - t_0 - T) - e(\beta, t - t_0)\} - \{(1 - \alpha T)e(\alpha, t - t_0 - T) - e(\alpha, t - t_0)\}]$$
(15)

and $t_0 = z_0 \sec(\theta)/c$, T = 2F/c.

It is also observed that e(u,t) defined in Eq. (13) can be expressed in terms of certain special functions termed incomplete Lipschitz–Hankel (ILHI) functions [11, 12] that have integral representation in terms of the Bessel functions, which are useful in numerical computation of the fields received by a receiver in the ionosphere.

5 Incident IRA Waveforms and Spectra

While the previous sections have formulated the problem and developed closedform expressions for the wave propagation through the ionosphere modeled as a cold plasma, we illustrate this propagation with numerical examples in this and the following section. Specifically, we consider the electromagnetic fields radiated from the prototype IRA [7, 8]. The geometry of this electromagnetic short-pulse launch from earth to its atmosphere is depicted in earlier Fig. 3. The boresight electric field radiated by the prototype IRA is expressed in Eqs. (12) and (13) in time and frequency domains. The equations are suitably modified [8] for the case of a four-arm feed. The following numerical values are considered in this illustration.

Antenna parameters

$$D = 3.66 \text{ m}, F = 1.2 \text{m} \Rightarrow T = 2F/c \cong 8 \text{ ns}$$

$$Z_{\text{feed}} \cong 400 \Omega, f_{\sigma} = 1.06(2 \text{ arms}), f_{d} = F/D = 0.33$$

· Pulser parameters

$$V_0 = 10^5 \text{V}, \beta \cong 5 \times 10^7/\text{s}, \alpha \cong 2.2 \times 10^{10}/\text{s}$$

 $t_{10-90}(\text{rise}) \cong 100 \text{ ps}, t_{e-\text{fold}}(\text{decay}) \cong 20 \text{ ns}$

· Propagation in air

$$z_0 = 60 \text{ km}, \theta_i = 0^\circ, 15^\circ, 30^\circ, 45^\circ, 60^\circ, \text{ and } 75^\circ$$

(Normal incidence is reported here.)



Fig. 4 The electric field incident at the air–ionosphere interface ($z_0 = 60$ km; z = 0 and $\theta = 0^\circ$)

• Propagation in the ionosphere

$$N_i = 1.12 \times 10^{12} (\text{m}^{-3}), f_p = 9.54 \text{ MHz}, \text{ and } z = 500, 600, 700, \text{ and } 800 \text{ km}$$

As a first step, we present the field incident at the air–ionosphere interface in the time and frequency domains. The antenna and the pulser parameters are fixed and the vertical distance to the D-layer is also fixed at $z_0 = 60$ km. The previously measured time-domain electric field at a distance of 304 m is extrapolated to the air–ionosphere interface, and the computed spectrum are shown in Fig. 4.

6 Numerical Results for the Prototype IRA

In this section, we consider four values of the "slab" thicknesses for the ionosphere as mentioned earlier. These are the four vertical distances: $z_i = 500, 600, 700$, and 800 km. Although two vales of electron density and several launch angles were calculated, we present the case of $f_p \cong 9.54$ MHz only. This case corresponds to an electron density N of 1.12×10^{12} electrons/m³. As before, we compute both frequency- and time-domain signals after trans-ionospheric propagation. We first display the frequency-domain results in Fig. 5. The main feature in these spectral magnitude plots is that the ionosphere acts as a high-pass filter and all of the incident frequencies below the plasma frequency f_p of 9.54 MHz are filtered out or reflected back at the air–ionosphere interface, and the frequencies above f_p are transmitted and dispersed. The dispersion has a pronounced effect on the timedomain signals shown in Fig. 6. The ionospheric path z is varied from 500 to 800 km. For each value of z two time-domain plots (long and early time) are shown. The ringing frequency asymptotically approaches the plasma frequency of 9.54 MHz (period $\cong 100$ ns). The high frequencies travel with a larger velocity and arrive at the observer first, leading to dispersion. For frequencies $\omega \gg \omega_p$



Fig. 5 Spectral magnitudes after trans-ionospheric propagation $\theta = 0^{\circ}$

$$v_{\rm p} = \frac{c}{n} = \frac{c}{\sqrt{1 - \left(\frac{\omega_{\rm p}}{\omega}\right)^2}} \qquad v_{\rm g} = nc = c\sqrt{1 - \left(\frac{\omega_{\rm p}}{\omega}\right)^2}.$$
 (16)

7 Summary

We have formulated the problem of short-pulse propagation through the ionosphere, modeled by a homogeneous cold-plasma medium. The plasma frequency is fixed at its high value. The launch angle from a transmitter on ground and the ionospheric propagation distances are varied. As an example, the prototype IRA transmitter is considered for this illustrative numerical study. The high value of $N \approx 1.12 \times 10^{12}$ electrons/m³ or plasma frequency of 9.54 MHz leads to significant dispersion. The main impulse from the prototype IRA propagates through with a significant dispersion. In other words, the high frequencies travel faster to the observer and then the lower frequencies catch up. The effect of the ionosphere on the short pulse is dramatic at this higher electron density.



Fig. 6 Trans-ionospheric propagation for the case of $N \cong 1.12 \times 10^{12}$ electrons/m³ or $\omega_p \cong 6 \times 10^7$ or $f_p \cong 9.54$ MHz and $\theta = 0^{\circ}$

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¹ Sensor and Simulation and Theoretical Notes can be downloaded at http://www.ece.unm.edu/ summa/notes. This paper is an adaptation of Theoretical Note 366, which has extensive additional information.

Electromagnetic Environment of Grounding Systems

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Abstract Electromagnetic compatibility (EMC) and lightning protection studies in large installations require knowledge of spatial and temporal distribution of electromagnetic fields in case of lightning and power system faults. A new hybrid method for modeling electromagnetic environment of grounding systems is developed in this work. The electromagnetic fields in the surrounding soil are determined from the previously calculated current distribution using dipoles theory with analytical formulas. The model can be used to predict the EM environment of grounding systems because it can calculate electromagnetic fields in any points of interest.

Keywords Component • Hybrid method • Electromagnetic environment • Grounding systems

1 Introduction

The numerical modeling methods for grounding systems under lightning strokes developed since the early eighties can be classified as follows:

- Transmission line approach
- · Circuit approach
- EM field approach
- Hybrid approach

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1.1 Transmission Line Approach

The transmission line approach was the first method that was used for simulating transient behavior of grounding systems. Recently the conventional transmission line approach has been extended from simple grounding wire to grounding grid [1] and has been improved from uniform per-unit parameters to nonuniform per-unit parameters [2]. This approach can be either in time and in frequency domain, it can include all the mutual coupling between the grounding wires; it can also include the soil ionization. Moreover this approach can predict surge propagation delay. Further the computation time required by transmission line approach is extremely less compared with the electromagnetic approach.

1.2 Circuit Approach

The circuit approach for the transient analysis of grounding systems was developed by Meliopoulos et al. [3]. The main steps involved in this method are as follows:

- Divide the grounding system into many finite segments.
- Create the equivalent lumped circuit for each segment and calculate its parameters.
- Solve the nodal equation of the equivalent circuit that represents the whole grounding system based on Kirchhoff's laws.

Circuit approach is easy to understand, can easily incorporate the nonlinear soil ionization phenomena, and can include all the mutual coupling between the grounding wires. The main drawback of this approach is that it cannot predict the surge propagation delay.

1.3 Electromagnetic Field Approach

Electromagnetic field approach is the most rigorous method for modeling the transient behavior of grounding systems, because it solves full Maxwell's equations with minimum approximations. This approach can be implemented either by:

- Method of moment (MoM)
- Finite element method (FEM)

The model for the transient behavior of grounding system based on MoM was first developed by Grcev et al. [4]. This model aims to transform the associated electric field Maxwell's equations to a system of linear algebraic equations with minimum assumptions. However, this model is too complex to be implemented. Further, when the grounding structure is large, the computation time is very large. Another disadvantage of electromagnetic field approach is that, because of its frequency domain solution procedure, it cannot be easily modified to include nonlinearity due to soil ionization.

The electromagnetic field approach for the transient analysis of grounding systems based on FEM was developed by Biro and Preis [5]. This model starts from electric or magnetic energy equation, which involves partial differential Maxwell's equations.

The difficulty in this approach is to transform the open boundaries of both air and earth environment into a closed boundary problem using spatial transformation [6], which will reduce the size of the problem. The main advantage of this electromagnetic field approach based on FEM is that the discretization of the domain (geometry of the medium) of the problem can be highly flexible nonuniform patches or elements that can easily describe complex shapes. That is the reason why the soil ionization can be easily included into this model. However, this method is more complicated to understand, because it is not directly solving the Maxwell's equations.

1.4 Hybrid Approach

Hybrid approach for the transient analysis of grounding system was first initiated by Dawalibi [7]. This model is the combination of both electromagnetic field approach and circuit approach. This method includes the frequency influence on series internal impedances, inductive components, and capacitive-inductive components which make this method more accurate than the conventional circuit approach, especially when the injection source frequency is high.

1.5 Our Approach

In this work, we propose a new hybrid approach [8, 9], where three methods are summarized: analytical formula for determining electromagnetic fields radiated by electrical dipole in infinite conductive medium, modified images theory for taking in account the interface in the half space instead of Somerfield's integrals, and transmission line theory for determining the longitudinal and leakage current.

In infinite medium, the total electromagnetic fields are the sum of the contributions from each dipole. In semi-infinite medium, two cases can be considered: the first case is the current source (dipole) and observation point in the same medium (the electromagnetic fields can be evaluated as a sum of the field of the current source and its image) and the second case is the current source in medium.1 and observation point in medium.2 (the electromagnetic fields can be evaluated as the field due only to the modified current source).
Our method can be either in time and in frequency domain, it's very easy to understand, reasonably accurate, and time efficient.

Expressions of Electromagnetic Fields in Frequency 2 Domain

The rigorous expressions for electromagnetic fields in the conducting medium are developed. Consider an electric dipole of length (dl) immersed in a conducting medium characterized by constitutive constants: conductivity (σ), permeability (μ), and permittivity (ε) , and excited by an impulse current.

When the dipole is located in the origin of a Cartesian coordinate system, and oriented in the z direction, the vector potential in the frequency domain is as follows:

$$\vec{A}(r,s) = \mu_0 \frac{ldl}{4\pi r} e^{-\gamma r} \vec{k}, \qquad (1)$$

where r = distance between the observation and source points and $\gamma = \sqrt{\mu s(\sigma + \varepsilon s)}; \quad s = j\omega.$ Using

$$\vec{H}(r,s) = \frac{1}{\mu_0} \operatorname{rot} \vec{A}(r,s)$$
(2)

we take the magnetic field components in the frequency domain:

$$dH_x(r,s) = \frac{-yIdl}{4\pi r^3} (1+\gamma r) e^{-\gamma r},$$
(3)

$$dH_{y}(r,s) = \frac{xIdl}{4\pi r^{3}} (1+\gamma r) e^{-\gamma r}, \qquad (4)$$

$$dH_z(r,s) = 0. (5)$$

And using Maxwell equations,

$$\vec{E} = \frac{1}{\sigma + j\omega\varepsilon} \operatorname{rot}\left(\frac{1}{\mu_0} \operatorname{rot}\vec{A}\right)$$
(6)

we take the electric field components in the frequency domain:

Electromagnetic Environment of Grounding Systems

$$E_x(r,s) = \frac{3xzIdl}{4\pi(\sigma + \varepsilon s)r^5} \left(1 + \gamma r + \frac{\gamma^2 r^2}{3}\right) e^{-\gamma r},\tag{7}$$

$$E_{y}(r,s) = \frac{3yzIdl}{4\pi(\sigma+\varepsilon s)r^{5}} \left(1+\gamma r+\frac{\gamma^{2}r^{2}}{3}\right)e^{-\gamma r},$$
(8)

$$E_z(r,s) = \frac{Idl}{4\pi(\sigma + \varepsilon s)r^3} \left(\frac{3z^2}{r^2} \left(1 + \gamma r + \frac{\gamma^2 r^2}{3} \right) \right) e^{-\gamma r}.$$
(9)

3 Expressions of Electromagnetic Fields in Time Domain

From the tables of Laplace transforms [10], we have

$$e^{-\gamma \cdot r} \Leftrightarrow e^{\frac{-r\alpha}{2}} \delta(t - r/\nu) + \frac{\alpha r}{2} e^{\frac{-t}{2\tau_0}} I_1(m) \frac{u(t - r/\nu)}{\sqrt{t^2 - r^2/\nu^2}},\tag{10}$$

where $m = \frac{\sqrt{t^2 - r^2/v^2}}{2\tau_0}$, $\delta(t - r/v)$ is the Dirac function, u(t - r/v) is the Heaviside step function, and $I_1(m)$ is the first-order modified Bessel function.

We also define the attenuation constant, the wave velocity, and the relaxation time, respectively: $\alpha = \sigma \sqrt{\mu/\epsilon}$, $v = 1/\sqrt{\epsilon\mu}$, and $\tau_0 = \epsilon/\sigma$.

Using Eq. (7), we take the expression of potential vector in time domain:

$$\vec{A}(r,t) = \frac{\mu I \Delta l}{4\pi r} \left[e^{\frac{-\alpha r}{2}} \delta(t-r/v) + \frac{\alpha r}{2} e^{\frac{-t}{2r_0}} I_1(m) \frac{u(t-r/v)}{\sqrt{t^2 - r^2/v^2}} \right] \vec{k}.$$
 (11)

3.1 Time-Domain Expressions of Magnetic Field

Using

$$\vec{H}(r,t) = \frac{1}{\mu} \nabla \wedge \vec{A}(r,t)$$
(12)

we take the magnetic field components in the time domain:

$$dh_{x}(r,t) = \frac{-yIdl}{4\pi r^{3}} \left\{ \left[\frac{r}{v} \frac{\partial}{\partial t} \delta(t-r/v) + \left(1 + \frac{\alpha r}{2} + \frac{\alpha^{2} r^{2}}{8} \right) \delta(t-r/v) \right] e^{\frac{-\alpha r}{2}} + e^{\frac{r^{2}}{2t_{0}}} \frac{\alpha^{2} r^{3} I_{2}(m)u(t-r/v)}{4v(t^{2}-r^{2}/v^{2})} \right\},$$

$$(13)$$

$$dh_{x}(r,t) = \frac{xIdl}{2t_{0}} \int \left[\frac{r}{2t_{0}} \frac{\partial}{\partial t} \delta(t-r/v) + \left(1 + \frac{\alpha r}{2} + \frac{\alpha^{2} r^{2}}{2} \right) \delta(t-r/v) \right] e^{\frac{-\alpha r}{2}}$$

$$dh_{y}(r,t) = \frac{xIdl}{4\pi r^{3}} \left\{ \left[\frac{r}{v} \frac{\partial}{\partial t} \delta(t-r/v) + \left(1 + \frac{\alpha r}{2} + \frac{\alpha^{2}r^{2}}{8} \right) \delta(t-r/v) \right] e^{\frac{-\alpha r}{2}} + e^{\frac{-t}{2r_{0}}} \frac{\alpha^{2}r^{3}I_{2}(m)u(t-r/v)}{4v(t^{2}-r^{2}/v^{2})} \right\},$$
(14)

where $I_2(m)$ is the second-order modified Bessel function.

Comparing the expressions of magnetic field components Eqs. (3), (4), and (13), (14) we take the inverse [11] of Laplace transform of

$$G_2(r,s) = (1+\gamma r)e^{-\gamma r}$$
 (15)

$$g_{2}(r,t) = e^{\frac{-t}{2r_{0}}}I_{2}(m)\frac{\alpha^{2}r^{3}u(t-r/\nu)}{4\nu(t^{2}-r^{2}/\nu^{2})} + \left[\left(1+\frac{\alpha r}{2}+\frac{\alpha^{2}r^{2}}{8}\right)\delta(t-r/\nu) + \frac{r}{\nu}\frac{\partial}{\partial t}\delta(t-r/\nu)\right]e^{\frac{-\alpha r}{2}}.$$
(16)

3.2 Time-Domain Expressions of Electric Field

The components of electric field can be rewritten as follows:

$$dE_x(r,s) = \frac{3xzIdl}{4\pi r^5\varepsilon} \left(G_2(r,s)G_3(r,s) + \frac{r^2}{3v^2}sG_1(r,s) \right),$$
 (17)

$$dE_{y}(r,s) = \frac{3yzIdl}{4\pi r^{5}\varepsilon} \left(G_{2}(r,s)G_{3}(r,s) + \frac{r^{2}}{3v^{2}}sG_{1}(r,s) \right),$$
(18)

$$dE_z(r,s) = \frac{Idl}{4\pi r^5 \varepsilon} \left[\left(3z^2 - r^2 \right) G_2(r,s) G_3(r,s) + \frac{r^2}{v^2} \left(z^2 - r^2 \right) s G_1(r,s) \right], \quad (19)$$

where

$$G_1(r,s) = e^{-\gamma r}, \ G_2(r,s) = (1+\gamma r)e^{-\gamma r}, G_3(r,s) = \frac{1}{s+1/\tau_0}.$$

Applying the convolution theorem of the Laplace transform, the electric field in the time domain can be obtained as

$$de_x(r,t) = \frac{3xzIdl}{4\pi r^5\varepsilon} \left(\frac{r^2\partial}{3v^2\partial t}g_1(r,t) + \int_0^t g_2(r,t)g_3(r,t-\tau)d\tau\right),\tag{20}$$

where

$$g_3(r,t) = e^{\frac{-i}{r_0}}u(t) \tag{21}$$

$$de_{x}(r,t) = \frac{3xzIdl}{4\pi r^{5}\varepsilon} \left\{ u(t-r/v) \left[\left(1 - \frac{\alpha r}{2} + \frac{\alpha^{2}r^{2}}{8} \right) e^{\frac{-(t-r/2v)}{\tau_{0}}} + \frac{\alpha^{2}r^{2}}{12v} \frac{1}{(t^{2} - r^{2}/v^{2})} \left(tI_{2}(m) - \sqrt{t^{2} - r^{2}/v^{2}} I_{1}(m) \right) + \frac{\alpha^{2}r^{3}}{4v} \int_{r/v}^{t} \frac{e^{-(t-\tau/2)/\tau_{0}}}{\tau^{2} - r^{2}/v^{2}} I_{2} \left(\frac{\sqrt{\tau^{2} - r^{2}/v^{2}}}{2\tau_{0}} \right) d\tau \right] + \frac{r}{v} e^{\frac{-\alpha r}{2}} \left[\delta(t-r/v) \left(1 + \frac{\alpha^{2}r^{2}}{24} \right) + \frac{r}{3v} \frac{\partial}{\partial t} \delta(t-r/v) \right] \right\}.$$
(22)

We can obtain with the same manner the components $de_y(t)$ and $de_z(t)$. Worth noting that up to here, no approximation has been made; therefore the expressions in the time domain of the magnetic and electric fields, $h_x(r, t)$, $h_y(r, t)$, and $e_x(r, t)$, given in Eqs. (13), (14), and (22), are considered to be rigorous and new.

4 Semi-infinite Medium

In the half space the interface soil-air is taken into account using modified images theory. The electric field radiated by a current element placed above or below the earth's surface can be evaluated by the modified method of images [12]. The following cases can be considered for the position of the current element and the observation point.



4.1 Current Source and Observation Point in Soil

The electromagnetic field can be evaluated as a sum of the field of the current source and its image I' as follows (Fig. 1):

$$I' = \frac{(\sigma_2 + j\omega\varepsilon_2) - (\sigma_1 + j\omega\varepsilon_1)}{(\sigma_1 + j\omega\varepsilon_1) + (\sigma_2 + j\omega\varepsilon_2)}I = R(\omega)I.$$
(23)

4.2 Current Source in Soil and Observation Point in Air

The electromagnetic field can be evaluated as the field due to the modified current source (I'') (Fig. 2):

$$I'' = \frac{2(\sigma_1 + j\omega\varepsilon_1)}{(\sigma_1 + j\omega\varepsilon_1) + (\sigma_2 + j\omega\varepsilon_2)} I = T(\omega)I.$$
(24)

5 Longitudinal and Leakage Currents

The transmission line approach for the transient analysis of buried grid based on the finite difference time-domain method (FDTD) has been developed.

To determine the longitudinal and leakage currents in the ground conductor, we propose the direct resolution of the propagation equation in time domain, by the FDTD method.

Transmission line equations in potential and current in time domain for one dimension are given by

$$\begin{cases} \frac{\partial U}{\partial \chi} = -RI - L \frac{\partial I}{\partial t} \\ \frac{\partial I}{\partial \chi} = -GU - C \frac{\partial U}{\partial t} \end{cases} \quad \chi = x \text{ or } y. \tag{25}$$

R, *L*, *C*, and *G* are the per-unit length parameters of the buried conductor [13].

The propagation equation is obtained by combination of two equations in system (25):

$$\frac{\partial^2 U}{\partial x^2} + \frac{\partial^2 U}{\partial y^2} - 2(RC + LG)\frac{\partial U}{\partial t} + 2RGU - 2LC\frac{\partial^2 U}{\partial t^2} = 0.$$
 (26)

The partial derivatives can be approximated by finite differences at point of coordinates (i, j, n); the subscripts *i*, *j*, and *n* are respectively associated to variables *x*, *y*, and time. Substituting the partial derivatives by their approximations into Eq. (26), we obtain

$$\frac{1}{\Delta x^{2}}U(i-1,j,n) + \frac{1}{\Delta y^{2}}U(i,j-1,n) + \frac{1}{\Delta x^{2}}U(i+1,j,n) + \frac{1}{\Delta y^{2}}U(i,j+1,n) \\
- \left[\frac{2}{\Delta x^{2}} + \frac{2}{\Delta y^{2}} + 2RG + \frac{2(RC+LG)}{\Delta t} + \frac{2LC}{\Delta t^{2}}\right]U(i,j,n) \\
= \frac{2LG}{\Delta t^{2}}U(i,j,n-2) - \left[\frac{2(RC+LG)}{\Delta t} + \frac{2LC}{\Delta t^{2}}\right]U(i,j,n-1).$$
(27)

By writing this equation on all points of the buried grid, we can generate the following linear matrix equation:

$$[A] [U] = [B]. (28)$$

The resolution of this system gives the node voltage on the buried grid. This resolution requires the knowledge of suitable conditions in extremities of the grid. Then, the voltage at the injection point and at the extremities (on borders of the grid) must be fixed.

Fig. 3 Buried grid



Once the transient voltages responses are computed, the currents in different branches of grounding grid are obtained by numerical integration of the following current line equation:

$$\frac{\partial U}{\partial \chi} = -RI - L \frac{\partial I}{\partial t} \quad \chi = x \text{ or } y.$$
⁽²⁹⁾

6 Application and Validation

For analysis we take the same example treated by Grcev et al. [4], a 60 m by 60 m square ground grid with 10 m by 10 m meshes, made of copper conductor with 1.4 cm diameter, and buried at a depth of 0.5 m under the earth's surface. The soil is assumed to be homogenous with a resistivity 100 Ω m, a relative permittivity 9 and a relative permeability 1.

In the following applications, we use the typical double exponential lightning current impulse given by

$$I(t) = I_0 (e^{-\alpha t} - e^{-\beta t})$$
(30)

with $I_0 = 1.635$ KA; $\alpha = 0.1421/\mu s$; $\beta = 1.0731/\mu s$.

6.1 Injection at the Corner Point of the Grid

In the first application, the lightning stroke is fed at the corner point of the grid (Fig. 3). Figure 4 shows the spatial distribution form 3D of the magnetic field, to remote ground at the soil surface (70 m \times 70 m) parallel to and centered on the grid, at $t = 10 \ \mu$ s.



Fig. 4 Magnetic field 3D ($t = 20 \,\mu s$) corner injection



Fig. 5 Magnetic field 3D ($t = 20 \ \mu s$) middle injection

6.2 Injection at the Middle Point of the Grid

In the second application, the lightning stroke is fed at the middle point of the grid. Figure 5 shows the spatial distribution form 3D of the magnetic field to remote ground at the soil surface (70 m \times 70 m) parallel to and centered on the grid, at $t = 10 \,\mu$ s.



Fig. 6 The physical situation for two electrodes

Presented results show large differences of the magnetic field to remote ground between points at the interface. High values of the magnetic field occur near the injecting point and are further spreading toward the rest of the ground surface while the values are decreasing.

The different locations of feed point are shown in Figs. 9 and 12. The curves in Figs. 10 and 13 compare the magnitude of magnetic field, for the feed point at one corner and the center of the grid with the same current injection.

For the same grounding grid, the maximal magnitude of magnetic field, for feed point at the center is much smaller than that for feed point at the corner. The location of feed point at the center is strongly recommended, instead of at the corner.

6.3 Injection at the Middle Point of Two Conductors

The physical situation is displayed in Fig. 6. We consider two electrodes connected at the middle. The conductor is 15 m long, located at 1 m beneath the air-soil interface. A typical double exponential lightning current impulse is injected at the connection point.

With $I_0 = 1.0167$ A, $\alpha = 0.0142/\mu s$, $\beta = 5.073/\mu s$.

Figure 7 illustrates the electric field (component Ex) along 20 m profile at the soil surface parallel to and centered on the conductor as depicted on Fig. 6.

For two conductors (2×15 m length, feed point at middle), the injected impulse current can flow into earth in four directions. We obtain in this case the several maximal electric fields near impact point (40 V/m).



Fig. 7 Electric field components Ex

7 Conclusion

In this work, a hybrid method for analysis transient electromagnetic fields behavior of grounding grid under lightning stroke is presented. This study is based on transmission line approach, electrical dipole theory, and modified images theory. In this formalism, the computation is carried out in two steps: one numerical, for determining the current distribution on the grid by direct resolution of the propagation equation in the temporal space, using the finite difference time-domain method (FDTD), and the other, analytical, for calculating electromagnetic fields components. The interface between earth and air is taken into account using modified image theory instead of Somerfield's integrals.

The computation results are based on a general formulation, in time domain, which permit the observation point in air or in soil, and are in good agreement with that based on electromagnetic fields approach. The location of feed point at the center is strongly recommended, instead of at the corner.

The computation results were successfully compared to those obtained in [7] and [14].

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A Global Circuit Tool for Modeling Lightning Indirect Effects on Aircraft

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Abstract The topic of this study is electromagnetic environment and electromagnetic interference (EMI) effects, specifically the modeling of lightning indirect effects on aircraft electrical systems present on embedded and highly exposed equipments, such as nose landing gear (NLG) and nacelles, through a circuit approach. The main goal of the presented work, funded by a French national project, PREFACE, is to propose a simple equivalent electrical circuit to represent a geometrical structure, taking into account mutual, self-inductances, and resistances, which play a fundamental role in the lightning current distribution. Then this model is intended to be coupled to a functional one, describing a power train chain composed of a converter, a shielded power harness, and a motor or a set of resistors used as a load for the converter. The novelty here is to provide a pre-sizing qualitative approach allowing playing on integration in pre-design phases. This tool intends to offer a user-friendly way for replying rapidly to calls for tender, taking into account the lightning constraints.

Keywords Lightning modeling • Short pulse • Equivalent electrical circuit • Lightning indirect effect • More electrical aircraft • More composite aircraft

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1 Introduction

The evolution of the aeronautical field, in the context of a "more composite" and "more electrical" aircraft, leads to a series of modifications in the design of the electrical systems. The introduction of the composite materials [1, 2] in the aircraft structure, in order to save weight and to improve robustness to aging, generates important environment changes. The composite fuselage cannot entirely comply, neither with the grounding and bonding nor with the lightning strike protection requirements. Thus, in order to ensure a proper functioning of the electrical system and to protect the electrical equipment or the High-Voltage Direct Current network, we have to estimate constraints on system and equipment levels in order to minimize the weight and optimize the distribution of innovative protections.

The principle of our circuit approach is to generate automatically an equivalent passive circuit in the shape of a SABER[®] circuit simulator macro-component. Inputs and outputs can be defined by the user in order to be coupled with other functional circuit models in the same simulator such as converters or harnesses.

Therefore, the computation of the lightning current distribution on the structure and the power harness shield is done in time domain by the circuit simulator and allows predicting common mode voltage and current transients on the three phases of the converter.

The advantage of this circuit approach, compared to the commonly used Finite Difference Time Domain method, is a very fast computation tool allowing parametrical and sensitivity studies that are not feasible with 3D simulations because of time-consuming issues. Also the point of impact of the lightning strike can be moved on the structure in order to search for worst-case scenarios.

Moreover, thanks to a global modeling approach, we aim to address the optimization and the distribution of the lightning protections either on system or equipment level.

The case study is the following: an NLG as shown in Fig. 1, which is composed of tubular pieces that can be easily approximated by equivalent cylindrical straight conductors. Therefore, passive R, L, M elements of the structure can be extracted through analytical engineer formulas such as those implemented in the partial element equivalent circuit (PEEC) [3] technique.

2 Equivalent Circuit Generation

We are interested in the generation of an equivalent circuit composed of passive components. The methodology used can be described with four major steps.

The first step is the simplification of the structure geometry. This is done by identifying the structure main parts, thus allowing their discretization as straight conductors or fixed diameter cylinders as described in the third step.



Fig. 1 A320 Nose landing gear 3D CAD

The second one is the identification of contacts between parts, giving rise to contact resistors. Then, it is necessary to know the value of contact resistors between the structure's parts. These resistors play a major role in the limitation of the current magnitude and in the impedance propagation path. Their value has been measured in the ONERA lab and introduced in our model by changing the conductivity of the conductors being in contact. The result on the apparent conductivity is the same as putting physically the contact resistors.

We emphasize here the fact that even without a theoretical model able to predict these contact resistor values and moreover their change with the current magnitude, e.g., breakdown effects, we can use the circuit parameter in order to check the main tendencies as the simulation duration is about a few seconds on a single core processor to be compared to hours of computation on multiprocessor calculators for FDTD simulations.

The third step is a coarse discretization of the whole structure but within the lengths l_k and the relative positions of the parts in space. Each part is represented by a cylinder with a resistivity ρ_k , associated to the material physical property. The diameter d_k of the cylinder is in fact an effective diameter corresponding to the piece cross section S_k .

The definition of d_k is therefore given in Eq. (1):

$$d_k = \sqrt{\frac{4S_k}{\pi}}.$$
 (1)

Hence, regardless of the shape of the part, like a tube or a more complex form, we model it by a straight cylindrical conductor.

The last step is the computation of all the passive components associated to the structure's parts such as linear resistors R_k , linear inductors L_k , and mutual inductors M_{ik} , respectively, defined by Eqs. (2), (3), and (4) [4]:

$$R_k = \frac{\rho_k l_k}{\pi d_k^2 / 4},\tag{2}$$

$$L_k = \frac{\mu_0 l_k}{2\pi} \ln\left(\frac{4l_k}{d_k} - 1\right),\tag{3}$$

$$M_{ik} = \frac{\mu_0}{4\pi} \oint_{c_1c_2} \frac{d\vec{l}_i \cdot d\vec{l}_k}{r_{ik}}$$
(4)

In Eqs. (3) and (4), μ_0 is the vacuum permeability, and the variables c1 and c2 stand for the lengths of two conductors *i* and *k*. As we want to propose a generic tool for lightning constraints pre-design evaluation, we developed a process in order to generate automatically the equivalent circuit to the structure and harnesses from their geometry and physical properties. At the moment, it still requires the intervention of an engineer for the structure simplification, especially in order to identify the material properties, the parts lengths, and the cross sections, but some work can be done in order to improve this step.

The idea is to obtain a macro-component, usable under a circuit simulator, whose inputs and outputs can be defined by the user, and all the parameter, computed with the previous equations under Matlab[®], are written into a netlist file associated to this component. By this way, we prevent ourselves from routing and schematic drawing issues and moreover from having to enter manually the parameter's values, especially the mutual inductors between all parts which can reach a few thousands different values.

The required steps are summarized in Fig. 2.

Finally, we can solve the indirect lightning current distribution, when applying a bi-exponential current generator, using a standard DO-160 "A"-type waveform with 200 kA peak value. The validation of the equivalent circuit is described in the next section through a comparison with experimental results.

3 Experimental Validation

Measurements of indirect lightning current distribution have been carried out, on an A320 nose landing gear, by the ONERA French laboratory. We can notice that the peak level of the "A"-type waveform has been reduced to 430 A instead of 200 kA, while the shape of the waveform was preserved. The results were normalized to the peak magnitude ratios in order to compare easily experimental measurements and



Fig. 2 SABER[®] macro-component generation

simulation results. We can see on the test setup in Fig. 3 that the waveform generator is located between the NLG compass tip and the copper tape, used for the current return.

The simplified structure is shown on Fig. 4, as described previously, a bi-exponential current generator is used to inject the lightning waveform and current distribution, represented by the variables J1–J5, is measured on output braids, the landing gear structure (J1–J5), and on the power harness shields (A0 and B0).

Comparison between simulations and measurements is shown on Fig. 5. For each current, the dashed black lines represent the measurements while the continuous colored lines stand for circuit simulations.

We can observe that the peak value and the global shape of the currents are all well estimated except for J3, corresponding to the current on the interface between the primary drag stay and the aircraft, with a relative overestimated error of 20 %.

However, we are more interested on the currents on the power harness shields, showing less than 3 dB error, as they will induce the threat by transfer impedance directly to the electrical equipment such as the converter used to control wheel breaks

This scenario is addressed in the next section.



Fig. 3 ONERA experimental test setup



Fig. 4 Simplified NLG structure



Fig. 5 Lightning current distribution comparison. *Dashed lines* correspond to measurements while *colored lines* represent simulation results

4 Global Simulation: Preliminary Results

The main goal of using a circuit approach, in addition to a gain in terms of computation time, is to be able to couple the effect of the lightning-induced currents on the complete chain of our electrical systems. In order to achieve that, a global simulation can be carried out by combining various equipment models in the same circuit simulator, e.g., the model of the converter, the power harness, and the electrical motor as shown schematically in Fig. 6.

Therefore, we can estimate the perturbation in terms of common and differential mode currents and voltages on the three phases of the converter in order to optimize the protections required for the equipment.

We can observe, on Fig. 7, the "A" standard waveform simulating a direct strike on the NLG structure, represented by the green curve in the top graph of the picture. The second graph represents the common mode (black curve) and the differential mode current (green curve) measured at the converter inputs and that would be seen by the aircraft High-Voltage Direct Current (HVDC) network. We can observe that the differential mode current is close to zero while common mode reaches 150 A maximum value.

The third graph shows the perturbation induced on the typical common mode voltage, reaching nearly 3.7 kV in absolute value.

In the last graph, the currents induced on the three phases of the converter are approximately the same, reaching 40 A. Obviously, all these values are destructives



Fig. 6 Global simulation schematic



Fig. 7 Lightning-induced current on the common and differential mode currents

both for the converter and the electrical motor and require the use of appropriate protections. Attention should be paid to the fact that this global simulation is still at a feasible study stage, demonstrating that it is possible to combine complex time domain models of electrical equipments with a lightning transient on a geometrical structure. Also, the value of the induced current and voltages are overestimated as we are using a geometrical mutual inductance coupling which did not yet take into account the transfer impedance of the harness shield.

5 Conclusion

In the present work, we have demonstrated the feasibility to model lightning indirect effects through a fast solving circuit approach. Simulation results for indirect lightning current distribution were compared with experimental measurements for the case of an A320 NLG, showing a great accuracy, similar to classical 3D FDTD technique, usually used for lightning modeling.

It was shown that it is possible to couple the circuit model in more global simulation approach, allowing estimating constraints directly on system and equipment levels, in order to optimize the required protections.

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Fully Absorbing Conditions in the Study of Axially Symmetrical UWB Radiators

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Abstract In this chapter we develop rigorous mathematical models for open resonators and radiators of pulsed and monochromatic waves with waveguide feed lines. Analysis of such structures is of considerable importance for solving a large number of theoretical and applied problems in microwave engineering, antenna design, and high-power electronics. The exact absorbing conditions described in this chapter provide a rigorous framework for these studies.

Keywords Exact absorbing conditions • Axially symmetrical structures

1 Introduction

Exact absorbing conditions (EAC) are used in computational electrodynamics of non-sinusoidal waves for truncating the domain of computation when replacing the original open initial boundary value problem by a modified problem formulated in a bounded domain. The main intention of this chapter is to construct EAC as applied to the analysis of compact axially symmetrical electrodynamic structures illuminated by the pulsed TE_{0n} - and TM_{0n} -waves and to prove the equivalency of the original (open) and modified (closed) initial boundary value problems. The analytic and physical results presented below—EAC for artificial spherical boundary, transport operators relating near-zone and far-zone pulsed fields, proof of the unique solvability of the modified closed problems, and their equivalence to original open problems and so on—represent a rigorous theoretical justification of the EAC method and show its numerical efficiency.

The essence of the method for open scalar problems, which are considered in the part Ω of the space \mathbf{R}^2 , is as follows. Assume that excitation sources and medium

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inhomogeneities are located in a bounded region Ω_{int} of the unbounded analysis domain Ω . The propagation velocity of the electromagnetic waves U(g, t) generated by appropriate sources and obstacles is finite. Therefore, during the observation time $t \leq T < \infty$, the signal U(g, t) will not cross the boundary Π_T of some bounded domain $\Omega_{\text{int, }T} \subset \Omega$: $U(g,t)|_{g \in \Omega_{\text{ext, }T}, t \in [0,T]} = 0$. Here, $\Omega_{\text{ext, }T}$ is the complement of domain $\Omega_{\text{int},T}$ with respect to Ω , that is, $\Omega = \Omega_{\text{int},T} \cup \Pi_T \cup \Omega_{\text{ext},T}$. This formula gives us the well-known exact radiation condition (ERC) for the outgoing (from the domain Ω_{int} where all sources and obstacles are located) pulsed waves U(g, t). The only but rather essential limitation of this simple condition is associated with the fact that with growing T, the domain $\Omega_{int,T}$ is expanding and the boundary Π_T is moving farther away from the domain Ω_{int} . Thus the ERC is not used for truncating the computational domain of open electrodynamic problems. In EAC method, the ERC is transferred from the field-free points $g \in \Omega_{\text{ext},T}$ onto some artificial boundary Γ located in the region, where the intensity of space-time field transformations can be arbitrary in magnitude: there is an operator D such that $D[U(g, t)]|_{g \in \Gamma} = 0$, $t \ge 0$. The electromagnetic waves U(g, t) must be *outgoing* in this case as well, or in other words, they are bound to intersect the boundary Γ in one direction only, moving away from the sources and obstacles.

The boundary Γ divides the unbounded domain Ω into two domains, namely, Ω_{int} and Ω_{ext} such that $\Omega = \Omega_{int} \cup \Omega_{ext} \cup \Gamma$. In the first one (bounded), we can formulate the initial boundary value problem with respect to the function U(g, t) with the help of EAC. This problem will be called the modified problem as distinct from the original problem formulated in the unbounded domain Ω with the ERC involved. In the domain Ω_{int} , the desired function U(g, t) can be determined by using the standard finite-difference algorithm. For the domain Ω_{ext} , the EAC method allows us to construct and use the so-called transport operators $Z_{q \in \Gamma \to g \in \Omega_{ext}}(t)[U]$ to determine the values of the function U(g, t) at points $g \in \Omega_{ext}$ from its values on the boundary Γ .

The EAC can be included into a standard finite-difference algorithm with the domain of calculation reduced to Ω_{int} . However, one can confidently assert that these finite-difference computational schemes are stable and convergent only when the modified problem is uniquely solvable and equivalent to the original problem. In the present work, we formulate the relevant theorem for the closed problems describing TE_{0n}- and TM_{0n}-pulse wave scattering in compact open axially symmetrical structures.

2 Formulation and Solution of the Problem

In Fig. 1, the cross section of a model for an open axially symmetrical $(\partial/\partial \phi \equiv 0)$ resonant structure with two waveguide feeding lines $(\Omega_1 \text{ and } \Omega_2)$ is shown, where $\{\rho, \phi, z\}$ are cylindrical and $\{r, \vartheta, \phi\}$ are spherical coordinates. By $\Sigma = \Sigma_{\phi} \times [0, 2\pi]$, we denote perfectly conducting surfaces obtained by rotating the curve Σ_{ϕ} about the



Fig. 1 Axially symmetrical node with two (coaxial and round) feeding waveguides (a) 3D geometry and (b) geometry of the model problem (1)

z-axis; $\Sigma^{\epsilon,\sigma} = \Sigma^{\epsilon,\sigma}_{\phi} \times [0, 2\pi]$ is a similarly defined surface across which the relative permittivity $\epsilon(g)$ and specific conductivity $\sigma_0(g) = \eta_0^{-1}\sigma(g)$ change stepwise. Piecewise constant functions $\epsilon(g)$ and $\sigma_0(g)$ are nonnegative inside Ω_{int} and take free space values outside ($\epsilon = 1$ and $\sigma_0 = 0$); $\eta_0 = (\mu_0/\epsilon_0)^{1/2}$ is the impedance of free space; ϵ_0 and μ_0 are the electric and magnetic constants of vacuum. The distribution of pulsed axially symmetrical waves in the open structures of this kind is given by

$$\begin{cases} \left[-\varepsilon(g)\frac{\partial^2}{\partial t^2} - \sigma(g)\frac{\partial}{\partial t} + \frac{\partial^2}{\partial z^2} + \frac{\partial}{\partial \rho}\left(\frac{1}{\rho}\frac{\partial}{\partial \rho}\rho\right) \right] U(g,t) = F(g,t), \quad t > 0, \ g \in \Omega \\ \\ U(g,t)|_{t=0} = \varphi(g), \frac{\partial}{\partial t}U(g,t) \Big|_{t=0} = \psi(g), \quad g = \{\rho, z\} \in \overline{\Omega} \\ \\ E_{tg}(p,t)|_{p=\{\rho,\phi,z\}\in\Sigma} = 0, \quad t \ge 0 \\ \\ E_{tg}(p,t) \text{ and } H_{tg}(p,t) \text{ are continuous when crossing } \Sigma^{\varepsilon,\sigma} \\ U(0,z,t) = 0, \quad |z| < \infty, \ t \ge 0 \\ \\ D_1[U(g,t) - U^{i(1)}(g,t)]|_{g \in \Gamma_1} = 0, \ D_2[U(g,t)]|_{g \in \Gamma_2} = 0, \quad t \ge 0. \end{cases}$$

$$(1)$$

Here, $U(g, t) = E_{\phi}(g, t)$ for TE_{0n}-waves and $U(g, t) = H_{\phi}(g, t)$ for TM_{0n}-waves. The domain of analysis Ω is the part of the half-plane $\phi = \pi/2$ bounded by the contours Σ_{ϕ} together with the artificial boundaries Γ_j (input and output ports) in the virtual waveguides Ω_j , j = 1, 2. The regions Ω_{int} and Ω_{ext} (free space) are separated by the virtual boundary Γ . The SI system of units is used. The variable *t*, being the product of the real time by the velocity of light in free space, has the dimensions of length. Dimensions are omitted throughout this work.

The functions F(g, t), $\varphi(g)$, $\psi(g)$, $\sigma(g)$, and $\varepsilon(g) - 1$, which are finite in closure of Ω , are supposed to satisfy the theorem on the unique solvability of problem (4) in the Sobolev space $\mathbf{W}_2^1(\Omega^T)$, $\Omega^T = \Omega \times (0,T)$, $T < \infty$. The "current" and "instantaneous" sources given by the functions F(g, t) and $\varphi(g)$, $\psi(g)$, as well as all scattering elements given by functions $\varepsilon(g)$, $\sigma(g)$ and by the contours Σ_{ϕ} and $\Sigma_{\phi}^{\varepsilon,\sigma}$, are located in the region Ω_{int} .

The operators $D_j[U]$ are described in [1] and provide an idealized model for fields emitted and absorbed by the virtual waveguides Ω_j . These operators specify the conditions for artificial boundaries Γ_j in the cross sections of the input coaxial waveguide (j = 1) and the output circular waveguide (j = 2). The function $U^{i(1)}(g, t)$ entering the first of these conditions defines the wave incident on the boundary Γ_1 . This function, as well as the source functions F(g, t), $\varphi(g)$, and $\psi(g)$, is assumed to be given.

We replace open problem (1) by the closed problem with a bounded domain of analysis Ω_{int} by imposing the additional boundary condition $D[U(g, t)]|_{g \in \Gamma} = 0$, $t \ge 0$. The operator D[U] is defined by the exact radiation condition

$$U(r,\vartheta,t) = \frac{L}{2r} \sum_{n=1}^{\infty} \left\{ \int_{t-(r-L)}^{t-(r-L)} \left[\left(\pm \frac{L-rq}{rL\sqrt{1-q^2}} P_n^1(q) - \frac{1}{L} P_n(q) \right) \int_0^{\pi} U(L,\vartheta_1,\tau) \widetilde{\mu}_n(\cos\vartheta_1) \right] \\ \times \sin\vartheta_1 d\vartheta_1 - P_n(q) \int_0^{\pi} \frac{\partial U(r,\vartheta_1,\tau)}{\partial r} \left| \widetilde{\mu}_n(\cos\vartheta_1) \sin\vartheta_1 d\vartheta_1 \right] d\tau \\ + \int_0^{\pi} \left[U(L,\vartheta_1,t-(r-L)) + (-1)^n U(L,\vartheta_1,t-(r+L)) \right] \\ \times \widetilde{\mu}_n(\cos\vartheta_1) \sin\vartheta_1 d\vartheta \right\} \mu_n(\cos\vartheta), \quad r \ge L, \ 0 \le \vartheta \le \pi,$$

$$(2)$$

if we set r = L. Here, $\tilde{\mu}_n(\cos \vartheta) = \sqrt{(2n+1)/(2n(n+1))}P_n^1(\cos \vartheta)$, $P_n^1(\ldots)$ are the associated Legendre functions, $P_n(\ldots)$ are the Legendre polynomials, and $q = [r^2 + L^2 - (t - \tau)^2]/2rL$.

As a consequence of [2], we can assert the following statement.

Let problem (1) have a unique solution in $\mathbf{W}_2^1(\Omega^T)$, $\Omega^T = \Omega \times (0,T)$, $0 \le t \le T < \infty$. Then, the modified (closed) problem is uniquely solvable in the space $W_2^1(\Omega_{int}^T)$, $\Omega_{int}^T = \Omega_{int} \times (0,T)$, and together with Eq. (2), it is equivalent to the open problem (1).

This result allows us to construct stable and convergent finite-difference algorithms for computing approximate values of the field U(g, t) everywhere in Ω for the times $0 \le t \le T < \infty$. The accuracy of the EAC implemented in this way has been demonstrated in [3] in the following ways. First, the accuracy of sources and their propagation in the discretized computational volume terminated by the EAC was rigorously checked against exact solutions. Secondly, for simple modelling problems, we determined how to choose parameters so that the reflection of waves from virtual boundaries had minimal influence on the accuracy of the calculated characteristics. Finally, for more complex problems with significant spatial inhomogeneities, we validated our result against those of other authors using different approaches (see, e.g., Sect. 4.2.2 of [3]).

3 Numerical Implementation and Results

Let us discuss some results obtained in numerical experiments based on the proposed algorithm. The case in point is radiation of pulsed and monochromatic waves by dual-reflector communication antennas, whose directional patterns can be controlled over sufficiently wide limits. Algorithms of construction of these antennas for efficiently radiating TE_{0n} - and TM_{0n} -waves in prescribed directions $\overline{\vartheta}$ are presented in the chapter [4]. Those techniques are based on the well-known results of the geometrical theory of diffraction and on the classical results of the theory of the second-order curves.

Below, we implement one of the above-mentioned schemes to determine the main geometrical parameters of the antennas composed of two mirrors—elliptic and parabolic mirrors—and to compute their electrodynamic characteristics.

For numerical analysis, we have taken an antenna whose parabolic reflector axis is directed at the angle $\vartheta = 90.0^{\circ}$ (Fig. 2a). The antenna was illuminated by a coaxial waveguide with dimensions $a_1 = 1.0$ and $b_1 = 0.12$ with the pulsed TE₀₁-wave $U_1^{i(l)}(g, t) = v_{11}(z, t)\mu_{11}(\rho) : v_{11}^{\rho}(-L_1, t) = F_1(t), \quad \tilde{k} = 9.5, \quad \Delta k = 5,$ $\tilde{T} = 25, \quad \overline{T} = 50, \quad F_1(t) = 4 \sin \left[\Delta k \left(t - \tilde{T}\right)\right] \cos \left[\tilde{k} \left(t - \tilde{T}\right)\right] \chi \left(\overline{T} - t\right) \left(t - \tilde{T}\right)^{-1},$ occupying the frequency range $4.5 \le k = 2\pi/\lambda \le 14.5$. Here, λ is the wavelength in free space, $v_{11}^{\rho}(z, t)$ is the space-time amplitude of the H_{ρ} -component of the incident wave, $\mu_{11}(\rho)$ is its transversal function, and $\chi(\ldots)$ is the Heaviside step function.

We have considered the range $4.5 \le k \le 14.0$ ($0.45 \le \lambda \le 1.4$). Within this range, the input waveguide sustains propagation of 1–3 undamped TE_{0n}waves. In the *normalized directional pattern* on the arc $r = M \ge L D(\vartheta, k, M)$ $= |\widetilde{E}_{\phi}(M, \vartheta, k)|^2 / \max_{0 < \vartheta < \pi} |\widetilde{E}_{\phi}(M, \vartheta, k)|^2$ ($0 \le \vartheta \le 180^\circ$, $K_1 \le k \le K_2$) of the antenna under consideration, a single main lobe directed at the angle $\overline{\vartheta}$ dominates in the range k > 9.0. Here $\widetilde{E}_{\phi}(M, \vartheta, k)$ denotes the transform $\int_0^T e^{ikt} \widetilde{E}_{\phi}(M, \vartheta, k) dt$ where *T* is the upper limit of the observation time interval. This angle differs from



Fig. 2 Geometry of the antenna (the radius of the first reflector is c = 4.2; the radius of the second reflector is d = 10.12; $L_1 = 8.0$, L = 16.0; the axis of the second reflector is directed at the angle $\vartheta = 90.0^\circ$) and the field pattern for $E_{\phi}(g,t)$ at the instant the principal part of the pulse propagates within the computational domain Ω_{int} (**a**); the directional pattern for k = 12.0 (**b**) and in the frequency band 4.5 $\leq k \leq 14$ (**c**); the radiation efficiency k (**d**)

the expected value only slightly (Fig. 2b). Most likely this distinction is caused by the fact that the antinodes of the wave $U_1^{i(l)}(g, t)$ do not coincide exactly with the first focus of elliptic mirror; also, with the current space step of the mesh $\overline{h} = 0.02$, not all geometric parameters of the antennas could be assigned sufficiently accurately.

The point k = 9.0 ($\lambda = 0.7$) appears to be considered as a threshold for the applicability of the geometrical theory of diffraction in the study of the antennas of this kind. The analysis in the range k < 9.0 requires more rigorous approaches. In the short-wave part of the frequency range under study, the directional pattern (Fig. 2c) and the radiating efficiency (Fig. 2d) remain practically unchanged over the band. In this range the antenna under consideration represents a rather wideband structure fully complying with the requirements of the impulse technology and high-power electronics devices. Below are given the results of the numerical experiment supporting our assertion.

Let the ultra wideband Gaussian TE_{01} -pulse be defined by $U_1^{i(l)}(g, t) = v_{11}(z, t)\mu_{11}(\rho) : v_{11}^{\rho}(-L_1, t) = F_2(t), \quad \tilde{k} = 12.0, \quad \tilde{\alpha} = 0.6, \quad \tilde{T} = 4.0, \quad \overline{T} = 8.0,$ $F_2(t) = \exp\left[-\left(t-\tilde{T}\right)^2/4\tilde{\alpha}^2\right] \cos\left[\tilde{k}\left(t-\tilde{T}\right)\right]\chi(\overline{T}-t)$, where $v_{11}(z, t)$ is the space-time amplitude of the H_{ϕ} -component of the incident wave. The width $B_k = 2(k_{upp} - k_{low})/(k_{upp} + k_{low}) \times 100 \%$ of the band occupied by the pulse $U_1^{i(l)}(g, t)$ equals 50 %; it occupies the frequency range $9.0 \le k \le 15.0$. Suppose the pulse is incident on the antenna whose parabolic mirror is directed at the angle $\vartheta = 90.0^{\circ}$. The pulse radiated by the antenna intersects the boundary r = L = 16.0 of the domain of computation Ω_{int} at times $38 \le t \le 42$ (Fig. 3a). That means that it is expected to pass the arc r = L = 30.0 at times $52 \le t \le 56$.

The pulsed directional pattern $D_P(\vartheta, t, M) = U(M, \vartheta, t) / \max_{\vartheta, t} |U(M, \vartheta, t)|$, $0 \le \vartheta \le 180^\circ$, $M = 30 \ge L$, $T_1 \le t \le T_2 \le T + M - L$ presented in Fig. 3b has been determined for that time interval; it is plotted for the observation angles ϑ varying from 75° to 105°. The pulse radiates from the antenna as a compact spacetime pulse retaining almost completely the temporal and amplitude-frequency characteristics of the primary pulse $U_1^{i(l)}(g, t)$. This result allows us to conclude that the structure under study can be considered as an "ideal" pulse antenna (for frequencies k > 9.0), i.e., the antenna radiates a waveform similar to the exciting pulse.

One additional remark needs to be made. At relatively large distances from the antenna, the amplitude centre of the radiated pulse is bound to pass the point $\vartheta \approx 89^{\circ}$ (Fig. 2b); however, with the distance r = M = 30, it is observed at the point $\vartheta \approx 94^{\circ}$ (Fig. 3b). The reason is that initially (see Fig. 3a) the pulse is located somewhat below the plane z = 0, and it is directed towards $\vartheta \approx 89^{\circ}$ only upon passing this plane.



Fig. 3 The excitation of the antenna whose parabolic mirror is directed at the angle $\vartheta = 90.0^{\circ}$ by the ultra-wideband Gaussian pulse. Geometry of the antenna and the field pattern for $E_{\phi}(g,t)$ at t = 35.0 (a); the pulsed directional pattern on the arc r = M = 30.0 (b)

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Part II Time Domain Computational Techniques

The New Vector Fitting Approach to Multiple Convex Obstacles Modeling for UWB Propagation Channels

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Abstract This chapter presents the new approach to time-domain modeling of UWB channels containing multiple convex obstacles. Vector fitting (VF) algorithm (rational approximation) was used for deriving the closed form impulse response of multiple diffraction ray creeping on a cascade of convex obstacles. VF algorithm was performed with respect to new generalized variables proportional to frequency but including geometrical parameters of the obstacles also. The limits of approximation domain for vector fitting algorithm follow the range of ultra-wideband (UWB) channel parameters that can be met in practical UWB channel scenarios. Finally, the closed form impulse response of a creeping UTD ray was obtained. As the result we obtained impulse response of the obstacles, time. It permits for calculation of channel responses for various objects without changing the body of a rational function. In that way the presented approach is general, simple, and effective.

Keywords UWB • UTD • Time domain (TD) • Frequency domain (FD) • Vector fitting (VF) • Inverse Fourier transform

1 Introduction

Ultra-wideband technology enables many beneficial features in transmission and radar area. In order to take the advantage of these features, careful UWB system analysis is performed. The important component of the UWB system is a propagation channel. The natural choice of the domain for the UWB propagation analysis is the time domain with the usage of impulse responses which can be found in an

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empirical way or a theoretical way. We deal with theoretical effective time-domain modeling of UWB channels that comprise obstacles (e.g., people) which can be modeled by convex objects (cylinders in 3D case or ovals in 2D case—analyzed in the chapter). Our aim is to present the procedure for obtaining the closed form impulse response for the multiple diffraction creeping ray. This problem was considered in [1], where two kinds of solution were given for obtaining the impulse response of a creeping ray. The first of them uses VF approximation of a creeping ray transfer function in frequency domain. The disadvantage of this solution is that VF approximation must be performed for each pair of values of parameters *R* and θ in ξ_d (Eq. 1). Therefore the values of poles and residues resulting from VF approximation are different for each creeping ray scenario and the solution is not universal. Alternatively the second solution from [1] gives the closed form impulse response of a creeping ray but in two separate formulas dedicated for two cases of values range of parameters ξ_d [1, 2]:

$$\xi_d(\omega, R, \theta) = \left(\frac{\omega \cdot R}{2 \cdot \nu}\right)^{1/3} \cdot \theta, \tag{1}$$

where ω is pulsation, *R* convex object radius, θ angular distance of creeping path of a diffraction (creeping) ray (Fig. 1), and *v* the speed of EM wave propagation.

The first range, defined by relation $\xi_d(\omega, R, \theta) \leq \xi_d^{\text{th}}$ [1, 3], relates, e.g., to the transmission problem of UWB pulse with a spectrum being in the range 0–10 GHz, with smaller values of θ and R (e.g., human being case [4]). The second range, defined by relation $\xi_d(\omega, R, \theta) > \xi_d^{\text{th}}$ [1, 3], concerns problems in which, e.g., a modulated UWB pulse, that travels the obstacle along greater distance, as in, e.g., radar and sensor scanarios. For the both formulas, the threshold value is given. Therefore this solution is not universal.

In this chapter we present the way (based on VF algorithm) for obtaining the universal, applicable for all "considered" UWB scenarios, closed form impulse response of an obstacles cascade. We are interested in the scenarios of channel with human being presence. Therefore we will present in the chapter values of scenario parameters that are applicable to the case of UWB EM wave diffraction by convex objects which can be used as models of human body [4].

The rest of the chapter is organized as follows. In Sect. 2 we introduce the concept of UTD transfer function for a creeping ray. In Sect. 3 the procedure of obtaining the closed form universal impulse of a creeping ray is described. Next, the impulse response of a cascade of convex objects is presented. In Sect. 4 illustrative examples are shown. Section 5 concludes the chapter.

2 The UTD Transfer Function of a Creeping Ray

The considered creeping ray scenario is shown in Fig. 1. There are first, second, and N^{th} convex obstacles in Fig. 1 in the form of 2D conducting cylinders [2].

The transmitting and receiving antennas are placed at the points T_A and R_A , respectively. The attachment points and shedding points are marked with Q_n' and Q_n , respectively (n = 1, 2, ..., N). The main parameters of the scenario are as follows: R_n , the radius of the n^{th} cylinder, and θ_n , the n^{th} angular arc length that ray creeps (the n^{th} creeping arc length). For the purpose of the presentation of formulas that will follow, the distances along which propagates EM wave in the air are marked by s_k , which is the length of the ray path between object number k and object number k + 1 ($k = \{0, 1, 2, ..., N\}$) where object number 0 (for the case of k = 0) and N + 1 (for the case of k = N) relate to transmitting antenna and receiving antenna, respectively. The amplitude term of the diffraction coefficient (or transfer function) of a separate n^{th} convex object in a cascade is given in [1, 3, 5] by

$$H_{A(n)}(\omega) = H_{A1(n)}(\omega) + H_{A2(n)}(\omega), \qquad (2)$$

where

$$H_{A1(n)}(\omega) = \sqrt{\frac{v}{2\pi \cdot \omega \cdot \theta_n^2}} \cdot e^{-j\frac{\pi}{4}} \cdot F(X_{d(n)}), \qquad (3)$$

$$H_{A2(n)}(\omega) = -\left(\frac{2 \cdot v \cdot R_n^2}{\omega}\right)^{1/6} e^{-j\frac{\pi}{4}} \cdot \begin{cases} p^*\left(\xi_{d(n)}\right) \\ q^*\left(\xi_{d(n)}\right) \end{cases},\tag{4}$$

$$X_{d(n)} = \frac{\omega \cdot L_{d(n)} \cdot \theta_n^2}{2 \cdot \nu},\tag{5}$$

$$L_{d(n)} = \frac{s_{n-1} \cdot s_n}{s_{n-1} + s_n},\tag{6}$$

$$\xi_{d(n)} = \left(\frac{\omega \cdot R_n}{2 \cdot \nu}\right)^{1/3} \cdot \theta_n.$$
(7)

Formula (2) contains two components. The first contains transition zone function— $F(X_d)$ [2]. The second comprises Fock scattering function [2], $p^*(\xi_d)$ for soft (TM) polarization case or $q^*(\xi_d)$ for hard (TE) polarization case. The derivations which will follow do not depend on the polarization of EM wave. Therefore we choose that an electric field of propagating wave is tangential to the cylinders, and consequently we will use only $p^*(\xi_d)$ in Eq. (4). It can be shown that when slope diffraction on the last convex object is not taken into account (the receiving antenna is sufficiently far from the last convex object), Fourier transforms of an electric field at the output of the transmitting antenna, $E_{Ta}(\omega)$, and at the input of the receiving antenna, $E_{Ra}(\omega)$, for one creeping ray (Fig. 1), are related by the following expression:

$$E_{\mathrm{Ra}}(\omega) = E_{\mathrm{Ta}}(\omega) \cdot H^{(N)}(\omega) \cdot \exp\left(-j \cdot \frac{\omega}{\nu} \cdot s_p\right) \cdot A_{\mathrm{c}}(s_0, s_1, \dots s_{N+1}), \qquad (8)$$

where s_p is the total length of the creeping ray, $A_c(s_0, s_1, \ldots, s_N)$ is a product of all the spreading factors [2, 5] of a scenario, and $H^{(N)}(\omega)$ is a transfer function of a ray that creeps on a cascade of N (N > 1) convex objects. Transfer function $H^{(N)}(\omega)$ doesn't include any delay factor. For the purpose of the presentation of formula for $H^{(N)}(\omega)$, we use the following indications. The upper index N is the number of convex obstacles in a cascade. The lower index n is dedicated for functions or arguments determined for the n^{th} convex object in a cascade; see Fig. 1 and formulas (2), (3), (4), (5), (6), and (7). Beside $H_{A(n)}(\omega)$, we use also functions $H_{Z(n)}(\omega, s_n)$ and $H_{D(n)}(\omega, s_n)$ in order to determine $H^{(N)}(\omega)$. Functions $H_{A(n)}(\omega)$, $H_{Z(n)}(\omega, s_n)$, and $H_{D(n)}(\omega, s_n)$ include also parameters θ_n , R_n , $L_{d(n)}$, and $L_{z(n)}$, which are hidden for clarity, but will be used when it will be necessary. Parameters $L_{d(n)}$ and $L_{z(n)}$ are the separation coefficients [2, 5]. The first of them is used in $H_{A(n)}(\omega)$ (amplitude term), while the second of them is dedicated for components relating to slope of EM field, $H_{Z(n)}(\omega, s_n)$ and $H_{D(n)}(\omega, s_n)$. Argument $L_{d(n)}$ is given by Eq. (6) and $L_{Z(n)}$ has the form $L_{Z(n)} = [(s_{n-1} + s_n)/s_{n-1}]^{-1/3} \cdot s_n$. With the above assumptions, $H^{(N)}(\omega)$ can be written in the recursive form

$$H^{(N)}(\omega) = H^{(N-1)}(\omega) \cdot H_{A(N)}(\omega) + H_M^{(N-1)}(\omega) \cdot \frac{s_N \cdot v}{j \cdot \omega} H_{Z(N)}(\omega, s_N), \quad (9)$$

where

$$H_{M}^{(N-1)}(\omega) = H^{(N-2)}(\omega) \cdot H_{Z(N-1)}(\omega, s_{N-1}) + H_{M}^{(N-2)}(\omega) \cdot H_{D(N-1)}(\omega, s_{N-1}),$$
(10)

$$H_{Z(N)}(\omega,\theta_N,s_N) = \frac{1}{s_N} \cdot \frac{dH_{A(N)}(\omega,\theta_N)}{d\theta_N},$$
(11)

$$H_{D(N)}(\omega,\theta_N,s_N) = \frac{v}{j\cdot\omega} \cdot \frac{dH_{Z(N)}(\omega,\theta_N,s_N)}{d\theta_N}.$$
 (12)

In order to initialize the above recursive procedure, the following starting functions must be applied: $H^{(0)}(\omega) = 1$, $H_M^{(0)}(\omega) = 0$.

3 Derivation of an Impulse Response of a Multiple Diffraction Ray

3.1 The Rational Approximation of Transfer Functions in Frequency Domain Using VF Algorithm

The procedure of derivation of the new universal impulse response of a multiple diffraction ray is based on VF approximation in a frequency domain [6]. Such an approach is in principle always possible [1], but with each change of channel parameters (scenario), the new approximation is required.

Our aim is to transform functions $H_{A(n)}(\omega)$, $H_{Z(n)}(\omega, s_n)$, and $H_{D(n)}(\omega, s_n)$ to such a form, which allows for approximation regardless of changing parameters (of course within reasonable bounds). Consequently the approximation can be done only once and used in cases of change of channel parameters in a predetermined range.

In order to be able to make that universal approximation, we introduce the two new normalized variables:

$$X_{d(n)}^{\text{sub}} = \frac{\omega \cdot L_{d/Z(n)} \cdot \theta_n^2}{2\nu},\tag{13}$$

$$\xi_{d(n)}^{\rm sub} = \frac{\omega \cdot R_n \cdot \theta_n^{\ 3}}{2\nu},\tag{14}$$

where $L_{d/Z(n)}$ is $L_{d(n)}$ in $H_{A(n)}(\omega)$ or $L_{Z(n)}$ in $H_{Z(n)}(\omega, s_n)$ and $H_{D(n)}(\omega, s_n)$.

Functions $H_{A(n)}(\omega)$, $H_{Z(n)}(\omega, s_n)$, and $H_{D(n)}(\omega, s_n)$, when decomposing $H_{Z(n)}(\omega, s_n)$ and $H_{D(n)}(\omega, s_n)$ into two components as in $H_{A(n)}(\omega)$ case, can be rearranged with respect to the new variables into the following forms:

$$H_{A1(n)}(\omega) = \sqrt{\frac{L_{d(n)}}{4\pi}} \cdot e^{-j\frac{\pi}{4}} \frac{F\left(X_{d(n)}^{\text{sub}}\right)}{\sqrt{X_{d(n)}^{\text{sub}}}} = \sqrt{\frac{L_{d(n)}}{4\pi}} \cdot V_1\left(X_{d(n)}^{\text{sub}}\right), \tag{15}$$

$$H_{A2(n)}(\omega) = \sqrt{R_n \theta_n} \cdot \left[-e^{-j\frac{\pi}{4}} \frac{P_*\left(\xi_{d(n)}^{\rm sub} \, 1/3\right)}{\sqrt{\xi_{d(n)}^{\rm sub} \, 1/3}} \right] = \sqrt{R_n \theta_n} \cdot V_{A2}\left(\xi_{d(n)}^{\rm sub}\right), \quad (16)$$

$$H_{Z1(n)}(\omega, s_n) = \frac{1}{s_n} \frac{\partial}{\partial \theta_n} \left[H_{A1(n)} \left(X_{d(n)}^{\text{sub}} \right) \right]$$
$$= -\frac{1}{s_n} e^{j\pi/4} \sqrt{\frac{\omega \cdot L_{Z(n)}^2}{2\nu\pi}} + \frac{j\omega}{s_n} \sqrt{\frac{L_{Z(n)}^3}{\pi}} \frac{\theta_n}{2\nu} V_1 \left(X_{d(n)}^{\text{sub}} \right), \tag{17}$$

$$H_{Z2(n)}(\omega, s_n) = \frac{\sqrt{R_n}}{s_n} \frac{\partial}{\partial \theta_n} \left\{ \sqrt{\theta_n} \left[-e^{-j\frac{\pi}{4}} \frac{p_*\left(\xi_{d(n)}^{\mathrm{sub}\ 1/3}\right)}{\sqrt{\xi_{d(n)}^{\mathrm{sub}\ 1/3}}} \right] \right\}$$
$$= \frac{1}{s_n} \sqrt{\frac{R_n}{\theta_n}} \cdot \left(-e^{-j\frac{\pi}{4}} \right) \cdot \left[p_*\left(\xi_{d(n)}^{\mathrm{sub}\ 1/3}\right) \right]' \xi_{d(n)}^{\mathrm{sub}\ 1/6}$$
$$= \frac{1}{s_n} \sqrt{\frac{R_n}{\theta_n}} \cdot V_{Z2}\left(\xi_{d(n)}^{\mathrm{sub}}\right), \tag{18}$$

$$H_{D1(n)}(\omega, s_n) = \frac{\nu}{j\omega} \frac{\partial}{\partial \theta_n} \left[H_{Z1(n)} \left(X_{d(n)}^{\text{sub}}, s_n \right) \right]$$

$$= -\frac{1}{s_n} e^{-j\pi/4} \sqrt{\frac{\omega \cdot L_{Z(n)}^4 \cdot \theta_n^2}{2\nu \cdot \pi}} + \frac{j\omega}{s_n} \sqrt{\frac{L_{Z(n)}^5}{\pi}} \frac{\theta_n^2}{2\nu} V_1 \left(X_{d(n)}^{\text{sub}} \right)$$

$$+ \sqrt{\frac{L_{Z(n)}^3}{4\pi}} V_1 \left(X_{d(n)}^{\text{sub}} \right), \qquad (19)$$

$$H_{D2(n)}(\omega, s_n) = \frac{\sqrt{R_n}}{s_n} \frac{v}{j\omega} \frac{\partial}{\partial \theta_n} \left\{ \frac{1}{\sqrt{\theta_n}} \left[-e^{-j\frac{\pi}{4}} \left[p_* \left(\xi_{d(n)}^{\text{sub} 1/3} \right) \right]' \xi_{d(n)}^{\text{sub} 1/6} \right] \right\}$$
$$= \frac{1}{s_n} \sqrt{\frac{R_n}{\theta_n^3}} \frac{v}{j\omega} \left\{ -e^{-j\frac{\pi}{4}} \left[p_* \left(\xi_{d(n)}^{\text{sub} 1/3} \right) \right]'' \sqrt{\xi_{d(n)}^{\text{sub}}} \right\}$$
$$= \frac{1}{s_n} \sqrt{\frac{R_n}{\theta_n^3}} \frac{v}{j\omega} V_{2D} \left(\xi_{d(n)}^{\text{sub}} \right).$$
(20)

Signs ()' and ()" in Eqs. (18) and (20) means the first- and second-order derivative, respectively.

These new forms of $H_{A(n)}(\omega)$, $H_{Z(n)}(\omega, s_n)$ and $H_{D(n)}(\omega, s_n)$ let us apply VF program in an effective way and consequently straight application of inverse Laplace transform in order to find the corresponding time-domain functions.

The functions which we approximate in Eqs. (15), (16), (17), (18), (19), and (20) by VF algorithm are $V_1(X_{d(n)}^{\text{sub}})$ and $V_{A2}(\xi_{d(n)}^{\text{sub}})$, $V_{Z2}(\xi_{d(n)}^{\text{sub}})$, and $V_{D2}(\xi_{d(n)}^{\text{sub}})$.

In order to apply VF approximation, we must determine ranges of values for both new variables. These ranges should follow the values of the UWB channel parameters that can be met in real life. In our case we focus on convex objects which can model human being presence in an UWB channel. These objects can be cylinders with radius 0.2–0.3 m, e.g., [4], and we use this range of *R* value in calculation of ranges of values of new variables. The rest of the parameters whose limits must be determined are frequency, *f*; angular distance that creeps a diffraction ray, θ ; and the separation coefficients $L_{d(n)}$ and $L_{Z(n)}$. We assume that $0.5 \le f \le 10$ (GHz) (incident UWB pulse spectrum limits), $10^{-4} \le \theta \le \pi$ (rad) and both values of separation coefficients $L_{d(n)}$ and $L_{Z(n)}$ in the range from 0.5 to 5 (m).

With the assumed above bounds for UWB scenario parameters, the bounds of the new approximation variables are the following: $10^{-8} \le X_{d(n)}^{sub} \le 10^3$, $10^{-12} \le \xi_{d(n)}^{sub} \le 10^3$. The log-scale sampling of approximation domain gives the best results of rational approximation (VF) of functions: $V_1(X_{d(n)}^{sub})$, $V_{A2}(\xi_{d(n)}^{sub})$, $V_{Z2}(\xi_{d(n)}^{sub})$, and $V_{D2}(\xi_{d(n)}^{sub})$. To keep the relative deviation of rational approximations under 1 % in the whole calculated range of values of $X_{d(n)}^{sub}$ and $\xi_{d(n)}^{sub}$, we used 10 poles for approximation of $V_1(X_{d(n)}^{sub})$ and 20 poles for approximation of $V_{A2}(\xi_{d(n)}^{sub})$, $V_{Z2}(\xi_{d(n)}^{sub})$, and $V_{D2}(\xi_{d(n)}^{sub})$. After approximating $V_1(X_{d(n)}^{sub})$ and $V_{A2}(\xi_{d(n)}^{sub})$, $V_{Z2}(\xi_{d(n)}^{sub})$, and $V_{D2}(\xi_{d(n)}^{sub})$. (13), (14), (15), (16), (17), (18), (19), and (20), with the usage of J = 10 and K = 20 poles $(A_{1j}, A_{A2k}, A_{Z2k}, A_{D2k})$ and residues $(C_{1j}, C_{A2k}, C_{Z2k}, C_{D2k})$, take the forms

$$H_{A1(n)}(\omega) \approx \sqrt{\frac{L_{d(n)}}{4\pi}} \cdot \sum_{j=1}^{J} \frac{C_{1j} \cdot 2\nu/L_{d(n)}\theta_n^2}{j\omega + A_{1j} \cdot 2\nu/L_{d(n)}\theta_n^2} = \sqrt{\frac{L_{d(n)}}{4\pi}} \cdot \sum_{j=1}^{J} P_{1j}(\omega), \quad (21)$$

$$H_{A2(n)}(\omega) \approx \sqrt{R_n \theta_n} \cdot \sum_{k=1}^K \frac{C_{A2k} \cdot 2\nu/R_n \theta_n^3}{j\omega + A_{A2k} \cdot 2\nu/R_n \theta_n^3},$$
(22)

$$H_{Z1(n)}(\omega, s_n) \approx -\frac{1}{s_n} e^{j \cdot \pi/4} \sqrt{\frac{\omega \cdot L_{Z(n)}^2}{2\pi \cdot \nu} + \frac{j\omega}{s_n}} \sqrt{\frac{L_{Z(n)}^3}{\pi} \frac{\theta_n}{2\nu}} \sum_{j=1}^J P_{1j}(\omega),$$
(23)

$$H_{Z2(n)}(\omega, s_n) \approx \frac{1}{s_n} \sqrt{\frac{R_n}{\theta_n}} \sum_{k=1}^K \frac{C_{Z2k} \cdot 2\nu/R_n \theta_n^3}{j\omega + A_{Z2k} \cdot 2\nu/R_n \theta_n^3},$$
(24)

$$H_{D1(n)}(\omega, s_n) \approx -\frac{1}{s_n} e^{j \cdot \pi/4} \sqrt{\frac{\omega \cdot L_{Z(n)}{}^4 \theta_n^2}{2\pi \cdot \nu}} + \frac{j\omega}{s_n} \sqrt{\frac{L_{Z(n)}{}^5}{\pi}} \frac{\theta_n^2}{2\nu} \sum_{j=1}^J P_{1j}(\omega) + \frac{1}{s_n} \sqrt{\frac{L_{Z(n)}{}^3}{4\pi}} \sum_{j=1}^J P_{1j}(\omega),$$
(25)
$$H_{D2(n)}(\omega, s_n) \approx \frac{1}{s_n} \sqrt{\frac{R_n}{\theta_n^3}} \frac{v}{j\omega} \sum_{k=1}^K \frac{C_{D2k} \cdot 2v/R_n \theta_n^3}{j\omega + A_{D2k} \cdot 2v/R_n \theta_n^3}.$$
 (26)

3.2 Impulse Responses Derivation

In the previous subsection, we obtained rational approximations of the components of transfer function of a multiple diffraction ray. The corresponding time-domain functions of the components in Eqs. (21), (22), (23), (24), (25), and (26) can be easily found by an application of inverse Laplace transform and one-sided inverse Fourier transform as in [1, 3].

The components in the sums in Eqs. (21), (22), (23), (24), (25), and (26) are the functions proportional to $(s + a)^{-1}|_{s=j\omega}$. Their inverse Laplace transforms are the functions proportional to e^{-at} . The factors $j\omega$ and $j\omega^{-1}$ are transformed to derivative and integral, respectively, in time domain.

The components proportional to the fractional powers of ω can be retransformed with inverse one-sided Fourier transformation. In a consequence the time-domain equivalents of Eqs. (21), (22), (23), (24), (25), and (26) have the forms

$$h_{A1(n)}(t) \approx \sqrt{\frac{L_{d(n)}}{4\pi}} \cdot \sum_{j=1}^{J} \frac{C_{1j} 2\nu}{L_{d(n)} \theta_n^2} e^{-\frac{A_{1j} 2\nu}{L_{d(n)} \theta_n^2} t} = \sqrt{\frac{L_{d(n)}}{4\pi}} \cdot \sum_{j=1}^{J} p_{1j}(t),$$
(27)

$$h_{A2(n)}(t) \approx \sqrt{R_n \theta_n} \cdot \sum_{k=1}^K \frac{C_{A2k} 2\nu}{R_n \theta_n^3} e^{-\frac{A_{A2k} 2\nu}{R_n \theta_n^3}t},$$
(28)

$$h_{Z1(n)}(t, s_n) \approx -\frac{1}{s_n} \sqrt{\frac{L_{Z(n)}^2}{2v(\pi \cdot t)^3}} \Gamma\left(\frac{3}{2}\right) + \frac{1}{s_n} \sqrt{\frac{L_{Z(n)}^3}{\pi}} \\ \times \frac{\theta_n}{2v} \sum_{j=1}^J \left(\delta(t) - \frac{A_{1j}}{L_{Z(n)}\theta_n^2} p_{1j}(t)\right),$$
(29)

$$h_{Z2(n)}(t,s_n) \approx \frac{1}{s_n} \sqrt{\frac{R_n}{\theta_n}} \sum_{k=1}^K \frac{C_{Z2k} 2\nu}{R_n \theta_n^3} e^{\frac{-A_{Z2k} - 2\nu}{R_n \theta_n^3}t},$$
(30)

$$h_{D1(n)}(t,s_n) \approx -\frac{1}{s_n} \sqrt{\frac{L_{Z(n)}^4 \theta_n^2}{2v(\pi \cdot t)^3}} \Gamma\left(\frac{3}{2}\right) + \frac{1}{s_n} \sqrt{\frac{L_{Z(n)}^5}{\pi}} \times \frac{\theta_n^2}{2v} \sum_{j=1}^J \left(\delta(t) - \frac{A_{1j}}{L_{Z(n)} \theta_n^2} p_{1j}(t)\right) + \frac{1}{s_n} \sqrt{\frac{L_{Z(n)}^3}{4\pi}} \sum_{j=1}^J p_{1j}(t), \quad (31)$$

The New Vector Fitting Approach to Multiple Convex Obstacles Modeling...

$$h_{D2(n)}(t,s_n) \approx \frac{1}{s_n} \sqrt{\frac{R_n^n}{\theta_n^3}} \frac{v}{j\omega} \sum_{k=1}^K \frac{C_{D2k} 2v}{R_n \theta_n^3} e^{-\frac{A_{D2k} 2v}{R_n \theta_n^3}t}.$$
(32)

Now the impulse response of a multiple diffraction ray, creeping on N convex obstacles, analogous in form to the transfer function of a multiple diffraction ray (Eq. 8), can be determined:

$$h^{(N)}(\omega) = h^{(N-1)}(t) * h_{A(N)}(t) + h_M^{(N-1)}(t) \cdot s_N \cdot v * \int h_{Z(N)}(t, s_N) dt, \qquad (33)$$

where

$$h_M^{(N-1)}(t) = h^{(N-2)}(t) * h_{Z(N-1)}(t, s_N) + h_M^{(N-2)}(t) * h_{D(N-1)}(t, s_N)$$
(34)

and $h_{A(n)}(t) = h_{A1(n)}(t) + h_{A2(n)}(t)$, $h_{Z(n)}(t,s_n) = h_{Z1(n)}(t,s_n) + h_{Z2(n)}(t,s_n)$, $h_{D(n)}(t,s_n) = h_{D1(n)}(t,s_n) + h_{D2(n)}(t,s_n)$, while the initial functions for the recursive procedure in Eq. (33) are $h^{(0)}(t) = 1 h_M^{(0)}(t) = 0$.

4 Numerical Verification of the Impulse Response

In this section we verify the new impulse response by comparing direct timedomain calculation results with IFFT results. We did it for the following examples. In the first the antennas are shadowed by a cascade of three convex objects. The lengths of the ray paths between the consecutive convex objects are equal to 1 m. The creeping arc lengths are also the same and equal to 0.1 rad. The distances from a transmitting antenna point to an attachment point on the first object and from a shedding point on the third object to a receiving antenna point are equal to 5 m. The radius of each object is set to 0.25 m. In the second scenario diffraction ray creeps one obstacle but on longer arc length which is equal to π rad. The radius of an obstacle is also 0.25 m. The results calculated for the first and second example are presented in Fig. 2a, b, respectively.

There are three graphs on each figure. One of them is an incident Gauss UWB pulse shape of 3 ns duration. The other two are distorted pulse shapes calculated directly in the time domain and with IFFT applied for original transfer functions (9). The figures show that there is a very good agreement between approximated (VF) and original (IFFT) output waveforms. The maximum values of the incident pulse are normalized to the maximum values of the distorted pulses. The delay factors and attenuation in the air are not taken into account in calculation of results presented in the figures.



+++ Distorted UWB pulse shape - TD

Fig. 2 The shape of the incident UWB Gauss pulse compared with the shape of the distorted pulse derived with time-domain calculations and with IFFT caused by a diffraction on (**a**) three convex objects with $R_1 = R_2 = R_3 = 0.25$ m, $\theta_1 = \theta_2 = \theta_3 = 0.1$ rad, $s_0 = 5$ m, $s_1 = s_2 = 1$ m, $s_3 = 5$ m and (**b**) one convex object with R = 0.25 m, $\theta = \pi$ rad, $s_0 = s_1 = 5$ m

5 Conclusions

In the chapter we presented the new effective and universal approach to modeling the UWB channel containing convex obstacles. The approach is based on VF algorithm in frequency domain. The main idea was to introduce normalized variables $X_{d(n)}^{\text{sub}}$ (Eq. 13) and $\xi_{d(n)}^{\text{sub}}$ (Eq. 14), in respect of which the rational functions approximation has been performed. Therefore the impulse response takes the form independent from the current parameters of obstacles in the channel. The impulse response was positively verified by numerical calculations. It contains amplitude as well as slope diffraction terms. Exponential form of the impulse response components allows for time-domain calculation of the output signal by means of accelerating recursive algorithms or by the usage of spice models. It should be noted that when, e.g., one needs to increase the upper limit of *R* values from 0.3 m to R_{max} , then the available upper limit of UWB signal would have to be decreased in order to maintain the given numbers of poles.

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FDTD Analysis of Shielded High- T_c Microstrip Resonators on Anisotropic Substrates

M.L. Tounsi, O. Madani, and M.C.E. Yagoub

Abstract An improved FDTD hybrid-mode technique is proposed to analyze shielded high-temperature T_c superconducting (HTS) microstrip resonators of arbitrary thickness. The temperature dependence on resonant frequency and bandwidth are investigated assuming an anisotropic dielectric substrate. Variations of resonant frequency with high- T_c superconducting thickness and shielding effect are also presented as well as *S* parameters over wideband.

Keywords Microstrip resonator • Superconductor • FDTD method • Anisotropy • Shielding • UWB

1 Introduction

High-temperature superconducting (HTS) passive microwave devices have shown significant preeminence over related devices fabricated with normal conductors such as gold, silver, or copper. Advantages of using HTS materials in passive microwave circuits include very small losses, low attenuation, and low distortion and dispersion, especially in the design of transmission lines, interconnects, microwave filters, and highly efficient antennas [1–3]. Therefore, owing to the advances in material technology and the ever-growing interest in high-frequency bands,

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microwave circuits with anisotropic media have been rapidly gaining importance in microwave and millimeter-wave industry [4–6].

UWB technology is among many promising modern applications [7], making UWB resonators essential components in UWB wireless and radar systems. The rapid growth in this field has prompted the development of various types of UWB filters based on microstrip resonators [8, 9].

Many studies have been published in the literature about passive microwave circuits like microstrip resonators and filters, but most of them assume perfect conductors. In fact, only few works have been done using superconductors, especially with the finite-difference time domain method (FDTD). To the best of our knowledge, the study of HTS microstrip resonators using FDTD method was only carried out by Okazaki et al. [10], using a nonuniform FDTD to resolve the very thin thickness of the superconductor.

In this chapter, we present an FDTD analysis of a shielded HTS microstrip resonator on anisotropic substrate which yields good agreement with published data [11]. The analyzed resonator can also operate within the ultra-wideband (UWB) simply by adjusting its dimensions. An improved formulation is used rather than a nonuniform FDTD to resolve the very thin thickness of the superconductor, assumed to be arbitrary.

2 Numerical Formulation

In the FDTD method, Maxwell's equations are solved by a difference method based on Yee's notation [12]. If a superconducting material is included in the simulation of a microwave circuit, the two-fluid model can be used to describe the superconductivity with a complex conductivity [13]:

$$\sigma = \sigma_1 - j\sigma_2 \tag{1}$$

with

$$\sigma_1 = \sigma_n \left(\frac{T}{T_c}\right)^4,\tag{2}$$

$$\sigma_2 = \frac{1}{\omega\mu_0(\lambda_L(T))^2},\tag{3}$$

where

$$\lambda_L(T) = \frac{\lambda_0}{\sqrt{1 - \left(\frac{T}{T_c}\right)^4}} \tag{4}$$

and where σ_n , $\lambda_L(0)$, ω , and T_c are the normal conductivity near the critical temperature, the zero temperature penetration depth, the pulsation, and the critical temperature of the superconductor, respectively.

So, the Maxwell-Ampere equation can be modified as

$$\vec{\nabla} \times \vec{H} = j\omega\varepsilon\vec{E} + \sigma_1\vec{E} + \frac{1}{j\omega\lambda_L^2(T)\mu_0}\vec{E},$$
(5)

where μ is the permeability of the medium and ε the permittivity that can be represented by the following matrix:

$$\varepsilon = \varepsilon_0 \begin{bmatrix} \varepsilon_x & 0 & 0\\ 0 & \varepsilon_y & 0\\ 0 & 0 & \varepsilon_z \end{bmatrix}$$
(6)

with ε_0 the permittivity of the free space. In time domain (5), can be formulated as

$$\vec{\nabla} \times \vec{H} = \varepsilon \frac{\partial \vec{E}}{\partial t} + \sigma_1 \vec{E} + \frac{1}{\lambda_L^2(T)\mu_0} \int \vec{E} \cdot dt.$$
(7)

Using central finite differences, the updated equations at $t = (n + 1)\Delta t$ for the electric fields at point (i, j, k) can be obtained as

$$E_{x,y,z}^{n+1}(i,j,k) = \left[\frac{1 - \frac{\sigma_{1}\Delta t}{2\epsilon_{x,y,z}}}{1 + \frac{\sigma_{1}\Delta t}{2\epsilon_{x,y,z}}}\right] E_{x,y,z}^{n}(i,j,k) + \left[\frac{\frac{\Delta t}{\epsilon_{x,y,z}}}{1 + \frac{\sigma_{1}\Delta t}{2\epsilon_{x,y,z}}}\right] \cdot \nabla \times H^{n+\frac{1}{2}} - \left[\frac{\frac{\epsilon_{0}}{\epsilon_{x,y,z}}\left(\frac{c\Delta t}{\lambda_{L}(T)}\right)}{1 + \frac{\sigma_{1}\Delta t}{2\epsilon_{x,y,z}}}\right] \sum_{p=0}^{n} E_{x,y,z}^{p}(i,j,k).$$

$$(8)$$

In the above equations, *c* is the velocity of light, and ε_x , ε_y , and ε_z are the permittivities in *x*, *y*, and *z* directions, respectively. Δx , Δy , and Δz are the respective spatial steps in the *x*, *y*, and *z* directions and Δt the time step which must satisfy the stability criterion [12]:

$$\Delta t \le \frac{1}{\nu_{\max}\sqrt{\frac{1}{\Delta x^2} + \frac{1}{\Delta y^2} + \frac{1}{\Delta z^2}}}\tag{9}$$

with v_{max} the maximum velocity in the computational volume.

3 Modeling the Thickness of the Superconductor

The very thin thickness of the superconductor requires a special treatment in FDTD method. Usually, a nonuniform mesh is used which can be very time consuming. In our case, we adopted the model of Maloney and Smith [14] which was initially developed for conductors with normal thickness. When the thickness *t* of the superconductor is very small with regard to the mesh height Δz , we can define a new cell similar to the one of the classical FDTD, but containing an additional field E_{zm} in the strip of superconductor (Fig. 1).

In this new mesh, the calculation of the electric fields E_x and E_y requires the multiplication of the conductivity σ of the superconductor by a geometrical factor that results from integrating the analytical current density over the strip. This factor, which takes into account the field attenuation in the superconductor [15], is defined as

$$m = \frac{\lambda_L(T) \left(1 - e^{-t/\lambda_L(T)}\right)}{\Delta z},$$
(10)

where *t* is the thickness of the superconductor assumed to be arbitrary, λ_L is the penetration depth, and *T* is the temperature.

By multiplying σ by *m*, a new updated equation for the superconductor can be obtained for the tangential field components (E_x and E_y), including the exponential decrease of the field inside the superconductor.

$$E_{x,y}^{n+1}(i,j,k) = \left[\frac{1 - \frac{m\sigma_1\Delta t}{2\varepsilon_{x,y}}}{1 + \frac{m\sigma_1\Delta t}{2\varepsilon_{x,y}}}\right] E_{x,y}^n(i,j,k) + \left[\frac{\frac{\Delta t}{\varepsilon_{x,y}}}{1 + \frac{m\sigma_1\Delta t}{2\varepsilon_{x,y}}}\right] \cdot \nabla \times H^{n+\frac{1}{2}} - \left[\frac{m\frac{\varepsilon_0}{\varepsilon_{x,y}}\left(\frac{c\Delta t}{\lambda_L(T)}\right)}{1 + \frac{m\sigma_1\Delta t}{2\varepsilon_{x,y}}}\right] \sum_{p=0}^n E_{x,y}^p(i,j,k).$$
(11)



Fig. 1 An elementary cell with taking into account the thickness of the superconductor

Note that E_{zm} is evaluated via Eq. (8). Also to calculate H_x and H_y , both E_z and E_{zm} must be used multiplied by $(1-t/\Delta z)$ and $t/\Delta z$, respectively. For the other field components (E_z , H_x , H_y , H_z), the classical FDTD updated equations can be used.

4 Geometry, Excitation Source, and Boundary Conditions

Figure 2 shows the simulated HTS resonator. The top layer is a microstrip pattern of a superconducting thin film YBCO. Its thickness is denoted by *t*, while T_c , λ_0 , and σ_n are set to 88 K, 0.16 µm, and 3.10⁶ S/µm, respectively, as given in [11]. The middle layer is a lossless r-cut sapphire substrate ($\varepsilon_x = \varepsilon_y = 10.18$, $\varepsilon_z = 10.03$, $\varepsilon_{xy} = 0.78$, $\varepsilon_{yz} = \varepsilon_{xz} = -0.7$). The bottom layer is a ground plane.

At t = 0, all EM field components are assumed to be zero in the computational volume. To excite the resonator, a Gaussian pulse excitation is used since its amplitude has a smooth variation. Also, its frequency spectrum can provide frequency-domain information from dc to the desired cutoff frequency by adjusting the width of the pulse. To introduce this pulse in the FDTD grid, several techniques can be used. A simple and efficient way is to introduce the pulse directly in the FDTD algorithm through the E_z component, using an additional term in Maxwell's equations [16].



Fig. 2 3D view of the simulated HTS resonator ($\Delta x = 0.1 \text{ mm}$, $\Delta y = 0.1173 \text{ mm}$, $\Delta z = 0.125 \text{ mm}$, and $\Delta t = 0.1668 \text{ ps}$)

As shown in Fig. 2, first-order absorbing conditions [17] are applied in order to eliminate the influence of the EM wave reflected by the walls, while perfect electric conductor (PEC) conditions were applied otherwise, especially on the ground plane.

5 Numerical Results

To validate the proposed FDTD numerical approach, we first calculated the *S* parameters of the above HTS resonator at T = 65 K. Figure 3 shows two different resonant frequencies where the energy is maximum.

We further investigated the effect of temperature on the first resonant frequency for an r-cut sapphire substrate. As shown in Fig. 4, good agreement has been obtained with [11] since the relative error does not exceed 1 %. It can be also shown that increasing the temperature will decrease the resonant frequency. This decrease, caused by the kinetic inductance of the superconducting electrons [18], is significant for temperatures near the critical temperature ($T_c = 88$ K). Note that in our simulation, we have used $\varepsilon_x = \varepsilon_y = 10.18$, $\varepsilon_z = 10.03$ for r-cut sapphire with $\varepsilon_{xy} = \varepsilon_{yz} = \varepsilon_{xz} = 0$, but the results agree well since the extra-diagonal elements of the tensor ε given in [11] are of low value compared to the diagonal ones.

Figure 5 illustrates the influence of the temperature on the bandwidth for an HTS YBCO HTS resonator for three types of anisotropic substrates. It can be observed



Fig. 3 *S* parameters of the HTS resonator on r-cut sapphire substrate at T = 65 K (w = 0.5 mm, G = 0.352 mm, L = 5.78 mm, h = 5 mm, a = 5.1 mm, b = 4h, t = 0.2 µm, $\varepsilon_x = \varepsilon_y = 10.18$, $\varepsilon_z = 10.03$)



that this effect is significant only for temperatures near the critical temperature where most of normal electrons are transformed into superconducting electrons. Note also that a low degree of anisotropy increases the bandwidth outside the critical temperature T_c , while it increases rapidly near T_c .

Thus, the thickness effect of the YBCO film was investigated for the same anisotropic substrates. It can be seen from Fig. 6 (for T = 50 K) that the thickness t of the superconducting film increases with the resonant frequency. Note that this increase is more important for small values of t ($t < \lambda_0$). When t exceeds λ_0 , increasing the superconducting film thickness will increase slowly the resonant frequency. Note also that high degree of anisotropy allows decreasing the resonant frequency.

Finally, in Fig. 7, the shielding effect on the resonant frequency is presented showing a decrease when the top cover height of the box *b* increases. This decrease is more important for small values of b (b = 2h and b = 3h). When *b* exceeds about 5*h*, almost no variation in the resonant frequencies is observed.



6 Conclusion

A hybrid-mode FDTD technique was presented for the analysis of shielded HTS microstrip resonators on anisotropic substrates. An improved formulation was used to take into account any arbitrary thickness of the superconductor without meshing its strip cross section. Numerical results for the effect of the temperature on resonant frequency and bandwidth have been presented. Variations of the resonant frequency versus the thickness of the superconducting film as well as the shielding effects have been also presented. Numerical results are in good agreement with those available in the literature.

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Parametric Evaluation of Absorption Losses and Comparison of Numerical Results to Boeing 707 Aircraft Experimental HIRF Results

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Abstract A broadband (100 MHz–1.2 GHz) plane wave electric field source was used to evaluate electric field penetration inside a simplified Boeing 707 aircraft model with a finite-difference time-domain (FDTD) method using EMA3D. The role of absorption losses inside the simplified aircraft was investigated. It was found that, in this frequency range, none of the cavities inside the Boeing 707 model are truly reverberant when frequency stirring is applied, and a purely statistical electromagnetics approach cannot be used to predict or analyze the field penetration or shielding effectiveness (SE). Thus it was our goal to attempt to understand the nature of losses in such a quasi-statistical environment by adding various numbers of absorbing objects inside the simplified aircraft and evaluating the SE, decay-time constant τ , and quality factor Q. We then compare our numerical results with experimental results obtained by D. Mark Johnson et al. on a decommissioned Boeing 707 aircraft.

Keywords HIRF • FDTD • Aircraft • Losses • Shielding effectiveness

1 Introduction

The certification process for an aircraft is a lengthy and expensive process. The high-intensity radiated fields (HIRF) environment forms only one segment of the certification requirements. A computational electromagnetics (CEM) model can provide a more detailed view of the HIRF coupling processes and levels throughout the aircraft than is generally possible in testing due to time and cost constraints.

It is, however, not possible to obtain all the material losses inside an aircraft. Therefore, a technique needs to be developed where absorption losses for Low

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Level Swept Fields (LLSF) evaluation can be reasonably approximated and compared with experimental measurements.

In this work, we investigate absorption losses numerically using EMA3D [1], a finite-difference time-domain (FDTD)-based code. We start with a simplified model of an empty Boeing 707 aircraft and illuminate it with a broadband plane wave at broadside incidence, horizontal polarization. Shielding effectiveness (SE) is then evaluated in three aircraft cavities, namely, the cabin, cockpit, and avionics bay. We then start adding lossy materials, which consist of randomly spaced cubes with low conductivity, to investigate how the SE changes as a function of the number of absorbing bodies, in the frequency range from 100 MHz to 1.2 GHz. We also observe how the statistical properties of electric fields degrade as absorption losses are added. Additionally, decay-time constant τ and quality factor Q are extracted as a function of frequency and the number of absorbers. Numerical results are then compared with experimental data [2]. We note that there are various discrepancies between the simplified model and the aircraft itself, as well as some uncertainties, which will be described in the sections below. Therefore, we use experimental data here only for reference, and we do not attempt to match the modeling results to experimental results.

2 Simplified Aircraft Model Setup

A simplified model of an empty (no absorption losses) Boeing 707 aircraft [3] is presented in Fig. 1. There is no landing gear present (landing gear wells are isolated, however), no engines, no seams, or slots (e.g., between aircraft skin panels). The interior is mostly empty except for the floor, three bulkheads (two in the cockpit, one in the tail), and four avionics boxes in the avionics bay below the cockpit floor. The skin, floor, bulkheads, and avionics boxes are all metal.

There were some uncertainties about the Boeing 707 aircraft that need to be pointed out. From [2], there may be a metallic layer on the cockpit windshield that was not characterized and therefore not implemented into the model. All windows in the Boeing 707 model have the material property of air. The cabin seats of the aircraft were absent during testing, and all seats are absent in the model. The exact geometry of the aircraft during test was not well known, thus some discrepancies may arise between the aircraft and the model. The coupling mechanisms between the cockpit, cabin, and avionics bay were uncertain; however, significant coupling was observed during testing. Exact antenna and probe locations of the test were not described in the test report.

A broadband Gaussian plane wave electric field source was used for external excitation. The angle of incidence was 90° (broadside), with horizontal polarization. The excitation waveform is shown in Fig. 2a, while the frequency content of the waveform is shown in Fig. 2b. Thus the source covers a frequency range between 100 MHz and 1.2 GHz.



Fig. 1 Simplified Boeing 707 model with windows and windshield (*white*), bulkheads (*blue*), floor (*black*), and avionics boxes (*green*)



Fig. 2 Gaussian pulse source (a), normalized frequency content of the excitation source (b)

2.1 Numerical Data and Data Processing

For numerical results, a principal field component E_x was obtained at three representative locations inside the cockpit, six inside the cabin (three forward, three aft), and three inside the avionics bay. While other field components also exist inside the aircraft, only E_x is obtained for comparison with measurements. The antennas were not modeled. Additionally, to compare the numerical results with experimental data as presented in [2], 50 MHz frequency stirring was applied to raw numerical data. Thus only the 50 MHz frequency stirred results will be presented here for SE, timedecay constant τ , and quality factor Q.

After running and analyzing the empty Boeing 707 model, absorption losses are gradually added to the cockpit, cabin, and the avionics bay. Absorption losses are

Simulation number	No. absorbers cockpit	No. absorbers avionics bay	No. absorbers cabin
1	0	0	0
2	1	5	25
3	2	10	50
4	4	21	100
5	6	21	150

Table 1 Absorbing bodies in each aircraft cavity for five different simulations

represented in the model as semi-randomly spaced cubes with the side length s = 0.5 m and conductivity $\sigma = 0.01$ S/m. Another four simulations were completed with a different number of absorbing bodies each, as summarized in Table 1.

3 Numerical Results

3.1 Shielding Effectiveness

The SE values were obtained after Fourier-transforming time-domain data at each location, then 50 MHz bandwidth averaging the frequency domain data. Then data were location averaged inside each cavity (i.e., averaged three locations inside the cockpit, three locations in the avionics bay, and six locations in the cabin). This method of data processing was chosen so that the numerical results could be consistently compared with the experimental results from [2]. Using the power balance method [4], the SE can be defined as

$$SE = -10\log_{10}\left(\frac{E_c^2}{E_i^2}\right),\tag{1}$$

where E_{c} is the electric field inside the cavity and E_{i} is the incident electric field.

Since aperture losses are the only losses present in the empty Boeing 707 model, the SE is less than 5 dB above 500 MHz, 15–30 dB lower than what is measured experimentally [2]. As absorbing bodies are gradually added to the model, the SE gradually increases, eventually making absorption the dominant loss mechanism inside the aircraft. This is shown in Figs. 3a–c for the cockpit, avionics bay, and cabin, respectively.

3.2 Decay-Time Constant τ and Q

The quality factor Q, which is a measure of how well a cavity stores energy, can be derived from extracting the decay-time constant τ as a function of frequency for each aircraft cavity. To extract τ , we first take an electric field time-domain



Fig. 3 Shielding effectiveness for a different number of absorbing bodies inside the cockpit (**a**), avionics bay (**b**), and cabin (**c**) and compared with experimental results for each cavity [2]. From *bottom* to *top*, the trend is simulation 1, 2, 3, 4, and 5, with experimental data at the *top* of (**a**) and (**c**). In (**b**), experimental data overlays the *top* two simulation curves

waveform E(t) obtained from numerical simulations, Fourier transform it into the frequency domain and square it to obtain $E^2(f)$. Then a band-pass filter is applied to obtain 50 MHz bandwidth averaging, and an inverse Fourier transform is performed to obtain a 50 MHz bandwidth averaged power time-domain waveform. The resulting waveform is then fit with an exponential decay to obtain τ . This procedure is applied for all probes inside the aircraft, and τ is then probe averaged for each cavity. Then Q can be obtained directly from τ [4]:

$$Q = \omega \tau, \tag{2}$$

where ω is the angular frequency. The Qs are shown in Fig. 4 for a different number of absorbers inside each aircraft cavity and compared with experimental results [2].



Fig. 4 Q for a different number of absorbing bodies inside the cockpit (**a**), avionics bay (**b**), and cabin (**c**) and compared with experimental results for each cavity obtained using band-limited white Gaussian noise excitation [2]

3.3 Boeing 707 Cabin Further Analysis

Given that the passenger cabin is the largest cavity in the aircraft, and with fewest discrepancies between the model and the aircraft, we choose the cabin for more detailed analysis. Figure 5a shows how the cabin SE varies with the number of absorbers at three different frequencies, 150 MHz, 750 MHz, and 1.15 GHz, and Fig. 5b shows how the cabin Q varies with the number of absorbers for the same frequencies. For each frequency, SE increases with the number of absorbers, while Q decreases. This is consistent with what has been observed experimentally inside a reverberation chamber loaded with salt water bottles [5].

A degradation of statistical uniformity with the added absorption losses is also observed. To show this, albeit not rigorously, the coefficient of variance (CV) is



Fig. 5 Cabin SE (a) and Q (b) as a function of the number of absorbing bodies at three different frequencies





chosen as the figure of merit to check for statistical uniformity inside the passenger cabin. For a two degree of freedom (2-DOF), Chi-squared distribution, where the random variable is E_x^2 , the CV is

$$CV = \frac{\sigma}{\mu} = \sqrt{\frac{2}{\nu}} = 1, \qquad (3)$$

where σ is the standard deviation, μ is the mean, and $\nu = 2$ is the number of degrees of freedom. This means that when CV begins to diverge from 1, the statistical distribution will no longer be Chi squared, and true statistical uniformity does not exist inside the cavity. Values of σ and μ for 50 MHz E_x^2 distribution, around various center frequencies, were used to calculate CV. Figure 6 shows how CV diverges from 1 as more absorbing bodies are added to the cabin, at two different frequencies, 750 MHz and 1.15 GHz. 750 MHz is chosen as the lowest frequency at which the empty aircraft cabin can support enough modes to be considered reverberant, and 1.15 GHz is chosen as the highest frequency available for analysis. It is clear from Fig. 6 that at both frequencies, the statistical uniformity degrades as more than 50 absorbing bodies are added.

4 Conclusions

In this evaluation of absorption losses inside a simplified aircraft model, it was found that SE increases and Q decreases with the increasing number of absorbing bodies. There were various discrepancies between the physical aircraft and the model: landing gear, engines, seams, slots, and most avionics boxes were not modeled, as they were unknown, and the cockpit windshield may have been coated with a conductive material. The results from this study, however, were expected and consistent with previous observations [5]. It was also observed that the statistical uniformity inside the aircraft cavities degrades with the increasing number of absorption losses. This implies that the statistical uniformity degrades with increasing SE. This raises a question as to whether aircraft with enough losses to provide SE as high as that measured for the Boeing 707 [2] can ever be considered reverberant, and if so, in what frequency range. One may be able to approximate absorption losses inside a reverberant aircraft using the statistical electromagnetics formalism [4], together with SE measurements for different aircraft, and classify absorption losses for each aircraft type. However, in a non-reverberant environment, as in the case of the Boeing 707, absorption losses are much more difficult to quantify, thereby making accurate implementation of absorption losses for CEM models more challenging.

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Part III Antennas

Directional Dependence of the Minimum-Phase Property of the TEM Horn Transfer Function

J.S. McLean, A. Medina, R. Sutton, and H. Foltz

Abstract The minimum-phase property of a broadband asymmetric TEM horn is investigated on and off the boresight axis. It is found that the transfer function conforms very closely to the minimum-phase criterion on boresight, but that off axis there is a significant deviation in the H-plane and in the E-plane at angles above the ground plane, but not below. This behavior can be qualitatively explained in terms of a simple model based on rays representing major sources of radiation. In general, the frequency-domain, minimum-phase criterion will be satisfied only if the time-domain field from the strongest radiation source is the first to arrive at the observation point.

Keywords TEM horn • Transfer function • Minimum phase

1 Introduction

Perfect pulse reproduction requires satisfaction of the distortionless transfer function criterion. Distortionless transfer functions are a small subset of a more general group, minimum-phase transfer functions. When a minimum-phase network exhibits a transfer function magnitude which is nearly constant with frequency, its phase function necessarily satisfies the distortionless transfer function criterion; however, many systems exhibiting flat or nearly flat transfer functions are not minimum phase. An example of a non-minimum-phase antenna is the log-periodic dipole antenna (LPDA), which exhibits a nominally flat magnitude, but which deviates greatly from minimum-phase behavior [1, 2]. On the other hand,

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Fig. 1 Farr Research model TEM-1. Dimensions are in mm. The height of the *top* of the plate above the ground plane at the aperture is 33.3 mm

the response of many broadband-ridged horns conforms closely to the minimumphase criterion when measured in the boresight direction, but deviates from it in directions off the principal axis [3].

The minimum-phase quality of an antenna is intimately associated with the propagation of energy through the system. A minimum-phase transfer function is obtained when there is equivalently a single path through the network or system. In [3] it was surmised that the off-axis deviation was due primarily to interference between the direct radiation from the horn's aperture and fields diffracted by the edge of the horn.

In this paper we investigate the minimum-phase behavior of an asymmetric or half TEM horn [3], on and off its principal axis. A symmetric TEM horn consists of two triangular tapered plates, which would correspond to the walls of a conventional horn which are perpendicular to the E-plane. There are no side walls, and the feeding source is connected to the tips of the plates at the horn's vertex.

A particular implementation of this design is the Farr Research model TEM-1 which is shown in Fig. 1.

The properly designed symmetric TEM horn has nearly perfect on-axis pulse reproduction, but requires the use of a balun if driven from a coaxial line. A half TEM horn consists of a single tapered plate over a larger ground plane and is fed at its tip via a coaxial probe through the ground plane. The half TEM horn with a finite ground plane cannot provide a symmetric radiation pattern, but does not require a balun. The imperfect pulse fidelity of realistic baluns means that from a pragmatic point of view, the half TEM horn may provide the best on-axis pulse fidelity and, moreover, is more easily matched to a 50- Ω source.

One might expect that the half TEM horn would suffer from the same off-axis deviation as a conventional-ridged horn, albeit to a lesser degree, since there are no conducting walls perpendicular to the H-plane and thus no conducting edge parallel to the electric field in the mouth of the horn. On the other hand, there is diffraction from the ground plane edge, and the rolled or chamfered edges used to reduce diffraction in commercial designs do not eliminate the diffraction completely. In this paper, we will examine measured off-axis data on a Farr Research model TEM-1 asymmetric TEM horn to determine if the expected effects appear.

2 Minimum-Phase Criterion

Given any transfer function $H(\omega)$, the minimum-phase function $\phi_{\rm m}(\omega)$ is defined as [4]

$$\phi_{\rm m}(\omega) = \frac{2\omega}{\pi} \int_0^\infty \frac{\ln|H(\omega')| - \ln|H(\omega)|}{{\omega'}^2 - \omega^2} d\omega'. \tag{1}$$

We will define an antenna transfer function $H(\omega)$ to be "minimum phase" if there is some constant *K* such that

$$\arg(H(\omega)) = \phi_{\rm m}(\omega) + K\omega, \tag{2}$$

where the term $K\omega$ represents flat time delay and the phase $\arg(H(\omega))$ has been unwrapped. It is important to note that if an antenna is *not* minimum phase, no adjustment of the time origin can make the transfer function exhibit the minimum-phase property.

As stated earlier, minimum-phase behavior is necessary although not sufficient for distortionless pulse transmission. However, there is additional practical significance when the problem of equalization is considered. In the case of transfer function represented by a rational function of finite order, the minimum-phase property implies that the function has no zeroes in the right-half complex plane [4]. If the poles and zeroes are only in the left-half plane, the system can be systematically equalized to a flat response (at least in principle) by adding a compensating network with zeroes and poles placed to cancel each of the poles and zeroes in the original transfer function. On the other hand, if there are zeroes in the right-half plane, cancellation requires a pole in the right-half plane and is thus not possible with a stable, passive compensating network. Equalization can still be attempted, but it will necessarily be approximated and limited in bandwidth.

An antenna transfer function, of course, cannot in general be represented by a rational function of finite order; nevertheless, such representations can be a good approximation over the useful bandwidth of an antenna. In particular, the combination of a relatively low-order rational function and a constant time delay can very

accurately represent the TEM horn. *Therefore, minimum-phase antennas should be inherently simpler to equalize for broadband, short-pulse applications*. A good example of the difficulty of equalizing *non*-minimum-phase antennas is the LPDA [1].

3 Measured Transfer Functions

The transfer functions were measured using the two-antenna method with nominally identical Farr Research TEM-01 horns, over the frequency range 10 MHz–20.0 GHz. Measurements were made on axis and at 45° off axis in both principal planes.

The E-plane transfer function magnitudes are shown in Fig. 2. Above the ground plane, the horn aperture and the ground plane edge are both visible from the observation point, and the nature of the frequency response, with nulls at multiple frequencies, is consistent with interference between two sources of radiation. Below the ground plane, the field is presumably due mainly to a single source, diffraction from the front edge of the ground plane, and therefore varies smoothly with frequency.

The measured phase in each case was first unwrapped and then compared to its respective minimum-phase function. The minimum-phase functions $\phi_{\rm m}(\omega)$ for each case were computed using Eq. (1) from the amplitude data shown in Fig. 2. The measured phases had a flat time delay extracted prior to comparison as in Eq. (2), with the constant *K* estimated using a least-square fit to the slope only. As can be



Fig. 2 Magnitude of the antenna transfer function in the E-plane for an asymmetric TEM horn for 45° above and below the ground plane and in the boresight direction



Fig. 3 Phase of transfer function of the TEM horn for three directions in the E-plane: 45° above the ground plane, on boresight, and 45° below the ground plane, compared with the minimum-phase function. On boresight and below the plane, there is good agreement, above the plane there is strong deviation



Fig. 4 Magnitude of the transfer function in the H-plane

seen in Fig. 3, there is close agreement between the measured, unwrapped phase and the minimum phase along the principal axis. There is also very close agreement at 45° below the ground plane. On the other hand, at 45° above the plane, the measured and minimum-phase functions are completely different. The H-plane transfer function is shown in Figs. 4 and 5. Minimum-phase behavior is maintained off axis up to 10 GHz and then deviates rapidly.



Fig. 5 Phase of the transfer function in the H-plane

4 Impulse Responses

The impulse responses corresponding to each of the transfer functions, in the E-plane, are presented in Fig. 6. As one would expect from the shapes of the transfer function magnitudes, the main pulse is much better formed on axis and below the ground plane. Above the ground plane, the pulse is broader in time and, significantly, has a pre-pulse prior to the main pulse. As will be discussed below, the existence of the pre-pulse is closely connected to the deviation from minimum-phase behavior.

As a thought experiment to demonstrate this connection, one can consider a fictitious minimum-phase transfer function synthesized by combining the measured magnitude with the computed minimum phase for the same magnitude. In the 45° above case, this allows us to see what the impulse response would look like if the non-minimum-phase behavior was somehow suppressed while leaving the rest of the transfer function undisturbed. The resulting comparison is shown in Fig. 7. Although the resulting function is nonphysical, it shows that the primary effect is to move the pre-pulse to after the main pulse.

5 Discussion

In [3] a physical interpretation of the minimum-phase property derived from the study of multipath propagation was presented. A simple two-ray model indicates that when the there are two paths and the amplitude of the signal on the longer path



Fig. 6 Time-domain impulse response for three directions: 45° above the ground plane, on the principal axis, and 45° below the ground plane



Fig. 7 Time-domain impulse response 45° above the ground plane, derived from frequency-domain transfer function (*top*) and from minimum-phase transfer function (*bottom*)

is greater in magnitude than that of the shorter path, the overall frequency-domain transfer function will have a zero in the RHP. That is, for a transfer function of $H(s) = 1 + Ae^{-s\tau}$, when $\tau > 0$ and A > 1, the zeroes of the transfer function are in the RHP, but for A < 1, the zeroes are in the LHP. More generally, the existence of multiple paths for energy transmission through a system creates the possibility of non-minimum-path behavior, but does not necessarily lead to it. Although the two-ray model is an oversimplification, one might glean that early arrivals of smaller signals from secondary radiation sources destroy the minimum-phase nature of the antenna response, while late arrivals do not.

In the TEM horn, we propose that below the ground plane, the radiation is dominated by a single mechanism (edge diffraction), providing essentially a single path, while above the plane, there is combined radiation from edge diffraction, the mouth, and the horn top plate, each with a different time delay. Thus, it appears that the precursors in the off-axis, time-domain response due to the diffraction from the edge of the horn and radiation along the horn cause deviation from minimum phase. Diffraction from the edge of the ground plane does not, as long as these contributions are smaller in amplitude than the primary aperture radiation.

Thus, the degeneration of the minimum-phase property of the transfer function with angular distance from the principal axis in the E-plane is extremely rapid and more pronounced than that in the H-plane. This behavior is essentially the opposite of that exhibited by the broadband double-ridged horn (DRH) [3]. Part of this may be explained by the absence of sidewalls in the TEM horn, which are an edge diffraction source in the double-ridged horn. In addition, the H-plane dimension of the TEM aperture is quite large (148.4 mm), while the other dimension is quite small, 30.2 mm. The unusual aspect ratio is necessary for the TEM horn to maintain a constant local 50- Ω impedance, thus providing the single localized reflection at the mouth which in turn is responsible for the nearly distortionless on-axis response. Because the DRH guides energy primarily by its ridge at the upper end of its operating frequency range, the DR horn actually has a narrower effective aperture than does the TEM horn.

The elegant simplicity of the half TEM horn is responsible for its excellent on-axis response which is nearly distortionless. The ground plane eliminates the need for a balun, and the taper combined with the aspect ratio then provides a very nearly constant $50-\Omega$ transmission line from the coaxial connector to the radiating aperture. However, off axis the response is degraded due to the dominance of early-time contributions.

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UWB Dual-Polarized Antenna for HPEM Sources

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Abstract A novel dual-polarized exponential tapered slot antenna for high power electromagnetic (HPEM) sources is designed and verified. This antipodal version of an ultra-wideband Vivaldi antenna offers compact sizes and sustains high input power of about 40 kV. Gains of about 8 dBi can be achieved. Both antenna planes are isolated with more than 20 dB.

Keywords Vivaldi • Dual polarized • UWB • HPEM

1 Introduction

Electromagnetic effects on electronic devices of, e.g., improvised explosive devices (IEDs) or critical infrastructures are dependent on polarization angles because of the position and orientation of transmission lines, antennas, etc. [1]. To defeat such devices, antennas for semiconductor-based HPEM sources mentioned in [2, 3] are needed that offer the possibility to change the polarization without rotating the antenna itself.

Thus, wideband dual-polarized antennas are one possible solution which will be investigated in this chapter. Most high power electromagnetic antennas are single linear polarized, like impulse-radiating antennas (IRA) and TEM-horns [2]. These types of antennas can handle input amplitudes of tens of kV up to 1 MV. A dual-polarized antenna avoids manually changing the orientation of existing antennas. In literature there are several dual-polarized antennas published. Some examples are depicted in Fig. 1 [4–8]. These antennas show feeding networks that are not feasible

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Fig. 1 Examples of dual-polarized wideband antennas. (a) Dual-polarized aperture-coupled Vivaldi antenna [4, 5], (b) dual-polarized dielectric rod antenna [6], (c) dual-polarized dielectric rod antenna with integrated feeding [7], (d) quad-ridged horn [8]

Fig. 2 Tri-IRA [9]



to handle high voltages because of their aperture coupling. High efforts are needed to avoid dielectric breakdowns. Antennas like the so-called Tri-IRA [9] in Fig. 2 are not very compact to be integrated in vehicles and similar mobile platforms.

Therefore, a novel compact antipodal dual-polarized Vivaldi antenna will be designed, simulated and characterized. Input impulse amplitudes of about 40 kV with rise times of about 150 ps are aspired.

The chapter has the following contents: Next section describes the main specifications and design procedure of the dual-polarized exponential tapered slot antenna. One focus is held on the selection of the substrate and their thickness in order to sustain the above input amplitudes. A 3D-EM model on CST Microwave Studio is prepared and simulated including the feeding network which is a transition from coaxial to parallel-plate transmission lines. The resulting antenna is then manufactured and their main characteristics are measured in terms of input matching, cross talk, and resulting radiated field with up to 40 kV input amplitude.

2 Antenna Design and Simulation

For the given application a linear dual-polarized and directive UWB antenna with minimized space requirements is necessary. An UWB antenna design that fulfills these requirements is the dual-orthogonal polarized Vivaldi antenna as proposed by the authors in [6, 7, 10, 11]. Since the referenced antenna is designed for low power applications, the design has to be adapted to withstand high impulse amplitudes.

The antenna is supposed to work with a HPEM source which generates pulses with amplitudes of about 40 kV and a rise time of about 150 ps. A radiation of the aforementioned pulses with possibly low distortion and without electrical break-downs must be possible. Based on the principle presented in [10], an improved concept for HPEM systems is shown here: Instead of using a ground metal, an antipodal structure is chosen. This has the advantage that both electrical potentials are always separated by the substrate and therefore a very high electric strength is achieved. Due to the feed of the antenna with a strip line, no balun or aperture coupling is needed.

This simplifies the structure, which results in less distortions and simultaneously the electric strength is further increased, as aperture couplings and slot lines tend to break through for high voltages. In the proposed solution the strip line is directly connected with a coaxial line. To avoid a high impedance mismatch and a possible electrical breakdown, the coaxial line is cut diagonally before the connection with the strip line, whereas simultaneously its outer conductor is deformed and matched the strip line shape (see Fig. 3).



Fig. 3 Simulation of the complete dual-polarized Vivaldi antenna



Fig. 4 Single high electric strength antenna. (a) Concept, (b) simulation results

The final dual-polarized antenna consists of two interlaced planar antennas similarly to the design in [10]. In the following the principle of a single antenna for one linear polarization is explained. As antenna substrate Arlon AD250C with a relative dielectric permittivity of 2.5 and a thickness of 6.23 mm is used. The antenna has a dimension of 270×440 mm. Figure 4b shows the simulation results of the single antenna. The input impedance matching is lower than -10 dB in the frequency range from 0.75 to 3 GHz. The lower frequency limit is directly related to the maximum dimension of the antenna effective aperture. This cut-off frequency allows for the radiation of impulses with the required rise time, which is discussed later in this chapter. As the impulse itself has no relevant energy parts at frequencies larger than 3 GHz, the upper frequency limit of 3 GHz is also sufficient.

Figure 5 shows the principle of intersection of two linearly polarized radiators, which create together a complete dual-orthogonal linearly polarized device. It consists of two of the before presented antennas, which are crossed orthogonally to each other. One antenna is shifted from the center in order to avoid an overlap of the metallization (strip lines). The performance of the single radiator is barely affected by the crossing (cf. Fig. 5b). The isolation between the two antennas is in the range of 30 dB in the relevant frequency range. As the S-parameters of the dual-orthogonal polarized Vivaldi antenna are similar to the single Vivaldi antenna, it can be concluded that the influence of nesting the antennas into each other has marginal influence on the performance. Figures 6 and 7 show the simulated gains of the two antennas in the setup of Fig. 5a. Both antennas exhibit a high maximal directivity of about 8 dBi and the E- and H-planes have a similar performance. Both antennas possess the same main beam directions in the whole designated frequency range. The main beam direction of Antenna 2 squints slightly in the E-plane. The reason for this is a presence of one metallized wing of Antenna 1 in the direct illumination area of Antenne 2. However, the influence is marginal and it can be accepted from the system point of view. Furthermore, by using an antenna array with a proper excitation, the squinting can be suppressed and a shift of the phase



Fig. 5 Antennas nested into each other. (a) Dual-orthogonal polarized Vivaldi antenna, (b) simulation results



Fig. 6 Simulated gain of Antenna 1. (a) E-plane, (b) H-plane

center can be avoided [12]. In the low frequency range, a nearly omni-directional radiation pattern in the H-plane can be observed. In that frequency region the antenna behaves like a dipole due to the small dimensioning w.r.t. the wavelength.

Since the antenna is developed for an impulse-based dual polarization usage, the polarization purity of the radiated signal and the time-domain behavior has to be analyzed. Both are simulated with the help of field probes in the far field and the results are shown in Fig. 8a. The polarization decoupling in the main beam direction is at least 15 dB for both antennas, and the performance for both polarizations is very similar to each other. The small differences in the cross polarizations are resulting from the non-symmetrical construction of the antenna. To evaluate the


Fig. 7 Simulated gain of Antenna 2. (a) E-plane, (b) H-plane



Fig. 8 Polarization characteristics and impulse distortion of the proposed antenna. (a) Polarization decoupling, (b) Gaussian impulse

impulse response, the antenna is excited by a Gaussian pulse with frequencies in the range from 0 to 3 GHz. Figure 8b shows that the pulse is differentiated, which is unavoidable during the radiation process. Furthermore only a weak ringing can be seen. Altogether the proposed and simulated antenna presents a good solution for the radiation of high power pulses in two orthogonal linear polarizations.

3 Measurement Results

Figure 9 shows the dual-polarized Vivaldi antenna with different coaxial transitions. The first measurements are made with a low power 50 Ω coaxial transition. Figure 10 shows the measured input reflection coefficients for both input ports for each single and combined antenna wings. Changes of the reflection parameters for



Fig. 9 Realization of the dual-polarized Vivaldi antenna. (a) With low power coaxial transition, (b) With high power coaxial transition



Fig. 10 Measurements of antenna configuration acc. to Fig. 9a. (a) S-parameters, (b) Radiated impulses

the combined measurements are rather low, so both antenna planes are well isolated. The measured field impulses in both planes are depicted in Fig. 10b and show good agreement to the simulation. These field measurements are performed with a 4 kV impulse generator with rise times of about 120 ps on 50 Ω impedance [2]. Both responses are symmetrical to each other and yield a figure of merit *rE/U* of about 0.78 for the single wings. For synchronous feeding of both antenna ports, the figure of merit increases to 0.98. This is equivalent to a gain of about 8 dBi. Simulation and measurement results show good agreement.

To investigate the behavior of high input voltages, the antenna is fed with the coaxial high voltage transition (see Fig. 9b). A 40 kV on 10 Ω , 150 ps rise time impulse generator is used with a matched transition and a resonator with a tunable center frequency in a range of 100–500 MHz. The resonator is a quarter wavelength impedance step from 10 to 100 Ω and is located between generator output and antenna input. The semiconductor-based impulse generator acts as approximate



Fig. 11 Radiated field impulses and their spectra with 40 kV input voltage and different feedings. (a) Matched feeding, (b) Resonator feeding

ideal voltage source and behaves as a short circuit after impulse generation. An impulse from the generator will be transmitted and reflected at the impedance step and the generator short circuit. Hence, the resonator yields a sequence of several UWB impulses with equidistant time delays according to the length of the $\lambda/4$ -resonator (compare [3]).

The measurement results are shown in Fig. 11. With the matched feeding the antenna radiates an ultra-wideband field impulse which can be seen by the bottom impulse spectrum. The ringing is caused by ground reflections of the anechoic chamber. In Fig. 11b three radiated field impulse sequences with the resonator are depicted. The pulse shape is approximately damped sinusoidal with highest stability which is typical for semiconductor-based impulse generators. The spectral density shows a stable resonance within the frequency range mentioned above. A quality factor of about 3 can be estimated from the diagram. This antenna sustains input impulse voltages of up to 40 kV without any dielectric breakdowns.

4 Conclusion

A dual-polarized exponential tapered slot antenna for high power applications is presented. Input impulse amplitudes of about 40 kV with 150 ps rise time can be radiated in several polarization planes without manually rotating the antenna itself. This very compact antenna design is feasible for integrating such antennas in mobile platforms, like vehicles and UAVs.

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Modal Analysis of Reflector Backed Hybrid Printed Bow Tie Antenna

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Abstract A reflector-backed hybrid of bow tie and elliptical antenna was designed and developed for UWB radar applications like ground-penetrating radar (GPR) and through-wall imaging radar (TWIR). Radiation pattern bandwidth of the antenna is found to be lesser than measured impedance bandwidth of the antenna. This anomaly was analyzed using characteristics modes of the antenna. Method of moments (MOM) using RWG basis function was used to calculate impedance matrix at each spot; frequencies and eigencurrents found were used to establish the optimum bandwidth of the antenna.

Keywords UWB • GPR • TWIR • Characteristic modes • RWG

1 Introduction

The demand of ground-penetrating radar (GPR) applications for better penetration and higher resolution requires lower operation frequencies and much wider bandwidth. As a consequence UWB antennas with a much lower center frequency are needed. There are plethoras of UWB antenna options (Vivaldi, spiral, elliptical, circular, tapered slot, etc.) available for different applications [1]. This wide scope of design choices is limited by the required lower center frequency of operation, radiation restricted to forward space, compactness, and good time-domain performance. We considered a wide variety of printed elements and found an acceptable

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antenna configuration using a hybrid design consisting of a bow tie and an elliptical dipole. One of the important observations found in the measured result of the reflector backed hybrid bow tie that the radiation pattern bandwidth does not match with the impedance bandwidth is brought up in this chapter.

In Sect. 2 of this chapter, design details of antenna element are presented. The analysis and measured results are discussed in Sect. 3. Section 4 presents analysis of characteristic modes of the antenna. Conclusion is presented in Sect. 5 (Figs. 1 and 2).

2 Antenna Element Design

The antenna is a hybrid antenna, which is derivative of a bow tie antenna and elliptical antenna. The major axis of the ellipse in the case of this antenna is 7.162 cm and the minor axis is of order of 4.162 cm. The bow tie element of the hybrid antenna has a design specification of length of one side of the triangle and value of one of its interior angles which is 85.34° and the length of the side opposite



to it is 6.7 cm [2]. The antenna element was designed with a 1.6 mm substrate RT/duroid 5880 ($\varepsilon_r = 2.2$) and the reflector was placed 5 cm behind the printed element as shown in Fig. 3.

3 Analysis and Measurement Results

The antenna geometry, as shown in Fig. 3, is a balanced structure; hence either an ultra-wideband balanced to unbalanced (balun) line or a differential excitation signal is required to feed the antenna. The antenna was analyzed using FEM solver (HFSS v12.1) (Figs. 4 and 5).

The photograph of realized antenna is shown in Fig. 6. Figure 7 shows the differential excitation signal feeding arrangement using two 50 Ω coaxial cables. The scattering parameters of the antenna were measured using four-port network analyzer having two differential sources. Measured return loss of the antenna is shown in Fig. 8.

The measured result of the realized antenna shows a bandwidth of 400 MHz–4 GHz as $|S11| \leq -10$ dB. Radiation pattern of the antenna was measured in spherical near-field measurement (SNFM) facility. Measured radiation



Fig. 4 Return loss (dB) vs. frequency (GHz)



Fig. 5 Input impedance (ohm) vs. frequency (GHz)









pattern at 1, 2, and 3 GHz is shown in Figs. 9, 10, and 11 respectively. Figure 11 shows that radiation pattern of the antenna shows that main beam splits beyond 3 GHz and there is null at the boresight, making antenna useful only up to 2.9 GHz.

This phenomenon is studied systematically using modal analysis wherein planar geometry was modeled using PDE tool in MATLAB. Method of moments with RWG basis function used to calculate input impedance matrix.

4 Analysis of Characteristic Modes

Characteristic modes are defined [3, 4] as the real currents on the surface of a conducting body that depends upon its shape size and are independent of any specific source or excitation. Modal currents or characteristic currents are defined as the eigenfunctions of a particular weighted eigenvalue equation that involves



Fig. 8 Measured return loss (dB) vs. frequency (GHz)



Fig. 9 Measured radiation pattern at 1 GHz



Fig. 10 Measured radiation pattern at 2 GHz

impedance matrix of the body. They are real and equiphasal and form an orthogonal set over the surface on which they exist. The formulation of the theory of characteristic modes for conducting bodies begins with the definition of an operator equation that relates the current J on the surface S of a conducting body with the tangential incident electric field:

$$\left[L\left(\vec{J}\right) - E^i\right]_{\tan} = 0. \tag{1}$$

Characteristic current modes can be obtained as the eigenfunctions of the following particular weighted eigenvalue equation:

$$X\left(\vec{J}_{n}\right) = \lambda_{n} R\left(\vec{J}_{n}\right), \qquad (2)$$

where λ_n are the eigenvalues, \overline{J}_n are the eigenfunctions or eigencurrents, and R and X are the real and imaginary parts of the impedance operator Z. It is inferred that the



Fig. 11 Measured radiation pattern at 3 GHz

smaller the magnitude of the eigenvalue λ_n is, the more efficiently the mode radiates when excited. A mode is at resonance when its associated eigenvalue is zero. The sign of the eigenvalue determines whether the mode contributes to store magnetic energy ($\lambda_n > 0$) or electric energy ($\lambda_n < 0$). Figure 12 shows variation of eigenvalue of the first current mode of the hybrid bow tie antenna with frequency. It is observed that eigenvalue starts being negative, next it resonates ($\lambda_n = 0$), and at the end it keeps a small uniform positive value. Hence, mode J_1 resonates at 800 MHz, 1.6 GHz and 2.9 GHz. Characteristic angles are defined as

$$\alpha_n = 180^\circ - \tan^{-1}(\lambda_n). \tag{3}$$

A characteristic angle models the phase angle, a characteristic current J_n , and the associated characteristic field E_n . When characteristic angle is close to 180° , the mode is a good radiator, while when the characteristic angle is near 90° or 270° , the mode mainly stores energy. Thus, the radiating bandwidth of a mode can be deduced from the slope near 180° of the curve described by the characteristic angles. Variation of characteristic angles of J_1 mode of the hybrid bow tie with respect to frequency. Figure 13 confirms the three resonant frequencies at 800 MHz, 1.6 GHz, and 2.9 GHz in the frequency range of 400 MHz–3 GHz.



5 Conclusion

The analysis and measured results of the reflector backed printed hybrid bow tie antenna show the impedance bandwidth of 400 MHz–4 GHz. Theory of characteristic modes was used to analyze the reason for the null observed in the boresight of the antenna beyond 3 GHz. Modal analysis shows that dominant current mode resonates at 800 MHz, 1.6 GHz, and 2.9 GHz making antenna useful only in the frequency range of 400 MHz–2.9 GHz.

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A Novel Class of Reconfigurable Spherical Fermat Spiral Multi-port Antennas

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Abstract Reconfigurability in antenna systems is a desired characteristic that has attracted attention in the past years. In this work, a novel class of spherical Fermat spiral multi-port antennas for next-generation wireless communications and radar applications is presented. The device modelling is carried out by using a computationally enhanced locally conformal finite-difference time-domain full-wave procedure. In this way, the circuital characteristics and radiation properties of the antennas are investigated accurately. The structure reconfigurability, in terms of frequency of operation and radiation efficiency, is technically performed by a suitable solid-state tuning circuitry adopted to properly change the feeding/loading conditions at the input ports of the antenna.

Keywords Fermat spiral multi-port antenna • FDTD • Radiation properties • Radiation efficiency

1 Introduction

In recent years, a growing effort has been devoted by the industrial and scientific research communities in the development of reconfigurable antennas [1, 2]. This interest is pushed by the need for versatile front ends in future microwave systems, which will support an ever growing number of functionalities such as radars, communication, direction and spectrum "sniffing" or control. Moreover, personal wireless or vehicle-to-vehicle communication devices must typically support a large number of standards (e.g. UMTS, Bluetooth, WiFi, WiMAX, DSRC) [3].

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A reconfigurable antenna can reuse its volume at different frequency bands so that a portion of or the entire structure is involved in a specific mode of operation [1]. Antenna reconfigurability in conventional front-end architectures may be readily achieved by changing the state of suitable switching devices in order to optimize the performance of the device for application in different operative scenarios. Several approaches have been proposed for implementing this concept [4–10]. Most of these approaches rely on either solid-state or electromechanical switches. The former includes switches based on PIN diodes, varactors or fieldeffect transistors (FETs), whereas the latter includes simple relays and a number of different types of microelectromechanical system (MEMS). In the presented study, a suitable solid-state tuning circuitry is used to dynamically adjust the circuital characteristics, such as the frequency of operation and/or radiation properties of a novel Fermat spiral multi-port antenna for next-generation wireless communications and radar applications. The device reconfigurability is technically achieved by changing the feeding/loading condition at the input ports and, thereby, the current distribution within the antenna structure.

The paper is organized as follows. In Sect. 2 the layout of the proposed radiating structure is detailed, whereas the full-wave modelling approach adopted for the relevant electromagnetic characterization is discussed in Sect. 3. A comprehensive investigation into the performance of the considered class of antennas is then presented in Sect. 4. The concluding remarks are summarized in Sect. 5.

2 Proposed Antenna Layout

The topology of the proposed spherical Fermat spiral multi-port antenna is shown in Fig. 1. As it can be readily noticed, the *n*-turn spiral arm, having width $w_1 = 50 \,\mu\text{m}$, is printed on a dielectric hemisphere with diameter $D_s = 4$ mm. To this end, the femtosecond-laser-induced silver electroless plating method detailed in [11] can be usefully employed in combination with a suitable two-photon-absorption-induced photopolymerization of the substrate which, in the presented design, is assumed to be a lossless silicon-like material with relative permittivity $\varepsilon_r = 11.9$. In particular, the proposed radiating structure is excited by means of two substrate-integrated coaxial feeding lines, whose geometrical characteristics have been selected in such a way as to achieve the characteristic impedance $R_0 = 50 \ \Omega$ and a TE₀₁ cut-off frequency $f_c = 61$ GHz. The external conductors of the coaxial cables are physically connected to the ground plane in the back of the dielectric substrate where the solid-state switching circuitry is printed. The adopted feeding structure allows for an easy wide-range tunability of the antenna loading conditions, whilst minimizing the level of spurious electromagnetic energy emission, potentially resulting in a degradation of the device performance at high frequencies.



3 The Full-Wave Antenna Modelling

The analysis and design of complex radiating structures require accurate electromagnetic field prediction models [12]. One such widely used technique is the FDTD technique. However, in the conventional formulation of the algorithm proposed by Yee [13], each cell in the computational grid is implicitly assumed to be filled in by a homogeneous material. For this reason, the adoption of Cartesian meshes could result in reduced numerical accuracy where structures having complex geometry are to be modelled [14]. In this context, the locally conformal FDTD scheme in [15] provides a clear advantage over the use of the staircasing approach or unstructured and stretched space lattices, potentially suffering from significant numerical dispersion and/or instability. Such scheme is based on the definition of suitable effective material tensors accounting for the local electrical and geometrical properties of the device under analysis [16].

By using the mentioned subcell method, the design and accurate full-wave analysis of the considered class of reconfigurable spiral multi-port antennas have been carried out. In doing so, the materials forming the radiating structures have been assumed to be lossless for the sake of simplicity. Furthermore, a voltage source of amplitude V_i and internal impedance Z_i has been employed to excite the *i*th antenna port (i = 1, 2). Under such hypotheses, the locally conformal FDTD characterization of the device has been carried out by making use of a non-uniform computational grid with maximum cell size $\Delta h_{\text{max}} = \lambda_{\text{min}}/340 \simeq 25 \,\mu\text{m}, \,\lambda_{\text{min}}$ denoting the operating wavelength within the dielectric substrate at the maximum working frequency $f_{\text{max}} = 10$ GHz in the excitation signals, which are Gaussian pulses described analytically by the following equation:

$$\Pi_{g}(t) = \exp\left[-\left(\frac{t-t_{0}}{T_{g}}\right)^{2}\right]u(t),$$
(1)

where $u(\cdot)$ denotes the usual unit-step distribution, and

$$t_0/10 = T_g = \sqrt{\ln 10} / (\pi f_{\text{max}}) \simeq 48.3 \,\text{ps.}$$
 (2)

The selection of T_g according to Eq. (2) results in the significant energy content (at - 10 dB cut-off level) of the excitation signals in the frequency band up to f_{max} . The source pulses are coupled into the finite-difference equations used to update the electromagnetic field distribution within the feeding sections. In this way, the incident a_i and reflected b_i power waves at the input ports (i = 1, 2) of the antenna can be evaluated in the time domain directly, whereas the frequency-domain behaviour of the relevant scattering matrix $\underline{S}(Z_1, Z_2)$ as a function of the loading conditions of the device can be determined by Fourier transformation.

In all calculations presented in this paper, a ten-cell uniaxial perfectly matched layer (UPML) absorbing boundary condition [17] has been used at the outer FDTD mesh boundary in order to simulate the extension of the space lattice to infinity. In doing so, a quartic polynomial grading of the UPML conductivity profile has been specifically selected in order to have a nominal numerical reflection error \mathcal{R}_{PML} $\simeq 10^{-4}$. In particular, the accurate numerical modelling of the individual antenna element (see Fig. 1) has required the adoption of a spatial grid consisting of $199 \times 167 \times 90$ cells in the *x*, *y*, *z* coordinate directions, respectively, and resulting in a computational time of about 1.1 s per step on a workstation equipped with a 2.99 GHz Intel Core Duo[®] processor and 3.25 G Byte memory.

4 Antenna Radiation Properties and Circuital Characteristics

The circuital characteristics and radiation properties of the spherical 3-turn spiral multi-port antenna (see Fig. 1) have been investigated in detail for different loading conditions at the relevant input ports.

In particular, where the nominal loading condition $Z_1 = Z_2 = R_0$ is considered, the scattering parameters S_{ij} (*i*, *j* = 1, 2) of the radiating structure feature the frequency-domain behaviour shown in Fig. 2. In this context, it is to be stressed that the considered device is in fact acting as an antenna at those frequencies where the input power reflection level is not perfectly compensated by the transmission coefficient between the two ports, namely, wherein the inequality $|S_{11}|^2 + |S_{21}|^2$ < 1 holds true. In Fig. 2, one can readily notice the natural resonant frequencies relevant to different modes of operation, each of them corresponding to a specific current distribution excited within the antenna volume [18]. As pointed out in



Fig. 2 Frequency behaviour of the scattering parameters of the reconfigurable 3-turn Fermat spiral multi-port antenna under the nominal loading condition $Z_1 = Z_2 = R_0$, $R_0 = 50 \Omega$ denoting the characteristic impedance of the substrate-integrated feeding coaxial cables (see Fig. 1)

Fig. 3, the radiation properties of the considered device are strongly dependent on the current distribution. In particular, at the resonant frequency f = 5.15 GHz, under the assumption that only the port # 1 is excited, the current density features a global maximum along the central segment of spiral arm on the top of the dielectric hemisphere, so resulting in an enhanced directivity at the broadside (see Fig. 3a). On the other hand, where the working frequency is selected to be f = 7.5 GHz, the current distribution is characterized by a local minimum at the midpoint of the spiral arm and, as a consequence, a stretched doughnut-like shape is featured by the antenna radiation pattern (see Fig. 3b). Therefore, in this specific mode of operation, the peak value of the radiated power is observed along the end-fire direction.

The natural resonant processes responsible for the antenna performance can be activated or inhibited by properly tuning the loading condition at the input ports. In doing so, particular attention should be put on the radiation efficiency $\eta(Z_1,Z_2)$ of the device in such a way as to maximize the input power into a radio-wave signal headed in a desired spatial direction. Under the hypotheses detailed in Sect. 3, it is straightforward to show that the power radiated by the antenna, regarded as a two-port network, can be easily expressed in terms of the incident and reflected wave vectors, $\mathbf{a} = \mathbf{a}(Z_1,Z_2) = [a_1(Z_1,Z_2) a_2(Z_1,Z_2)]^T$ and $\mathbf{b} = \mathbf{b}(Z_1,Z_2) = [b_1(Z_1,Z_2) b_2(Z_1,Z_2)]^T$, respectively, as follows:

$$P_{R} = \frac{1}{2} \left(\left\| \mathbf{a} \right\|^{2} - \left\| \mathbf{b} \right\|^{2} \right) = \frac{1}{2} \mathbf{a}^{H} \cdot \mathbf{Q} \cdot \mathbf{a},$$
(3)

where the superscript H denotes Hermitian transposition and



Fig. 3 Angular behaviour of the directivity of the reconfigurable 3-turn Fermat spiral multi-port antenna under the nominal loading condition $Z_1 = Z_2 = R_0$ at frequency f = 5.15 GHz (**a**) and f = 7.5 GHz (**b**), wherein only the port # 1 is excited

$$\mathbf{Q} = \underline{\mathbf{1}} - \underline{\mathbf{S}}^H \cdot \underline{\mathbf{S}} \tag{4}$$

is the so-called dissipation matrix [19], $\underline{1}$ being the usual unit dyadic. In order to facilitate the analysis and optimization of the device, whilst gaining a useful insight in the physical mechanisms responsible for the relevant circuital behaviour, a singular-value-decomposition-based representation of the *S*-parameters is adopted. This factorization is given by

$$\underline{\mathbf{S}} = \underline{\mathbf{U}} \cdot \underline{\mathbf{\Sigma}} \cdot \underline{\mathbf{V}}^{H},\tag{5}$$

where $\underline{\mathbf{U}}$ and $\underline{\mathbf{V}}$ are unitary matrices of left and right singular vectors, respectively, and $\underline{\boldsymbol{\Sigma}}$ is the diagonal matrix of corresponding ordered singular values σ_n . As a consequence, the *S*-matrix relation $\mathbf{b} = \mathbf{S} \cdot \mathbf{a}$ can be conveniently written as

$$\underbrace{\underline{\mathbf{U}}^{H} \cdot \mathbf{b}}_{\mathbf{b}'} = \underline{\mathbf{\Sigma}} \cdot \underbrace{\underline{\mathbf{V}}^{H} \cdot \mathbf{a}}_{\mathbf{a}'}.$$
(6)

Upon considering the normalized wave vectors $\mathbf{a}' = \underline{\mathbf{V}}^H \cdot \mathbf{a}$ and $\mathbf{b}' = \underline{\mathbf{U}}^H \cdot \mathbf{b}$, the device behaviour can be described as the superposition of the natural responses of decoupled one-port junctions, where the *n*th junction presents a positive reflection coefficient σ_n . Thus, one can readily infer that the circuital characteristics and the radiation properties of the antenna are primarily defined by the singular values of

the relevant scattering matrix. As a matter fact, by combining Eqs. (3) and (4) with Eq. (6), the expression of the termination-dependent radiation efficiency is found to be, after some algebra:

$$\eta = \frac{P_R}{P_{\rm in}} = \frac{\mathbf{a}^H \cdot \mathbf{Q} \cdot \mathbf{a}}{\mathbf{a}^H \cdot \mathbf{a}} = \frac{\sum_n \left(1 - \sigma_n^2\right) \left|a_n'\right|^2}{\sum_n \left|a_n'\right|^2},\tag{7}$$

where $P_{\text{in}} = \frac{1}{2} \|\mathbf{a}\|^2 = \frac{1}{2} \|\mathbf{a}'\|^2$ denotes the power accepted at the terminals of the structure. Therefore, the optimal antenna operation is clearly achieved by minimizing the singular values σ_n . To this end, a suitable loading condition (Z_1, Z_2) of the input ports is to be selected properly.

By using the presented circuit-theory-based formulation, the performance of the proposed reconfigurable 3-turn Fermat spiral multi-port antenna has been thoroughly investigated under the hypothesis that the port #2 (see Fig. 1) is terminated in a load with complex impedance $Z_2 = Z_L$ synthesized by means of the mentioned solid-state tuning circuitry printed on the back of the antenna substrate. In this way, it has been numerically found that, upon selecting $Z_L = R_0 + j 2\pi fL$ with L = 10 nH, the global antenna efficiency at the natural resonant frequency f = 9.25 GHz can be increased by about 65% (see Fig. 4), which in turn results in the enhancement of the radar detectability of possible targets or in a larger channel capacity of the radio link. Conversely, the nominal loading condition $Z_L = R_0$ is useful to inhibit the antenna radiation process in a relatively wideband around the aforementioned frequency.

5 Conclusion

A novel class of reconfigurable spherical Fermat spiral multi-port antennas for next-generation radar and wireless communications has been presented. The antenna modelling has been carried out by using a computationally enhanced locally conformal FDTD full-wave procedure. In this way, the circuital characteristics and radiation properties of the devices have been investigated accurately for different loading conditions of the input ports. The structure reconfigurability, in terms of frequency of operation and radiation efficiency, has been technically performed by a suitable solid-state tuning circuitry. The proposed antenna element can be used to design reconfigurable arrays featuring a superior performance in terms of pattern diversity which is, in turn, useful to dynamically increase the channel capacity in different operative scenarios.



Fig. 4 Frequency behaviour of the radiation efficiency of the reconfigurable 3-turn Fermat spiral multi-port antenna under the assumption that the port # 2 (see Fig. 1) is terminated in a load with impedance $Z_L = R_0$ or $Z_L = R_0 + j X_L$, the reactance $X_L = 2\pi f L (L = 10 \text{ nH})$ being synthesized by means of the solid-state tuning circuitry printed on the back of the antenna substrate

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Highly Directive Epsilon Negative Metamaterial-Loaded Circular Patch Antenna for Triple-Band Performance

M.H. Ababil, M.F. Saimoom, S.M.H. Chowdhury, M.R.C. Mahdy, and M.A. Matin

Abstract Based on the idea of additional modified mode(s) (Mahdy et al., IEEE Antennas Wirel Propag Lett 10:869–872, 2011), metamaterial-loaded triple-band rectangular patch antennas have been reported in (Mahdy et al., Prog Electromagn Res Lett 21:99–107, 2011). But the idea of additional modified mode(s) to achieve multiband performance in circular patch antenna has not been reported so far. Recently we have reported the idea of "additional modified mode(s)" in circular shaped patch antennas loaded with metamaterials to achieve multiband performance (Ferdous et al., IET Microw Antennas Propag J 7:768-776, 2013). On the basis of the design algorithm reported in (Ferdous et al., IET Microw Antennas Propag J 7:768–776, 2013), in this chapter, a triple-band circular patch antenna loaded with ENG metamaterial has been proposed. The proposed antenna not only provides good resonance, but also ensures satisfactory radiation performances (directivity, radiation efficiency, and gain) for all the three bands. Achieving a triple-band performance has been possible by modifying $TM_{\delta 10}(1 < \delta < 2)$ mode (using ε negative metamaterial) along with TM₂₁₀ mode modification (due to symmetrical slotting). It is expected that this sort of antenna will be really effective in multiband highly directive applications, especially in satellite communication.

Keywords Triple band • ENG metamaterial • Circular patch • Mode modification

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1 Introduction

In recent times, the application of highly directive and multiband antennas in satellite communications, wireless communications, surveillance, weather, radar, etc. has gained increased popularity. Achieving multiband performance by using a unique radiating structure may simplify the exigency of complex electronic circuitry required to tune many bands at a time. A dual band circular patch antenna has been reported by Wong et al. [1] using conventional dielectrics as substrate and by using symmetrical slots. In Wong's work, conventional TM₂₁₀ mode was modified to yield dual band performance. In that case the first-order mode (TM₁₁₀ mode frequency) and second-order mode (TM_{210} mode frequency) were determined by the antenna geometry structure (i.e., radius of the patch). Alu et al. proposed design method to obtain electrically small rectangular patch antennas using DPS (double positive) ENG metamaterial juxtaposed layer [2]. But broadside null radiation pattern was obtained for such rectangular patches at sub-wavelength regime, illustrated in Fig. 2 of [3]. Consequently, it was then forecasted that all these electrically small rectangular antennas loaded with metamaterial can only be good resonators, not good radiators. But theoretically it is possible to achieve electrically small size without deterioration of radiation performance, if the shape is circular [2, 4] or elliptical [5].

Although the additional modified mode(s) has been reported in rectangular patch antennas [6], in case of circular patch antenna additional, modified mode has rare citation in the literature. A proper MATLAB-based design algorithm for achieving additional unconventional mode in ENG metamaterial-loaded circular patch antenna has been developed in [7]. Based on that design algorithm, in this chapter a novel design of triple-band ENG metamaterial-loaded circular patch antenna has been shown. Its main feature is in its flexibility of choosing resonant frequencies according to user's will for a particular band. Apart from conventional TM₁₁₀ and TM_{210} mode, newly produced unconventional $TM_{\delta 10}$ (1 < δ < 2) mode yields highly directive radiation performance. This designed antenna has two degrees of freedom (radius of the patch and filling ratio of the metamaterial). In our case, the first band, i.e., first-order mode is determined by the patch geometry only. The use of ENG metamaterial instead of conventional dielectrics causes resonance in between first- and second-order mode frequency. The biggest advantage here is that the designer can tune the second band frequency, i.e., $TM_{\delta 10}(1 < \delta < 2)$ mode almost anywhere in between first- and second-order mode by just changing the filling ratio of the ENG metamaterial. The third band is obtained by the modification of TM_{210} mode by using symmetrical slots following Wong's work [1]. The practical implementation of MNG (μ negative) metamaterial is possible with the current technology available [8, 9]. But, achieving the practical design of ENG metamaterial-loaded antenna using Lorentz model in microwave regime [10] is a tough task. Nevertheless, the directivity and gain performance of such antennas are attractive with conventional size. In our research work, we have shown how three highly directive bands can be obtained by using ENG metamaterial as substrate in circular microstrip patch antenna and using symmetrical slots.

2 Antenna Design Structure and Parameter Specification

The geometry of a circular patch with metamaterial block shown in Fig. 1, has been used in designing our proposed antenna. ENG metamaterial with a permittivity of ε_1 is concentric with regular dielectric material having a permittivity of ε_2 . The antenna consists of a metallic patch of thickness ($t_p = 1 \text{ mm}$) and radius (a = 20 mm) at the top. A metallic plate placed at the bottom with radius ($2 \times a = 40 \text{ mm}$) and thickness ($t_g = 2 \text{ mm}$) acts as ground. This chosen optimum radius and thickness of the ground plate causes maximum reflection from the ground that enhances the directivity. As a feed we have used coaxial cable which has a characteristic impedance of $Z_p = 50 \Omega$ with an inner radius of $r_{in} = 0.4 \text{ mm}$ and an outer radius $r_{out} = 1.0456 \text{ mm}$. The feed position of the cable is set at a distance $f_p = 15 \text{ mm}$ from the center so that good impedance matching property is obtained over the operating frequency range. Dielectric substrate height is taken as h = 4.3 mm. Of all the dielectric parameters, $\varepsilon_2 = 1.3\varepsilon_0$, $\mu_2 = 1\mu_0$, $\mu_1 = 1.0\mu_0$ are chosen as the optimum values for these parameters.

However, the other two controlling parameters such as ENG metamaterial's permittivity (ε_1) and filling ratio (η) have been calculated by a MATLAB-based parameter optimization algorithm [7]. The algorithm has been developed by equating dispersive equation (1) which is derived by applying boundary condition of electric and magnetic fields at three separate locations of the patch



Fig. 1 Geometry of a circular microstrip patch antenna partially loaded with metamaterial (ENG). (a) Substrate and ground plane with parameters: a = patch radius = 20 mm, $\eta a = \text{radius of NIM}$ loading area from center = (0.56)(20) mm = 11.2 mm, h = substrate height = 4.3 mm, $t_g = \text{ground thickness} = 2 \text{ mm}$, $D_g = \text{diameter of ground plane} = 40 \text{ mm}$. (b) Metallic patch with parameters: $a_r = \text{slot's inner arc radius} = 19.3 \text{ mm}$, $a_t = \text{slots width} = 0.5 \text{ mm}$, $\beta = \text{slot's arc angle} = 80^\circ$, $f_p = \text{feed position from center} = 15 \text{ mm}$

 Table 1
 ENG metamaterial-loaded antenna's selected permittivity and filling ratio

4.2686 -1.90 0.56	$\varepsilon_1(\operatorname{at} f_r)$ Filling ratio (η)
	-1.90 0.56



Fig. 2 S₁₁-parameter for 20 mm circular patch antenna using ENG as core material

$$2\sqrt{\frac{\mu_{1}\varepsilon_{2}}{\mu_{2}|\varepsilon_{1}|}}\frac{I_{m}(|k_{\rho1}|b)}{I_{m-1}(|k_{\rho1}|b)+I_{m+1}(|k_{\rho1}|b)} = \frac{J_{m}(k_{\rho2}a)Y'_{m}(k_{\rho2}a)-J'_{m}(k_{\rho2}a)Y'_{m}(k_{\rho2}b)}{J'_{m}(k_{\rho2}b)Y'_{m}(k_{\rho2}a)-J'_{m}(k_{\rho2}a)Y'(k_{\rho2}b)}$$
(1)

In Table 1 those optimum parameters are shown at our desired resonant frequency of 4.26 GHz.

Actually, SNG or DNG metamaterials are inherently dispersive and lossy [4, 11]. So, without using dispersive lossy model the simulated results cannot give proper realistic results. Here we have used Lorentz model for ENG metamaterial's dispersive relation:

$$\varepsilon(\omega) = \varepsilon_{\infty} + \frac{(\varepsilon_s - \varepsilon_{\infty})\omega_0^2}{\omega_0^2 - \omega^2 - j\omega\delta}.....(2)$$

where $\varepsilon_{\infty} = 1.0 \ \varepsilon_0$, $\varepsilon_s = 1.23 \ \varepsilon_0$, $\omega_0 = 26.6 \ \text{Grad/s}$, $f_r = 4.26 \ \text{GHz}$, and damping frequency $\delta = 1 \ \text{MHz}$. CST microwave studio [12] simulation by using all the above stated material and geometric parameters gives the following results: *S*-parameter, 3D radiation patterns (Figs. 2, 3, 4, and 5).



Fig. 3 3D view of (a) conventional TM₁₁₀ mode at 3.6665 GHz (b) E and H plane view



Fig. 4 3D view of (**a**) unconventional $TM_{\delta 10}$ (1 < δ < 2) mode's radiation pattern (**b**) E and H plane view



Fig. 5 3D view of (a) conventional TM_{210} mode's radiation pattern at 5.2169 GHz (b) E and H plane view

3 Resonance and Radiation Characteristics of the Antenna

Figure 2 shows that at frequencies $f_{110} = 3.6665$ GHz, $f_{\delta 10}$ ($1 < \delta < 2$) = 4.2686 GHz, and $f_{210} = 5.2169$ GHz resonance occurs which causes the corresponding return losses at these frequencies to fall well below -10 dB. It ensures satisfactory resonance. However satisfactory resonance does not always guarantee good radiation. So in order to ensure good radiation and thus call it a good radiator, we need to look into its radiation patterns at those particular resonant frequencies also. The radiation patterns are obtained at the resonant frequencies, i.e., f_{110} , $f_{\delta 10}$, and f_{210} , respectively.

From the above 3D radiation patterns, we see that the conventional TM_{110} and TM_{210} modes (Figs. 3a and 5a) as well as the unconventional modified $TM_{\delta 10}$ ($1 < \delta < 2$) mode (Fig. 4a) show satisfactory *z*-directed broadside radiation. The conventinal TM_{110} mode has the gain 9.31 dB and another conventional



Fig. 6 On the y = 0 plane, at $f_{110} = 3.6665$ GHz, (**a**) electric field distribution and (**b**) current distribution



 TM_{210} mode has the gain 8.38 dB. However, the gain of the modified mode $TM_{\delta 10}$ (1 < δ < 2) at frequency 4.2686 GHz is 9.46 dB which is even higher than that of the conventional modes. From the E and H plane view of the radiation patterns (Figs. 3b, 4b, and 5b), satisfactory radiation performance also becomes evident.

4 Electric Field and Current Distribution Over the Patch

From the electric field distribution at the plane y = 0, it is apparent that for the conventional TM₁₁₀ mode (Fig. 6a) and TM₂₁₀ mode (Fig. 8a), electric field flips its sign passing from one side to the other side of the patch, thus satisfying the condition for broadside radiation. Again, in case of modified TM_{$\delta 10$}($1 < \delta < 2$) (Fig. 7a) electric field also flips from one side to the other side of the patch. As a



Fig. 8 On the y = 0 plane, at $f_{210} = 5.2169$ GHz, (a) electric field distribution and (b) current distribution

result of which the electric field distribution of unconventional modified $TM_{\delta 10}(1 < \delta < 2)$ mode looks like that of conventional TM_{110} mode which also satisfies the condition for broadside radiation. It is mentionable here that neither the gain nor the radiation characteristics of the conventional modes are hardly influenced on account of choosing the resonant frequency in between TM_{110} and TM_{210} modes. It provides us a great deal of flexibility in choosing resonant frequency of $TM_{\delta 10}$ mode ($1 < \delta < 2$). By the comparison of Figs. 7b and 8b with Fig. 6b, it becomes evident that these three current distributions are almost the same. That is why, $TM_{\delta 10}$ mode ($1 < \delta < 2$) and TM_{210} mode frequencies show z-directed radiation patterns like that of TM_{110} mode. The overall performance of this sort of circular patch antenna, when compared with that of its rectangular counterpart [13] is very good.

5 Conclusions

In this chapter, a novel tri-band circular patch antenna with high directivity performance in all three bands has been reported. Using ENG metamaterial as substrate provides us with an additional modified mode with high gain. By controlling ENG metamaterial's filling ratio, different frequencies can be tuned in. Moreover, the use of symmetrical slots yields satisfactory broadside radiation of another conventional mode. One modified mode along with two conventional modes yields a triple-band antenna performance. This novel antenna design can be a suitable choice where highly directive gain along with user's flexibility in choosing resonant frequencies is a prime concern. In our proposed triple-band antenna design loaded with ENG metamaterial, all three bands have almost equal gains. So, all the three bands can be used separately for highly directive multiband antenna applications such as in satellite communication, radars, and so on.

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Part IV Pulsed Power

A New Set of Electrodes for Coaxial, Quarter Wave, and Switched Oscillators

F. Vega, F. Rachidi, D. Giri, B. Daout, F. Roman, and N. Peña

Abstract We propose a new profile for the electrodes forming the radial transmission line of a switched oscillator (SWO). The profile is formed using a curvilinear orthogonal coordinate system, based on a specific 2D transformation called Logarithmic-Tangent (Ln-Tan). The proposed profile fulfills the geometric criteria for an optimal distribution of the electric field. Basic equations describing the Ln-Tan coordinate system are presented and discussed.

Keywords Switched oscillator • Conformal electrodes

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1 Introduction

A switched oscillator (SWO) is a mesoband radiating system proposed by Baum in [1]. The system consists of a DC-charged low-impedance coaxial transmission line that is discharged at one end by a spark gap and is connected at the other end to a high-impedance antenna. During the charging phase, the capacitor formed by the SWO is slowly charged until breakdown occurs at the pressurized spark gap end. The produced wave propagates towards the antenna. Due to the mismatch between the coaxial transmission line and the antenna, only part of the energy is radiated; most of the energy is reflected back to the spark gap, where the wave is re-reflected by the low-impedance arc. The reflections occurring at this point, opposite in sign, produce a damped sinusoid-like signal with a central frequency of about $f = v_p/4L$, where *L* is the length of the line and v_p is the wave propagation velocity.

The overall geometry of an SWO is depicted in Fig. 1. Notice that the electrodes of the spark gap form a radial transmission line (RTL) that progressively becomes a low-impedance coaxial transmission line which is further connected to the antenna.

2 Background

In order to prevent distortion of the signal transmitted to the antenna, the discharge should be produced exactly on the axis of symmetry of the SWO so that all the wave front originated at the discharge point would arrive to the antenna in phase. The



Fig. 1 Quarter wave coaxial switched oscillator (SWO). Notice the radial transmission line (RTL) existing at the low-impedance end of the SWO. The SWO is connected to an arbitrary antenna
location of the discharge point depends, in turn, on the electrostatic field distribution, previous to the disruption.

The problem of the distribution of the electrostatic field in an SWO has been discussed by Giri et al. in [2]. A more detailed discussion and design technique was proposed by Armanious et al. in [3], where an iterative method, based on the equivalent charge distribution principle, permitted the generation of two exponential profiles forming the electrodes. The formed geometry produces an electric field that is maximum on the axis of symmetry. However, the resulting electric field is not monotonically decreasing as we move away from the axis.

We propose in the next section a new set of electrodes that maximizes the electric field on the axis and additionally generates a field distribution that smoothly decreases as we move away from the axis.

3 Conditions for Optimal Electrostatic Distribution

The probability of producing breakdown on the axis of symmetry maximizes if the magnitude of the electrostatic field at the time of occurrence of the discharge is maximum at the axis of symmetry of the SWO, between the electrodes. The conditions necessary to produce the desired field distribution are the following:

- (i) The distance between the electrodes should be minimum at the axis of symmetry.
- (ii) The distance between the electrodes should monotonically increase as we move towards the coaxial line.
- (iii) The profile of the electrodes as well as its first space derivative should be continuous.

On the other hand, the interelectrodic distance at the axis d_{gap} and the cross section of the coaxial transmission line r_i , r_o (as defined in Fig. 1) are generally specified parameters and can be included in the analysis as two additional conditions:

- (i) The distance between the electrodes at the axis of symmetry should be d_{gap} .
- (ii) The distance between the electrodes at the junction with the coaxial line should be $\Delta r = r_{o} r_{i}$, coinciding with the dimensions of the coaxial transmission line.

A set of curves fulfilling these conditions can be formed using an orthogonal curvilinear space, based on the conformal Ln-Tan transformation proposed by Moon in [4]. The electrodes are generated by two parallel surfaces, conformal to one of the axis of the curvilinear coordinates. This idea is analogous to the one used by Rogowski in [5], with the difference that he used the Maxwell transformation instead.

4 A Method for Generating a Curvilinear Coordinate Space from Conformal Transformation

The method of generating a curvilinear space starting from a conformal transformation was proposed by Moon and Spencer in [6]. The method starts by performing a conformal transformation from the W to the Z plane. The orthogonal curved lines produced on the Z plane can be regarded as a 2D curvilinear coordinate system, which can be either translated or rotated, in order to generate a 3D coordinate system. The procedure can be summarized as follows:

The transformation from the W to Z complex planes is

$$Z = f(W), \tag{1}$$

where *f* is an analytical function and *W* and *Z* are complex planes:

$$W = u + iv \quad Z = x + iy. \tag{2}$$

The parametric form of this new set of orthogonal curves can be obtained from the real and imaginary parts of Eq. (1) as

$$x = f_1(u, v) = \operatorname{Re}[f(W)]$$
 $y = f_2(u, v) = \operatorname{Im}[f(W)]$. (3)

The resulting curvilinear grid can be used to generate new coordinate systems. For example, if the Z plane is extruded in a perpendicular direction, a cylindrical coordinate system (u, v, w) can be obtained, where the relationship with the Cartesian coordinates is

$$x = f_1(u, v)$$
 $y = f_2(u, v)$ $z = w$. (4)

If, on the other hand, the Z map is rotated around the original y-axis, we obtain a rotational coordinate system (u, v, w). The new relationships with the Cartesian coordinates are given by

$$x = f_1(u, v) \sin(w)$$
 $y = f_1(u, v) \cos(w)$ $z = f_2(u, v)$. (5)

A similar procedure can be applied if the map is rotated around the *x*-axis. On this new coordinate system, the infinitesimal arc length element (dl) is [4]

$$dl = \sqrt{g_{11}du^2 + g_{22}dv^2 + g_{33}dw^2},\tag{6}$$

where the metric coefficients g_{ii} are defined as

$$g_{11} = \left(\frac{\partial x}{\partial u}\right)^2 + \left(\frac{\partial y}{\partial u}\right)^2 + \left(\frac{\partial z}{\partial u}\right)^2 g_{22} = \left(\frac{\partial x}{\partial v}\right)^2 + \left(\frac{\partial y}{\partial v}\right)^2 + \left(\frac{\partial z}{\partial v}\right)^2$$

$$g_{33} = \left(\frac{\partial x}{\partial w}\right)^2 + \left(\frac{\partial y}{\partial w}\right)^2 + \left(\frac{\partial z}{\partial w}\right)^2.$$
(7)

5 The Logarithmic-Tangent Coordinate System

The conformal transformation generating the Ln-Tan coordinate system is

$$Z = \frac{2a}{\pi} \ln(\tan(W)) - ia, \qquad (8)$$

where u and v are defined on the domain

$$0 \le u < \pi/2 \quad 0 \le v \tag{9}$$

while *x*, *y* are defined on the domain

$$-a \le y \le a \quad -\infty < x < \infty \tag{10}$$

in which a > 0 is a constant.

The 3D space system (u, v, w) can be generated by rotating the transformed plane as

$$x = \frac{a}{\pi} \ln\left(\frac{\sin^{2}(u) + \sinh^{2}(v)}{\cos^{2}(u) + \sinh^{2}(v)}\right) \cos(w) \quad y = \frac{a}{\pi} \ln\left(\frac{\sin^{2}(u) + \sinh^{2}(v)}{\cos^{2}(u) + \sinh^{2}(v)}\right) \sin(w)$$
$$z = \frac{2a}{\pi} \tan^{-1}\left(\frac{\sinh(2v)}{\sin(2u)}\right) - a \tag{11}$$

For the sake of simplicity, we'll work on the *x*–*z* plane (w = 0), for which

$$x = \frac{a}{\pi} \ln\left(\frac{\sin^2(u) + \sinh^2(v)}{\cos^2(u) + \sinh^2(v)}\right) y = 0 \quad z = \frac{2a}{\pi} \tan^{-1}\left(\frac{\sinh(2v)}{\sin(2u)}\right) - a.$$
(12)

The x-z equations represent two perpendicular sets of parametric curves. The first set (called here the *v*-set) can be generated using *u* as parameter and *v* as a constant. The second set (called here the *u*-set) uses *v* as parameter, while *u* is held



Fig. 2 Logarithmic-Tangent curves: u-set curves (blue), v-set curves (black)

constant. Figure 2 shows some examples of this family of curves. Notice that if $0 < u < \pi/2$, only the bottom half of the space (-a < z < 0) is generated.

We propose to form the electrodes of the SWO taking a pair of curves belonging to the *v*-set and rotating them around the *z* axis. The coaxial line section can be connected at the extremities of the curves, at the points where u = 0, $u = \pi/2$ (the horizontal axis z = 0).

The parameters of the generated profiles are $v = v_1$ and $v = v_2$ with $0 < u < \pi/2$. If a constant difference of potential is applied between v_1 and v_2 , the formed isopotential lines will be conformal to the *v*-set and the electric field stream lines will be conformal to the *u*-set [7]. Notice that, at the *x*-axis, the stream lines are parallel to the *x*-axis, coinciding with the direction of the stream lines inside the coaxial line (which are radial).

The choice for this coordinate system is based on the fact the curves belonging to the *v*-set fulfill conditions (i), (ii), and (iii):

Conditions (i) and (ii): by simple inspection of Fig. 2, it can be concluded that the distance between any pair of curves belonging to *v*-set is minimum at the axis of symmetry (*y*-axis, $u = \pi/4$) and increases as we move away from the axis of symmetry ($u - > \pi/2$, u - > 0). This can be demonstrated by calculating the length of an arc u = constant, between v_1 and v_2 . From Eq. (6),

$$l_{\nu}(\nu_{2},\nu_{1},u) = \int_{\nu_{1}}^{\nu_{2}} dl_{\nu} = \int_{\nu_{1}}^{\nu_{2}} \sqrt{g_{22}} d\nu = \frac{a4}{\pi}$$
$$= \int_{\nu_{1}}^{\nu_{2}} \sqrt{\frac{1}{\left(\sin\left[2u\right]^{2} + \sinh\left[2v\right]^{2}\right)}} d\nu.$$
(13)

This integral can be expressed as

$$l_{\nu}(\nu_{2},\nu_{1},u) = -\frac{2ia}{\pi}\csc[2u]F\Big[2i\nu,\csc[2u]^{2}\Big]\Big|_{\nu_{1}}^{\nu_{2}},$$
(14)

where $F(\varphi, k)$ is the incomplete elliptic integral of the first kind:

$$F(\varphi,k) = \int_0^{\varphi} \frac{d\theta}{\sqrt{1 - k^2 \sin^2[\theta]}}.$$
(15)

For a fixed v, Eq. (14) minimizes at $u = \pi/4$ and increases monotonically as u moves towards either 0 or $\pi/2$; see, for example, [8, p. 593].

Condition (iii): The full derivative of Eq. (8) is

$$\frac{dZ}{dW} = \frac{2a}{\pi} \csc[W] \sec[W].$$
(16)

The above equation is continuous for all W except for W = 0, which corresponds to the plane z = a.

6 Calculation of the Surfaces

Conditions (iv) and (v) deal with the calculation of the profiles, namely, the selection of the constants v_1 , v_2 , and a, defining the profiles of the electrodes satisfying the constraints on the interelectrodic distance at the axis of symmetry and at the junction with the coaxial line.

Condition (iv): The distance between the curves is d_{gap} at $u = \pi/4$ (x = 0, the axis of symmetry). Using Eq. (12) for *z*, d_{gap} can be calculated as

$$d_{\rm gap} = \frac{4a}{\pi} (\operatorname{Arc} \operatorname{Tan} [\operatorname{Sinh}[2v_2]] - \operatorname{Arc} \operatorname{Tan} [\operatorname{Sinh}[2v_1]]).$$
(17)

Condition (v): The coaxial line is connected at the *x*-axis, z = 0. At this axis, the curvilinear coordinates are $(u = \pi/2, v = v_1)$ and $(u = \pi/2, v = v_2)$ and the rectangular coordinates are $(x_1 = r_1, z = 0)$, $(x_2 = r_0, z = 0)$, where r_1 , r_0 are the inner and outer radii of the coaxial transmission line and x_1 , x_2 belong to the v_1 and v_2 curves, respectively.

Using Eq. (12) for x, this can be calculated as

$$x_2 = \frac{2a}{\pi} \text{Log}[\text{Tanh}[v_2]]; \ x_1 = \frac{2a}{\pi} \text{Log}[\text{Tanh}[v_1]]$$
(18)

The resulting 3×3 system can be solved numerically, and the value of the constants *a*, v_1 and v_2 can be calculated.

7 Conclusions

A new profile for the electrodes forming the radial transmission line of a switched oscillator (SWO) was presented. The profile is formed using a set of conformal curves generated using a specific 2D transformation called Logarithmic-Tangent (Ln-Tan). The proposed profile is shown to guarantee a smooth continuity (up to the first space derivative) at the junction between the electrodes and the coaxial line. Basic design equations describing the Ln-Tan electrodes were presented and discussed.

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Performances of a Compact, High-Power WB Source with Circular Polarization

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Abstract This paper presents the design and the performances of an embedded high-power microwave (HPM) wideband source, developed and built at the French-German Research Institute of Saint-Louis. The system was intended for dual use, homeland security, and military applications. It is powered by a 400 kV compact Marx generator with specificities in coaxial design and low energy. The slow monopolar signal from the Marx is sharpened using a pulse-forming stage, made of a switching module pressurized with nitrogen, followed by a monopulse-tomonocycle converter. The duration and rise times of this signal could be adjusted by varying the pressure and space between electrodes. Repetitive operations were performed up to 100 Hz during 10 s without a gas flow. Two kinds of antennas can be connected to the source. The first one is a TEM horn, with an optional dielectric lens, that radiates a vertically polarized UWB short pulse. The second one is a nine-turn helix, working in Kraus monopolar axial mode and radiating a circularly polarized wideband signal along the main axis. A dedicated conical reflector increases its directivity and bandwidth. The whole source is designed to be embedded inside an aluminum trailer, powered by batteries and remote controlled through an optical fiber.

Keywords High-power microwaves • Wideband and ultra-wideband sources

• Marx generator • Pulse-forming line • Helical antenna • TEM horn antenna • Dielectric lens

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1 Introduction

UWB short-pulse sources offer new possibilities for defense and security applications. These include fast boat and car neutralization, improvised explosive device disruption, and wideband jamming. In order to explore the offered possibilities of high-power UWB transient signals for security purposes, a project was launched in 2005 by the French-German Research Institute of Saint-Louis (ISL) which is supported by both the French and German Ministries of defense. The aim of this study is to design and build a mobile, autonomous, and remote-controlled prototype that could radiate WB or UWB pulsed electric fields in the MV figure of merit class (product of the distance by the electric field in the far-field region). This article deals with the improvements brought to our first lab prototype, which was detailed in the previous UWB SP 9 book.

2 Dedicated Compact Marx Generator

In order to improve the global efficiency of the system under strong constraints of compactness, a dedicated Marx has been designed to feed the system. This generator has been dimensioned in such a way as to deliver only the amount of energy required to supply the pulse-forming line. In comparison with typical X-ray Marx generators that were used previously, a decrease in its equivalent capacitance achieves this goal and helps to reduce the output voltage rise time. Thus, the reliability of the peaking switch breakdown is improved and the dielectric stress of the PFL components is minimized. Moreover, this design represents an advantage regarding long bursts at a high pulse repetition frequency and even more when the prototype is battery operated. This compact high-voltage generator built at ISL is composed of 11 stages which are inserted inside a metallic tube fitted with a polyethylene jacket. Each stage is hardened to withstand discharges on loads that present significant mismatch. They are made of a couple of capacitors of 1.1 nF and two 17 μ H charging inductances. A hole in the center of the stages is fitted with two spherical electrodes that connect stages together during the discharge. This design allows the UV flashlight emitted by successive breakdowns to propagate throughout the generator; it produces a pre-ionization which helps the erection mechanism. All these parts are packed in a resin compound in order to improve both electrical and mechanical strengths. In this configuration, the charging voltage can reach 50 kV with a rise time lower than 10 ns (measured on a 50 Ω load). The operating voltage reaches values up to 500 kV (in open circuit). Nevertheless, a nominal voltage of 30 kV is sufficient to drive the shaper with a 300 kV pulse. Depending on the power supplies, this Marx generator is able to operate, without gas flow, at repetition rates up to 120 Hz during 10 s or 1 h at 10 Hz. The volume of the Marx generator is about 30 L and it weighs around 50 kg.

3 Pulse-Forming Stage

3.1 Motivation

The pulse shaper is of major importance in a high-power short-pulse source. Indeed, the expected bandwidth requires an ultrashort pulse with fast slopes. Even if our dedicated Marx generator has a rise time in the order of 5–10 ns, the output pulse must be sharpened in order to cover the specified wide spectrum.

We found that a bipolar (monocycle) signal presents several interesting features for high-power WB/UWB radiation. As shown in Fig. 1, there is no DC component in the spectrum of the bipolar pulse. This kind of pulse avoids residual charges on the antenna after several bursts and thus the short circuiting of radiators is not required, increasing the human operator's safety. It also reduces the risk of parasitic flashovers for a given peak-to-peak amplitude. Moreover, as there is only a small low-frequency content in the input signal of the antenna, the radiated fields can be focused more efficiently. Finally, the fast mid-cycle falling edge of the bipolar pulses enhances the figure of merit due to the high dV/dt.

3.2 Design

The presented prototype is the last version of the one previously developed at ISL. Both are based on the active conversion of a monopulse to a monocycle [1, 2]. The "slow" Marx generator output is sharpened by a switching stage providing a fast monopulse signal. This pulse is then converted into a monocycle through a pulse-forming block. In practical terms, the PFS of GIMLI is composed of two spark gaps (as pulse sharpener) and a Blumlein pulse-forming line (as converter). Both spark gaps are mounted in a 60-mm outer diameter aluminum vessel that can be pressurized up to 9 MPa with N₂ (Fig. 2).

3.3 Manufactured Devices

Our last version of the PFS is made of aluminum under strong constraints of compactness. It is a modular system allowing the output signal to be tuned by changing several parameters such as gases, pressure, distance between electrodes, and the Blumlein part. In fact, the bipolar signal at the output of the PFS has a total duration which can be adjusted from 1 to a few ns by using different lengths of the Blumlein module (Figs. 3 and 5).



Fig. 1 Comparison between monopolar pulse and monocycles in time $\left(a\right)$ and frequency $\left(b\right)$ domains

3.4 Typical Results

Figure 6 shows the first tests of the generator connected to a PFL. At a charging voltage of 33 kV, the output voltage of the Marx generator reaches 325 kV with a



Fig. 2 Simplified coaxial design of the bipolar pulse shaper



Fig. 3 Pulse sharpener associated with different Blumlein modules

10–90 % rise time of 7.5 ns. Otherwise, the peak-to-peak voltage generated by the PFL reaches 360 kV with a 10–90 % rise time of 200 ps (Fig. 4).

Three different Blumlein were manufactured for our applications. For example, the shortest one generates a signal composed of two opposite Gaussian pulses separated by less than 500 ps (Fig. 5).



Fig. 4 Output voltage waveforms (black, Marx output; red, pulse-forming stage output)



Fig. 5 Typical output signals in the time and frequency domains (*red*, short; *green*, medium; *blue*, long Blumlein modules)

4 High-Power Antenna

4.1 Ultra-wideband TEM Horn Antenna with Lens

At the very beginning of the project, our first objective was to radiate a short signal with the maximum amplitude and bandwidth. The first antenna developed was based on a traveling wave antenna and a double-ridged TEM horn. In order to minimize the radiating element size and to avoid the problem of balanced excitation at a very high voltage, we design a half antenna, lying on a metallic ground plane. First experiments were performed inside our (semi)anechoic chamber. The measured S_{11} parameter was lower than -15 dB over [120 MHz–6 GHz] and radiated fields in the range of 140 kV/m at 10 m were reported [2–4]. In order to equip our mobile demonstrator, a new design was achieved. We tried to reach similar EM performances with a more compact structure. Due to integration constraints (the local ground plane is too small, the high-voltage input is located at a distance from the floor), a full TEM horn was chosen. The input of the antenna consists of a dedicated transition between a coaxial guide and a strip line; the dielectric used is PTFE. An optional dielectric lens could be placed at the aperture of the horn, which improves the axial gain and directivity (Fig. 6).

As can be seen in Fig. 7, the radiated fields computed at 10° from the main axis decrease by a factor >3. At 20° we only record 11 % of the peak-to-peak value in both planes. The excitation was set to be a 300 kV peak-to-peak impulse with a dt between pulses of about 450 ps (i.e., typical signal at the output of the smallest pulse-forming line). The expected axial performance is higher than 40 kV/m at 50 m, which is a good performance compared with our first lab prototype.



Fig. 6 Plane wave at the output of the mobile prototype, numerical simulation



Fig. 7 Radiated E-fields, E-plane (a), and H-plane (b) (*red*, 0° ; *blue*, 10° ; *green*, 20° from the main axis)

4.2 Wideband Helical Antenna with Reflector

Most of the time, it can be difficult to define with a good accuracy the relative position of the transient HPM source and of the sensitive circuits which are embedded inside a potential target. With the TEM horn antenna that radiates a linearly polarized electrical field, the coupling to the target—which could be



Fig. 8 Manufactured prototype during the first high-power experiments

"polarization selective"-varies by many dB from the worst case to the best one. In order to obtain a more constant range of effects when trying to disrupt an unknown or a moving system, an antenna which radiates circularly polarized fields over some pseudo-period signals was developed. The selected structure is a helical antenna, radiating one main lobe with a maximum intensity along the main axis (Kraus monofilar axial mode). To excite this mode, the diameter and the spacing of the helix should be defined with respect to the frequency/wavelength considered. The antenna gain increases with the number of turns of the helix. Unfortunately, this leads to a more frequency-selective device. In order to maintain the antenna efficiency over a wideband, a nine-turn structure was selected as a good compromise. A small ground plane at the bottom of the antenna and a conical reflector allow the radiated field intensity to be increased and minor lobes to be minimized at oblique angles and over a wide frequency band (Fig. 8). The typical input impedance for helical antennas is close to 120 Ω . In order to feed the radiating element with our standard 50 Ω output, a wideband impedance transformer was implemented as an input transition. Unfortunately, the matching is slightly degraded by the plastic parts here to avoid risks of parasitic flashovers.

In order to determine the axial radiated fields, the emitter and a probe were placed in the ISL anechoic chamber, facing each other at a distance of 8 m from phase centers. The probe is polarization selective, so that only the projected E-fields over one plane are recorded.

The electrical source was set to deliver the three different typical signals that can be generated, depending on the Blumlein length, as described previously. Measurements in the time domain exhibit a duration of 10 ns with some pseudo-periods. The Fourier transform shows a first cutoff frequency of 380 MHz (depending on the impedance transformer bandwidth) and a wideband coverage up to 1.2 GHz (Fig. 9). The figure of merit varies from 220 to 400 kV, depending on the selected Blumlein. In order to check the signal polarization, the probe was rotated along the main axis over 180° . The shape of the signal slightly evolves, but the amplitude remains similar over the whole of the measurements, which is a good point regarding potential applications (Figs. 10 and 11).



Fig. 9 Radiated fields at 8 m from the antenna (*red*, short; *green*, medium; *blue*, long Blumlein modules)



Fig. 10 Radiated fields inside the anechoic chamber-circular polarization



Fig. 11 Final prototype in wideband configuration

4.3 Integration on the Trailer

The described electrical source including a capacitor charger, remote control through an optical fiber, and one antenna is embedded inside an aluminum trailer. The size of the prototype is $3 \times 1.25 \times 1$ m; it weighs about 700 kg, including lead batteries.

5 High-Power Wideband Radiation

First outdoor tests were performed in wideband configuration. The transient signal radiated close to the ground was recorded in order to check the radiation patterns and the propagation of circularly polarized signals on the battlefield. With the smallest Blumlein and a 30 kV charging voltage, the amplitude of the E-field reaches 27.5 kV/m at 10 m from the trailer on both the vertical and horizontal polarizations. At 90° and 180° of the main axis, the maximum amplitude is below 1 kV/m.

6 Conclusion

The proposed electrical source achieves a very good compromise between size, reliability, and performances. Using embedded accumulators, a gas tank, and a full remote control, it can operate during hours at low pulse repetition rates or deliver bursts in the range of 100 Hz/10 s. The use of a coaxial 50 Ω output allows this source to be connected to many different devices, such as different types of antennas. A couple of radiating structures has been built. They allow the emission of strong WB or UWB electromagnetic signals with some interesting features concerning the covered adjustable bandwidth. These elements have been integrated into a trailer for outdoor experiments. Further experiments are currently under progress to validate efficiency and long-term operations at high levels.

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Design Considerations for a Switch and Lens System for Launching 100 ps, 100 kV Pulses

P. Kumar, S. Altunc, C.E. Baum, C.G. Christodoulou, and E. Schamiloglu

Abstract Recent research has shown that it is possible to kill certain skin cancers by the application of fast, high-amplitude, electromagnetic pulses. An impulse radiating antenna (IRA) with a prolate-spheroidal reflector is one such device that can be employed for this purpose. In this chapter the design of a switch system to launch high-voltage (>100 kV), fast (100 ps) spherical TEM waves from the first focal point of a prolate-spheroidal IRA is described. Spherical and cylindrical pressure vessel designs are considered, where the pressure vessel also serves as a (launching) lens.

Keywords Dielectric lens • Hydrogen switch • Impulse radiation

1 Introduction

Recent research has shown that it is possible to kill certain skin cancers (melanoma) by the application of fast, high-amplitude, electromagnetic pulses [1-4]. An impulse radiating antenna (IRA) with a prolate-spheroidal reflector is one such device that can be used for this purpose. A prolate-spheroidal IRA (PSIRA) can noninvasively deliver narrow beam, 100 ps, 100 kV, electromagnetic pulses to a

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biological target in the near field [5-7]. The side-view schematic of a 60° four feedarm PSIRA is shown in Fig. 1.

A 100 ps ramp-rising step launched from the first focal point $(-z_o)$ of the PSIRA is guided by the feed arms and focused into a target located at the second focal point (z_o) . The PSIRA will ultimately be sourced with a few 100 kV to obtain high fields at the second focal point. A switch system and "launching" lens are necessary to avoid dielectric breakdown at such high voltages and to ensure the launch of (approximate) spherical TEM waves from the first focal point. The switch system consists of switch cones, a pressure vessel, and a gas (typically hydrogen) chamber. The design and numerical simulations of such a switch system and launching lens are presented in this chapter.

CST Microwave Studio[®], a three-dimensional, finite-integral time domain (FITD), commercially available software, was used to numerically simulate our switch system designs. Although the input for the numerical simulations presented here is only 1 V, analogous designs for the prototype IRA tested at high voltages (>100 kV) indicate that similar results can be expected for the switch system configurations in this chapter [8].

2 Initial Launching Lens Designs

The initial approach to the design of a launching lens followed methodologies similar to those used in the prototype IRA where the feed point and the focal point are assumed to be spatially isolated [9]. The objective of the lens design is to ensure that within the lens a spherical TEM wave is centered on the feed point. However, outside the lens an approximate spherical TEM wave is centered at the first focal point. The uniform and nonuniform designs explored are summarized below.

2.1 Uniform Lens Design

Analytical calculations defining the boundary of a dielectric lens, with fixed relative permittivity, are detailed in [10]. These equations are derived using high-frequency (optical) approximations. It is shown that the dielectric constant required for such a design must be greater than 25. This constraint, a serious drawback since materials with such high dielectric constants have a large loss and dispersion, was the motivation to explore nonuniform lens designs that would lead to materials with lower dielectric constants.

2.2 Nonuniform Lens Design

Lower dielectric constants can be obtained by designing a lens in which the ε_r varies (discretely) across the lens boundary. To simplify the design, the lens is assumed to be a body of revolution. An analytically aided simulation approach is used; the lens boundary, dielectric constants of various layers, and thicknesses of the layers are first estimated analytically. Simulations are then used to examine the time of arrival of electric fields on a spherical surface to ensure that a spherical TEM wave originates from the first focal point. This procedure is iterated until the design is achieved within the desired tolerance. Two designs, planar and conical, that were explored are described below [11–14].

2.2.1 Planar

The planar lens is one in which the relative permittivity of the layers varies along the (rotational) axis of the lens. All layers have the same electrical length, i.e., the design is log periodic. 3-layer and 6-layer designs were investigated. Simulations indicated that the time spread in the rays originating from the source was over 100 ps. This time spread is highly undesirable as it indicates that the spherical wave front is distorted well beyond the acceptable tolerance of 20 ps.

2.2.2 Conical

In this design, the relative permittivity of the layers is a function of the polar angle, θ . Simulation results from a 7-layer configuration were almost identical to the planar design.

Due to the very large distortion of the spherical wave, the planar and conical designs were not iterated as this would have been too resource intensive. The large distortion in the spherical wave front could be due to many reasons, such as the

optical approximations breaking down at lower frequencies and reflections from inner layers. This approach was abandoned in favor of the simpler designs described in the following sections.

3 Investigation of Various Switch Configurations

In the design of the launching lens described above, it was assumed that the feed point and first focal point were spatially isolated. This assumption imposed a severe constraint on the lens design. The problem is greatly simplified if the geometric center of the feed point coincides with the first focal point. Before proceeding to the launching lens designs, various switch configurations are described where the geometric center of the switch cones is the first focal point. The focal impulse amplitude, $E_{\rm max}$, and beam width were compared for the following switch configurations [15]:

- 1. Four Feed Arms with Switch Cones (4FASC): A 200 Ω bicone switch, of height 1.0 cm, centered at the first focal point transitions to the feed arms of the PSIRA. The connection between the switch cone base and the feed arms is called the loft connection.
- 2. Truncated Four Feed Arms with Switch Cones (T4FASC): This configuration is identical to the 4FASC except that the feed arms are truncated at a distance of 19 cm from the first focal point.

One notes that the rise time of t = 100 ps corresponds to a physical distance of $ct_{\delta} = 3$ cm (in air). Therefore, a switch of appropriate dimensions can also serve as a guiding structure. Two such configurations were studied:

- 3. Vertical Bicone Switch (VBCS): A vertical bicone is used as the source where the height of each cone is 6.0 cm. This structure also serves to guide the spherical TEM waves originating from the feed point. The impedance of the cones was also varied.
- 4. Slanted Four-Cone Feed Arms (S4CFA): This structure is almost identical to the T4FASC except that the feed arms are replaced by 200 Ω cones. Each cone is 6.0 cm in height.

Results from numerical simulations for E_{max} and the spot size, at the second focal point, for the various configurations are summarized in Table 1.

One observes that the VBCS-75 and T4FASC designs are the most promising. Both these configurations are easy to fabricate and yield a relatively high peak electric field and an acceptable spot size compared to the other configurations in Table 1. Additionally, it is also found that the pre-pulse is shorter with these structures, compared to those in [7], which is desirable. Therefore, these configurations are further considered for the launching lens designs described in the next section.

Configuration	$E_{\rm max}$ (V/m)	Spot size (cm)
VBCS-75 VBCS-100 VBCS-125 VBCS-150 4FASC T4FASC	6.727 5.619 4.708 4.057 6.363 7.217	3.894 4.065 4.304 4.414 3.546 2.880
14FASC S4CFA VBCS Z denotes :	7.217 5.114	3.889 3.39 tch with impedance
	VBCS-75 VBCS-100 VBCS-125 VBCS-150 4FASC T4FASC S4CFA VBCS-Z denotes	Configuration E_{max} (V/m) VBCS-75 6.727 VBCS-100 5.619 VBCS-125 4.708 VBCS-150 4.057 4FASC 6.363 T4FASC 7.217 S4CFA 5.114 VBCS-Z denotes a vertical bicone switt

 $2Z \Omega$

4 Pressure Vessel Design

The gas chamber, pressure vessel, and launching lens are necessary as, in the final experimental realization of the PSIRA, 100 kV or more will be applied across the switch gap. The gas chamber typically contains hydrogen or SF_6 under high pressure. Hydrogen is used as an example in this chapter. As mentioned in the previous section, the location of the geometric center of the switch cones at the first focal point greatly simplifies the design. The design is further simplified if one considers the pressure vessel to serve the function of the launching lens. Then the switch system consists of only three components: (1) switch cones and guiding structures, (2) hydrogen chamber (HC), and (3) pressure vessel (PV). The design of the last two components is the focus of this section.

The pressure vessel may be cylindrical or spherical. The peak focal impulse amplitudes of the VBCS-75 configurations, with a spherical PV, were approximately 30 % lower than the corresponding T4FASC designs. Therefore, numerical simulation results for only the T4FASC configuration with the spherical pressure vessel (T4FASC-SPV) and cylindrical pressure vessel (T4FASC-CPV) designs are presented here [16, 17].

4.1 T4FASC with a Spherical Pressure Vessel

To provide structural support to the pressure vessel, a cylindrical support section (CSS), of height H_{css} , is added to the T4FASC design. The pressure vessel and the hydrogen chamber are both spherical. The relative permittivity of the pressure vessel medium is assumed to be 3.7. The impedance of the switch cones is 200 Ω in the pressure vessel medium, i.e., the cone half angle is $\theta = 45.58^{\circ}$. A spherical container, with $e_r = 2.25$ (transformer oil), surrounds the pressure vessel and is used to denote the oil "bath" that would be used in the final, practical design. Figure 2 shows the side view of the various components of the switch system. Dimensions of these components are summarized in Table 2.



Fig. 2 Side view of the T4FASC configuration with a spherical pressure vessel and spherical hydrogen chamber

Table 2 Dimensions of switch system components for the T4FASC-SPV design	Component	Height(cm)	Radius (cm)
	Switch cone	h = 0.5	0.51
	Cylindrical support	$H_{css} = 0.5$	0.51
	Hydrogen chamber	_	0.71
	Pressure vessel	-	$r_{\rm pv} = 2.0$
	Oil medium	_	$r_{\rm oil} = 5.0$

Probes to monitor the time of arrival of electric fields were placed on a (virtual) sphere of radius 10 cm (near field). The normalized responses from these probes are shown in Fig. 3 (each response is normalized with respect to its maximum). Please note that the E_{φ} component in the -yz-plane and the E_{θ} in the -zx-plane are not shown, as they are zero. The time spread in the electric fields is less than 20 ps, which is within the acceptable tolerance.

The focal waveform and spot size are shown in Fig. 4. These results indicate that the electric field is enhanced, compared to the T4FASC configuration in Table 2, with the pressure vessel by approximately 91 %, $E_{\rm max} = 13.76$ V/m, with a corresponding increase of only 19 % in the spot diameter, and with beam width = 4.64 cm. The enhancement in the electric field is due to propagation through the pressure vessel dielectric medium.



Fig. 3 Normalized E_{θ} and E_{φ} , components of the responses from the electric field probes on the *xy*-, *yz*-, and *zx*-planes for the T4FASC-SPV configuration. (a) Normalized E_{φ} in the *xy*-plane. (b) Normalized E_{θ} in the *xy*-plane. (c) Normalized E_{θ} in the *-yz*-plane. (d) Normalized E_{φ} in the *-zx*-plane

4.2 T4FASC with a Cylindrical Pressure Vessel

Analytical calculations using optical approximations for a T4FASC configuration with a cylindrical pressure vessel and cylindrical hydrogen chamber are detailed in [18]. It is shown that for a given relative permittivity, radius, and height of the pressure vessel, the surrounding oil medium can be used as the launching lens. Figure 5 shows the side view of the various components of the switch system.

The dimensions of the components are summarized in Table 3. The radius and height of the pressure vessel are determined using the formulas in [18] for $r_{\rm hc} = h$ tan, $\theta = 0.5$ tan, (45.58°) = 0.51 cm, and $\varepsilon_{\rm rll} = 2.25$; $\varepsilon_{\rm rll}$ is the permittivity of the launching lens. Note that $h_{\rm pv} = h + H_{\rm css}$ in Table 3.

The time of arrival of the electric fields on probes placed on a spherical surface, of radius 10 cm, showed a time spread similar to that of the T4FASC with a Spherical Pressure Vessel, i.e., less than 20 ps. The focal waveform and spot size are shown in Fig. 6. These results indicate that the electric field is enhanced, compared to the T4FASC configuration in Table 3, with the pressure vessel by



approximately 62 %, $E_{\text{max}} = 11.7$ V/m. The corresponding increase in the spot diameter is only 18 %, and beam width = 4.95 cm. Although the E_{max} is less than the T4FASC-SPV, the T4FASC-CPV configuration is attractive since it is easier to fabricate.

4.3 Optimization of Switch Components

Various components of the switch and pressure vessel design were optimized so that a large peak electric field, with a small spot size, was obtained. For the spherical pressure vessel configuration, the following ranges yield a reasonable electric field amplification:

- Feed arm length: 18 cm $\leq l \leq 22$ cm
- Height of cylindrical support structure: 0.2 cm $\leq H_{css} \leq 0.6$ cm
- Pressure vessel radius: 1.50 cm $\leq r_{pv} \leq 3.0$ cm



Fig. 5 Side view of the T4FASC configuration with a cylindrical pressure vessel and cylindrical hydrogen chamber

of switch system components for the T4FASC-CPV design	Component	Height (cm)	Radius (cm)
	Switch cone	h = 0.5	0.51
	Cylindrical support	$H_{\rm css} = 0.545$	0.51
	Hydrogen chamber	$h_{\rm hc} = 1.0$	0.51
	Pressure vessel	$h_{\rm pv} = 1.045$	$r_{\rm pv} = 1.905$
	Oil medium	_	$r_{\rm oil} = 5.0$

The height and radius of the pressure vessel for the cylindrical pressure vessel design are fixed by the formulas in [18]. The range of the feed arm length, l, for the T4FASC-CPV configuration is identical to the T4FASC-SPV design.

5 Conclusions

This chapter has described the design of a switch system to launch high-voltage (>100 kV), fast (100 ps) spherical TEM waves from the first focal point of a PSIRA. Spherical and cylindrical pressure vessel designs have been considered. The pressure vessel also serves as a (launching) lens.





The design of the switch system presented above is conceptually complete. The system can deliver <200 ps pulses with a spot area of approximately 2 cm², which is adequate for biological applications. However, there remain major technological challenges before a prototype device can be field tested. For example, one hurdle pertains to high-frequency switching. Numerical simulations used in the design of the switch system presented in this chapter have not considered the physics of gas discharges. For the switch designs considered, current technology limits the repetition rate to a few hundred Hz. Higher repetition rates, of the order of a few kHz, are required to effectively kill melanoma cells [19]. Hence, future work involves, amongst other venues of research, exploring techniques to increase the frequency of switch discharges.

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Part V UWB Interaction

EMI Risk Management with the Threat Scenario, Effect, and Criticality Analysis

F. Sabath

Abstract EMI risk assessment is a process for identifying, analyzing, and developing mitigation strategies for risks caused by electromagnetic interferences (EMI). The EMI risk assessment incorporates risk analysis and risk management, i.e., it combines systematic processes for risk identification, and evaluation, and how to deal with these risks. This chapter introduces a novel methodology for EMI risk assessment at system level, the Threat Scenario, Effect, and Criticality Analysis (TSECA). The TSECA is based on the general principle of the well-established Failure Mode, Effects, and Criticality Analysis (FMECA), which has been modified to include.

Keywords EMI • HPEM • Intentional EMI • Risk analysis • Risk assessment • TSECA

1 Introduction

The last decades have seen an increase in the vulnerability of civil society to interference caused by high-power electromagnetic (HPEM) environments. This is due to an increasing dependence on computer networks, wireless communication, microelectronics, and other sensitive electronic systems that can be easily disturbed by environments generated by HPEM sources. Today those HPEM sources range from advanced military sources to simple homemade devices.

Since the XXVI General Assembly in Toronto, 1999, the URSI adopted a resolution on criminal activities using electromagnetic tools [1]. The interaction of HPEM environments with essential electronic components and systems has been

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investigated in various papers and scientific reports [1, 2]. Some research groups focused on the susceptibility of electronic components and circuits to electromagnetic (EM) environments [3–9].

The comparison of results of those susceptibility investigations with the known capabilities of available HPEM sources highlights the necessity to protect critical systems and infrastructure against HPEM environments. The common way to specify protective measures starts by estimating the effects caused by the worst-case (strongest) EM environment. Identified effects are evaluated if they can be tolerated or if a protection is needed. In case protection measures are needed, they are specified with regard to the utilized worst-case EM environment. The application of this procedure on intentional EMI (IEMI) scenarios faces serious challenges, as the worst-case IEMI environment:

- 1. Depends on the particular characteristics of the target system
- 2. Is not the most likely EMI threat environment
- 3. Does not occur in real scenarios
- 4. Does not exist

Due to items 1–3, the estimation step yields an overestimation of the existing EMI risk and, as an immediate consequence, an oversized specification for the protection measures. As the protection measures are related to costs and extreme EM environments occur seldom, project management tends to accept the risk of EMI-induced failures and waive protection measures. As a consequence the whole system is left unprotected to any kind of IEMI environment that might occur more often that the considered worst case.

To prevent this unwanted situation, methods for a systematic risk management and assessment are required. The main goal of a systematic risk management must be the provision of adequate levels of functional safety and the avoidance of unreasonable high technical protective measures. A key element of such a systematic EMI risk management is the EMI risk analysis based on an analysis methodology that enables the identification of possible EMI-induced effects and the evaluation of its severity for the system performance is wanted.

This chapter is intended to introduce the concept of the Threat Scenario, Effect, and Criticality Analysis (TSECA), a novel risk analysis methodology. The TSECA is based on the general principle of the well-established Failure Mode, Effects, and Criticality Analysis (FMECA) [10, 11], which has been modified to analyze HPEM scenarios.

2 EMI Risk Management

The safety and reliability of critical infrastructure and their electronic subsystems are questioned when they are exposed to a high-power electromagnetic (HPEM) environment. In particular, cases in which the electromagnetic (EM) environment is originated with the intention to cause malfunctions or destructions in electronic

systems are in the focus of increasing public concern. Therefore methods for a systematic EMI risk management are required that provide adequate levels of functional safety and avoid unreasonable high technical protective measures and a high level of uncertainty about the functional reliability.

2.1 Introduction to Risk Management

The risk management is a systematic process for the identification, assessment, and prioritization of risks followed by coordinated and economical application of resources to minimize, monitor, and control the probability and impact of unfortunate events [12–14]. Principle and generic guidelines on risk management are provided by the ISO 31000 standards series [15].

According to ISO 31000 the process of risk management consists of the following steps:

- 1. Establishing the context
- 2. Identification of potential risks
- 3. Risk assessment

Once the context has been established and possible risks have been identified, the identified risk must then be assessed as to their potential severity of impact and to the probability of occurrence.

In the particular case of risks caused by the exposure to an EM environment, the establishment of the substitutes is as follows:

- The EMI risk is given by a malfunction or destruction of the system under consideration or its essential parts.
- Stakeholders are the designer, operator, and owner of the system on one site and the potential offender on the other site.
- Risks will be evaluated with regard to the desired or main function of the system under consideration.
- Possible mitigations or solutions are all operational and technical measures that reduce the coupling to the system or prevent that an EMI source can come close to the target system.

Generally, methods for risk identification can start with the source of our problems or with the problem itself. In the particular case of EMI risks, the source of problems is by definition set to the EM environment. Consequently it is highly recommended to utilize a method that starts from the EM environment as the source. After different EMI scenarios are created, consequences of the EM exposure (caused effects), their potential severity of impact, and the probability of occurrence must be assessed in the EMI risk assessment.

2.2 EMI Risk Assessment

The main objective of the EMI risk assessment is to provide qualitative and quantitative information about the risk that the exposure of the system under consideration to an EM environment causes failure or destructions as well as its consequences on system level. Therefore, EMI risk assessment answers three basic questions:

- 1. If the system under consideration is exposed to an EMI environment, what can go wrong with it, or which effects occur, that lead to adverse consequence(s)?
- 2. What and how severe is(are) the adverse consequence(s) that the system under consideration may be eventually subjected to as a result of the exposure to an EMI environment?
- 3. How likely will the adverse consequence(s) occur?

The answer to the first two questions requires technical knowledge of electromagnetic field propagation and coupling with systems as well as system performance and susceptibilities. The propagation and coupling of the EM field to the system under consideration are usually analyzed by appropriate analytical and numerical methods, e.g., topological analysis. Possible effects might be identified by checking the determined EM environment against known susceptibility data of subsystems or components.

Several industrial sectors, including aviation, space industry, and military systems, have adopted the Failure Mode and Effects Analysis (FMEA) to focus a reliability and safety analysis the most important initial effects. The FMEA is a bottom-up, inductive analytical method that examines the effect of a single-point failure on the overall performance of a system [10, 11]. Failure Mode, Effects, and Criticality Analysis (FMECA) is an extension of the process with the addition of a risk (criticality) assessment. The result highlights failure modes with relatively high probability and severity of consequences.

The answer to the third question can be obtained by employing Boolean logic methods like event tree analysis, event sequence diagrams, or fault tree analysis [16].

2.3 Benefits of EMI Risk Assessment

A proper implementation of an EMI risk assessment provides project management as well as the user with the following benefits:

- 1. A documented approach for selecting a design with a high probability of successful and safe operation in an EM environment
- 2. A documented uniform method of assessing potential failure mechanisms, failure modes, and their impact on system operation, resulting in a list of failure

modes ranked according to the seriousness of their system impact and likelihood of occurrence

- 3. Early identification of EMI caused failure, which may be critical to mission success and/or safety
- 4. An effective method for evaluating the effect of proposed changes to the design and/or operational procedures on the vulnerability to EM environments
- 5. Criteria for early planning of EMI hardening tests

3 Threat Scenario, Effects, and Criticality Analysis

In reliability engineering failure mode analysis is employed to identify redundancies and potential weaknesses, to discover interactions between elements of a given system, and to find appropriate measures to avoid failure propagation. A reliability and safety analysis tool that has been adapted by several industrial sectors is the Failure Mode and Effects Analysis (FMEA). The FMEA is a bottom-up, inductive analytical method that examines the effect of a single-point failure on the overall performance of a system. Failure Mode, Effects, and Criticality Analysis (FMECA) is an extension of the process with the addition of a risk (criticality) assessment. The result highlights failure modes with relatively high probability and severity of consequences.

The application of the conceptual procedure of the FMECA, as documented in [11], to analyze the risk of an EMI scenario results into a methodology, which consists of the following logical steps (Fig. 1):

- 1. Define the threat scenario.
- 2. Construct scenario interaction model and system structure model.
- 3. Determine effects and failure modes.
- 4. Evaluate each effect and failure mode and assign a severity classification category.
- 5. Identify failure detection methods.
- 6. Identify corrective measures for failure modes.
- 7. Document analysis.

In contrast to the (unmodified) FMECA, the modified methodology considers the impact of an EMI threat scenario on the performance of the system under consideration. Therefore, the analysis starts with the definition of the threat scenario to be analyzed. Consequently the (second) modeling step encompasses the electromagnetic coupling and the structural model of the target system. By limiting the analysis to effects that are caused by interaction of the target system with the EM environment, the determination of causes becomes obsolete. The described novel risk analysis methodology is termed the Threat Scenario, Effects, and Criticality Analysis (TSECA) [17, 18].
Fig. 1 TSECA procedure



3.1 Step 1: Define the Threat Scenario

The initial step of the TSECA consists of the definition of the threat scenario to be analyzed (Fig. 2). It is essential for the success of the analysis that the threat scenario definition covers all aspects of the HPEM scenario. The complete threat scenario definition includes possible HPEM source categories, alignment of the source, as well as the target system, environmental conditions (e.g., material properties), the architecture of the target system, identification of internal and interface functions, expected performance at all indenture levels, system restraints, and failure definitions. As depicted in Fig. 3, the aspects can be arranged into groups as:

- 1. HPEM source
- 2. Environmental topology
- 3. Target system

The group titled EMI Source contains information on possible EMI sources as:

- Source categories
 - Threat level or generated electromagnetic environment: The parameter *threat level* describes the ability of the HPEM environment to cause EMI effects. The *threat level* is perhaps the aspect of HPEM which has been mostly investigated and evaluated by the EMC community. The more general term



Fig. 2 Example EMI scenario



Fig. 3 Aspects of an EMI scenario

generated electromagnetic environment serves better the purpose of the risk analysis.

- Mobility: The parameter *mobility* summarizes the capability of an EMI source to come close to a target system. It compiles aspects like dimensions, weight, need of special supplies, integrability into platform systems or shelters, and ability to operate in motion.
- Likelihood of occurrence.
- Duration of exposure: The duration of exposure is the maximum time span over which the source would emit electromagnetic energy without being detected, stopped, or disabled by protective measures.

The group titled *Target System* contains all information on the system to be analyzed, which needed for the analysis:

- Structure of the system (including the lowest level that needs to be considered)
 - Interfaces
 - Interdependencies
- Main function and required performance
 - Definition of failures
 - · Operational modes
 - Mission phases
- Accessibility of the system
- Susceptibility to electromagnetic quantities
- Environmental profile (particular the electromagnetic environment)

The last group *Environmental Topology* contains all information on the topological setup and the propagation of the electromagnetic environment:

- Location of the target system.
- Possible locations of the EMI source: The occurrence of a given EMI source category is determined by its mobility as well as the accessibility of the location.
- Electromagnetic topology.

3.2 Step 2: Construct Scenario Interaction Model and System Structure Model

After the threat scenario is defined, a model that describes the propagation (coupling) of the electromagnetic environment, generated by the EMI source, into the target system, as well as a block diagram illustration of the operation and functional interrelationships of the target system must be constructed.

As illustrated in Fig. 4, the first part of the model, the scenario interaction model, is needed to describe the propagation of the electromagnetic environment,



Fig. 4 Interaction sequence diagram for a system exposed by an external HPEM source

generated by the EMI source, into the ambient of selected equipments or components of the target system. A structured approach to visualize and analyze the interaction of an EM environment with complex systems is the topological concept [19–21]. It permits subdividing a complex chain of interactions into a number of simpler parts. The topological analysis starts with the source EM environment and runs through a sequence of simpler steps propagating the EM environment (radiated and conducted coupling) until the component level of the target system is reached. Within each step the propagation of the EM field is characterized by measurements or described by appropriate analytical and numerical models. Often the EM environment is characterized by norms/quantities of interest, like the peak voltages, peak currents, peak electric and/or magnetic fields, the total signal energy, peak signal power, peak time integral, and rate of change of the aforementioned quantities. A topological decomposition for the example EMI scenario in volumes and networks is depicted in Fig. 5.

The second part of the model, the system structure model, consists of block diagrams which illustrate the operation, interrelationships, and interdependencies of functional entities of the target system. The structure model provides the ability for tracing failure mode effects through all levels up to the system level (Fig. 4). Both functional and reliability block diagrams are required to show the function flow sequence and the series dependence or independence of functions and operations.



Fig. 5 Topological decomposition of the example EMI scenario (volumes and networks)

Depending on the complexity of the threat scenario and the target system, the construct of the scenario interaction model as well as the system structure model can be comprehensive and time consuming. In order to adjust the effort to the expected result, it is mandatory to focus this step on the objective of the risk analysis. Particular in distributed, complex, and multilayered systems, a limitation of analyzed structure levels is desired. It is recommended to stop the decomposition of the target system at the highest sublevel that consists of elements (e.g., equipment, component) which provide deterministic or stochastic relations between the electromagnetic environment and caused effects (response).

3.3 Step 3: Determine Effects and Failure Modes

The equipment or component response is determined by evaluation of the HPEM induced environment with regard to the susceptibility data of the component. For example, a particular effect (failure) occurs whenever the applied EM environment exceeds the susceptibility threshold of the component. Several common effects include destruction, complete loss of function, uncontrolled output, and premature/ late operation.

Level	Criticality	Severity	Description
U	Unknown		Unable to determine due to effects on another component or not observed
0	No effect	Undisturbed	No effect occurs or the system can fulfill its mission without disturbances
1	Interference	Limited	The appearing disturbance does not influence the main mission
2	Degradation	Severe	The appearing disturbance reduces the efficiency and capability of the system
3	Loss of main func- tion (mission kill)	Very severe	The appearing disturbance that prevents the system is able to fulfill its main function or mission
4	Loss of system	Catastrophically	The appearing effect caused serious destruction or physical loss of the system

 Table 1
 Classification scheme for effects at system level (adapted from [22])

3.4 Step 4: Evaluate Each Effect and Failure Mode and Assign a Severity Classification Category

The consequences of each determined effect on component operation, function, or status shall be identified, evaluated, and recorded. Based on the block diagram, the consequence of an identified effect (e.g., output and behavior of a component) on the next system level shall be analyzed. All effects must be traced either through the top system level or the level at which impact on the system function/performance is compensated. The effect probability can be determined by appropriate methods (qualitatively or quantitatively).

Each effect shall be classified with regard to the severity of its consequence on the system performance. Suitable classification scheme for system-level effects are shown in Tables 1 and 2.

3.5 Step 5: Identify Failure Detection Methods

A description of the methods by which occurrence of the failure mode is detected by the operator shall be recorded. The failure detection means, such as visual or audible warning devices, automatic sensing devices, sensing instrumentation, other unique indications, or none shall be identified.

Level	Description	Individual item	Fleet/inventory	Quantitative
A	Frequent	Likely to occur often in the life of an item	Continuously experienced	Probability of occurrence greater than or equal to 10^{-1}
В	Probable	Will occur several times in the life of an item	Will occur frequently	Probability of occur- rence less than 10^{-1} but greater than or equal to 10^{-2}
С	Occasional	Likely to occur sometime in the life of an item	Will occur sev- eral times	Probability of occur- rence less than 10^{-2} but greater than or equal to 10^{-3}
D	Remote	Unlikely, but possible to occur in the life of an item	Unlikely, but can reason- ably be expected to occur	Probability of occur- rence less than 10^{-3} but greater than or equal to 10^{-6}
Е	Improbable	So unlikely; it can be assumed that occurrence may not be experienced in the life of an item	Unlikely to occur, but possible	Probability of occurrence less than 10^{-6}
F	Eliminated	Incapable of occurrence within t potential hazards are identifi	he life of an item.	This category is used when nated

 Table 2
 Failure probability levels [23]

3.6 Step 6: Identify Corrective Measures for Failure Modes

The compensating provisions, either design provisions or operator actions, which circumvent or mitigate the effect of the failure shall be identified and evaluated. This step is required to record the true behavior of the item in the presence of an internal malfunction or failure.

3.7 Step 7: Document Analysis

The TSECA shall be documented in a report that contains:

- Details of the scenario that was analyzed
- Assumptions made in the analysis
- The way the scenario was modeled
- The results
- The criticality and probability level of identified effects (failure modes)
- Any recommendations for further analyses, design improvements, or test plans

4 Conclusion

In this article a novel methodology for EMI risk assessment at system level, the Threat Scenario, Effect, and Criticality Analysis (TSECA), was introduced. The TSECA is based on the general principle of the well-established Failure Mode, Effects, and Criticality Analysis (FMECA). The output from TSECA includes a rating of importance based on the probability and severity of effects and adverse consequence(s) that result from an exposure of the target system to an EMI environment.

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On the Use of Probabilistic Risk Analysis for Intentional Electromagnetic Interference

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Abstract Due to the interconnection of modern electronic systems such as IT networks, it becomes impossible to predict the reliability of the whole system solely deterministically. The system needs to be decomposed in smaller elements which are easier to analyze by itself. Despite the functional structure of the system, also the electromagnetic topology needs to be taken into consideration. Critical system elements may be placed in shielded rooms where the deterministic calculation of electromagnetic fields is not reasonable because slight changes in this environment can drastically change the electromagnetic field structure in that enclosure. Furthermore, there are also a big variety of electromagnetic threats that the system can be exposed to. Due to these uncertainties, a complex system must be analyzed statistically. Hence, there is a need for a systematical analysis method which takes the uncertainties from different sources into consideration and which combines this knowledge in order to predict the risk. Moreover, the analysis should enable to identify the main contributors to the risk. The objective of this article is first to define the expression risk and second to present some of the aspects of probabilistic risk assessment and to show how those can be applied to IEMI problems.

Keywords Intentional EMI • Risk analysis • Critical infrastructure

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1 Introduction

Due to the complexity of modern electronic systems such as IT networks, it is impossible to predict the reliability of the whole system exposed to electromagnetic threats deterministically. There are uncertainties associated with the coupling paths, with the behavior of the victim and with the Intentional Electromagnetic Interference (IEMI) sources. One of the typical types of IEMI is an UWB pulse.

The system could be decomposed into smaller components which are easier to analyze by itself. However, the question remains: how to combine the behavior of single components in order to predict the risk for the whole system? For that, methods known from probabilistic risk analysis under functional safety aspects may be helpful.

The objective of this contribution is to introduce the expression risk and to define it for the case of IEMI. For that, the quantitative and qualitative consequences of IEMI are discussed briefly. Hereafter, the probabilities of the consequences are examined. The probabilities are decomposed into the likelihood of sources and probabilities of coupling and failure. Simulation results of random coupling to a generic system are presented.

In previous research electronic systems have been investigated as single systems. In this paper a measurement setup is shown in which the reliability of a system consisting of more than one element is examined. It is suggested to use fault tree analysis for the description of that system. The necessity of including the common cause failures into the description of the system is explained.

2 Definition of Risk

Often the expression "risk" is misleadingly used in EMC for describing the sources or the effects of the EMI. However, the most common definition of "risk" is the combination of the consequence of a threat and the probability of this consequence [1]. Mathematically the risk R can be expressed as a set of pairs:

$$R = \{C_i, p(C_i)\}, \quad i = 1, 2, \dots, N,$$
(1)

where C_i denotes the *i*th consequence and $p(C_i)$ its probability. Depending on the situation, a more quantitative definition of risk can be given:

$$R = \sum_{i=1}^{N} C_i \cdot p(C_i).$$
⁽²⁾

Using Eq. (2), risk can be represented as one number and thus different scenarios or systems can be compared easily. For example, that number could be a loss of a

certain amount of money per year. After having defined the risk generally, the next section describes how risk can be calculated in case of IEMI.

3 Probabilistic Risk Analysis in Case of IEMI

3.1 Consequences

First, possible scenarios and consequences have to be defined. As already mentioned, there are several ways to treat the consequences of an IEMI attack. On the one hand, consequences can be described in terms of financial lost. For that the system or mission needs to be analyzed with regard to financial consequences of different scenarios. Later, the causes and probabilities of these scenarios can be determined with the help of fault tree analysis, which is one possibility (see Sect. 3.5). The fault tree analysis is a deductive (top-down) approach. That means that starting with an undesirable event, the causes are determined.

Instead of describing the consequences quantitatively, it is possible and often easier to use qualitative measures. For that, the consequences can be classified. Depending on the level of the system different classification schemes can be applied. That means that for the description of the whole system, other classification might be more useful than for a component which is a part of the whole system. In [2] it was found that a very suitable scheme for classification at system level is the description by duration and criticality. After having determined all consequences of interest, the question arises how likely these consequences are.

3.2 Probabilities of Consequences

It is helpful to separate the probability of a consequence into the likelihood of the source which caused the consequence and the conditional probability of this consequence given the appearance of the source. Mathematically this can be expressed as

$$p(C_i) = p(S_i) \cdot p(C_i | S_i), \tag{3}$$

where $p(S_i)$ denotes the likelihood of the source S_i and $p(C_i|S_i)$ the conditional probability of C_i given S_i . If there are several sources leading to the same consequence, then Eq. (3) can be extended to

Fig. 1 The likelihood of IEMI source depends on its mobility and technological challenge



$$p(C_i) = \sum_{m=1}^{M} p(S_m) \cdot p(C_i | S_m).$$

$$\tag{4}$$

In Eq. (4), M is the number of all possible sources that are taken into consideration. Using Eq. (4), the probabilities of the sources and the corresponding consequences can be determined independently. Hence, the probabilities of sources could be calculated first and then the probabilities of certain consequences could be measured using the given sources. Therefore, the subsequent step is to determine the probability of each source and then the corresponding consequences.

3.2.1 Probabilities of Sources

For the calculation of Eq. (4), the likelihood of each source is needed. Since it is very difficult to detect if a disturbance happened because of an IEMI attack, no or only few data exists on the likelihood of different sources. Hence, the likelihood of the sources needs to be predicted in some other way. One possibility was presented by Sabath and Garbe [3]. They suggested that the likelihood of an IEMI source depends on the mobility and technological challenge of that source.

The smaller and lighter the source is, the higher is its mobility. The technological challenge combines quantities such as the level of knowledge needed to design or operate the system, availability of components, and costs. The lower the technological challenge and the higher the mobility of a source is, the higher is its likelihood (see Fig. 1).

This description of the likelihood does not provide absolute values of the probability or frequency (such as once in 5 years). However, a relative comparison can be done. It means that a statement can be done such as 20 % of IEMI attacks are made by a system similar to a JOLT [4].

There is ongoing research and development trying to identify IEMI attacks which can help to collect the data about the likelihood of IEMI attacks [5]. This data can help to update the expectation deduced from the prediction schemes presented by Sabath and Garbe. After determining the probability of a source, the next step is to determine the probability of consequences given the appearance of the source.



Fig. 2 Illustration of breakdown failure rate and destruction failure rate

3.2.2 Probabilities of Consequences for a Given Source

In previous research, the failure probabilities of single systems exposed to an UWB pulse were measured and predicted [6, 7]. For that, expressions such as breakdown failure rate (BFR) and destruction failure rate (DFR) were introduced. The result of the measurement produced curves similar to those in Fig. 2.

Through the use of mathematical norms [7], the breakdown threshold (the amplitude where the BFR reaches the 5 % level) can be predicted for other pulse shapes. This description of failure probability includes only a definite environment. The location and polarization of the source does not alter. Hence, there is a need to include somehow further effects such as the distance of the source to the system and the angle of incidence of the incoming pulse. For that the probabilities can be used again. Depending on the structure of the building or the fence around the system, a probability distribution for the distance from the source can be guessed. The next section describes the probability of coupling for random angle of incidence.

3.3 Simulation of Coupling for Random Angle of Incidence

Assuming that the BFR in Fig. 2 was measured for the maximum coupling case, the question arises: what would be the BFR if the angle of incidence is random and has a uniform distribution over a virtual sphere surrounding the system? In order to analyze this simulation, a two-conductor transmission line which can be assumed as a simplification of a more complex system was analyzed. The transmission line (length, 2 m) is excited by a double-exponential pulse, plane wave (rise time, 100 ps; full-width half-max, 2.5 ns) using random angle of incidence, and polarization. Then the current *I* flowing in one of the terminating impedances is calculated. Afterwards, the angle of incidence and the polarization are changed randomly and the current is calculated again. After a lot of times (e.g., 1,000), a dataset of different current peak values is available from which the empirical probability



Fig. 3 Probability density function of the peak current in a two-conductor transmission line excited by a double-exponential pulse from random angle and random polarization. Also shown is the fit to the beta distribution

density function (PDF) and the cumulative distribution function (CDF) can be plotted. The results are shown in Fig. 3.

Figure 3 also shows the fitting of a beta distribution to the empirical data. For that, the maximum-likelihood method was applied. The advantage of using the beta distribution is that the coupling to this transmission line can be described with a few parameters only and that the beta distribution is bounded to both of its ends.

3.4 Combination of the Breakdown Failure Rate and the Randomness of the Angle of Incidence

After describing the statistical coupling to a transmission line and the breakdown behavior in the worst case, the next step is to combine these two distributions. The result is the failure probability of a component given a certain amplitude of the incident pulse and assumed a random angle of incident. This probability $p(C_i|S_m)$ (probability of the *i*th consequence given the source S_m) can be calculated using the strength-stress model [8, pp. 155–157]:



Fig. 4 System under test consisting of two active components

$$p(C_i|S_m) = \int_0^\infty f_{BFR}(I_{BFR}) \left[\int_{I_{BFR}}^\infty f_{coup}(I_{coup}) dI \right] dI,$$
(5)

where $f_{BFR}(I_{BFR})$ is the derivative of the BFR. Camp [6] has shown that $f_{BFR}(I_{BFR})$ can be approximated by the Weibull distribution. The term $f_{coup}(I_{coup})$ denotes the probability density function of the coupled current I_{coup} shown in Fig. 3. Hence, Eq. (5) combines the BFR and the randomness of the angle of incidence. Similar approaches can be used in order to include the distance of the source to the victim and the attenuation through barriers.

The discussion in previous subsections concentrated on the BFR, which is a measure of the failure behavior of a single system. However, if there are multiple components which contribute to the system, then a method is needed in order to combine the component failure behavior to calculate the system behavior.

3.5 Measurement and Analysis of a Multicomponent System

In the previous research, a lot of work has been done in order to define and predict the failure behavior of single components. Normally, systems consist of multiple components. In order to analyze that, measurements on a system consisting of two microcontrollers shown in Fig. 4 were performed.

The system consists of two microcontrollers connected over the main board which serves as the power supply and also as the indicator of the proper working of the two-component system. The two microcontrollers make two LEDs on the main board blink. As soon as one of the microcontrollers fails, the other takes the control over both LEDs. Thus, the system is working redundantly.

In order to describe the behavior of that system, a fault tree can be constructed [9]. On the top of the tree, the undesired event is placed. That is the failure of the system. Then the causes for the failure are developed step for step. In that simple case, the fault tree is shown in Fig. 5a.

In previous research the components have been examined individually. From that the assumption might be made that the failures of the events are independent. However, our measurements have shown that depending on the wiring of the system, there is a strong dependency of the failures of the two microcontrollers.



Fig. 5 Fault tree for the two-microcontroller system. (a) No common cause failures are taken into consideration. (b) Common cause failures are included in the fault tree

Therefore, the "simple case" fault tree cannot be used for the analysis of such systems. In classical risk analysis, this type of failure is called common cause failures (CCF). There are several models that try to include the CCF into the overall model [10, pp. 140–152]. In order to include the common cause failures in our model, the fault tree can be extended as shown in Fig. 5b. It is part of the future work to predict the dependencies of the failures in interconnected systems.

4 Conclusion

This paper explained how to use probabilistic methods for describing the reliability of interconnected electronic systems exposed to IEMI. A way to describe the risk of that system is presented.

In order to determine the risk, consequences have to be defined. Those descriptions may be quantitative or qualitative. Next, the probabilities of the consequences need to be determined. It is suggested to separate the probability of a consequence into the likelihood of a source and the probability of this consequence given the appearance of the source. The likelihood of a source can be estimated using through the assessment of the mobility and technological challenge of the potential source. The probability of the consequence depends on several random factors such as the angle of incidence of the electromagnetic pulse, the distance from the source, or the reaction of the victim to this pulse. The results of a simulation of the random coupling to a transmission line are presented. In Eq. (5) it is shown how the breakdown failure rate known from [6] can be combined with the random angle of incidence. From that the failure probability of a component can be calculated for the case where the angle of incidence of the electromagnetic pulse is not known.

After the failure probabilities of single components are determined, fault tree analysis can be used in order to predict the failure behavior of the whole system. In the measurement setup a strong dependency of the two components building a whole system was observed. Hence, there is a need to describe and predict this dependency, known as common cause failures in the classical risk analysis. Predicting these common cause failures is part of the future work.

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Susceptibility of Electrical Systems to UWB Disturbances Due to the Layout of Exit Cables

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Abstract In this chapter, the effects of the exit cable layout variation for electrical large equipment under test (EUT) are analyzed. Susceptibility effects have been evaluated using antenna reciprocity theorem. The results show that slight variations of the cable layout cause drastic differences in the incident angle of the maximum susceptibility for high frequencies. At high frequencies, an EUT with a straight laid cable shows the maximum susceptibility compared to variations of the cable layout but only for a smaller spherical incident angle. The low probability of detecting this angle during radiated immunity testing with an antenna shows the advantage of using test alternatives with statistically uniformly distributed power (reverberation chamber). Moreover, it is shown that slight cable layout variations do not affect the results of this testing method.

Keywords Electromagnetic interference • Susceptibility • Layout variation

1 Introduction

The importance of susceptibility testing of electrical systems to ultra-wideband (UWB) disturbances has grown during the past years due to new communication standards and the rising threat of intentional electromagnetic interference (IEMI). Such tests shall evaluate the accurate functionality of the device in a worst case scenario. Hence, the electromagnetic illumination from the incident angle with maximum susceptibility during the test is very important. Most systems need power supply and data cables, which connect the EUT to the outside world (exit cable). Hitherto the effect of these cables has not been taken into account for high frequencies.

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Here, the influence of the layout of the cable on the susceptibility shall be analyzed by using numerical simulation techniques. Since only the disturbances that are coupled directly into the EUT can cause malfunctions, the antenna reciprocity theorem is used to evaluate the influence of cable layouts. The coupling port is substituted by a source and the radiation characteristics are analyzed.

Different cable layouts are compared to a straight cable layout and analyzed for two simulation models which are introduced in Sect. 2. In Sects. 3 and 4 the effects of the different cable layouts are discussed. The article closes with a summary of the cable layout effects on the EUT's susceptibility in Sect. 5.

2 Modeling

The simulations were carried out with the numerical software tool CONCEPT II [1], which is based on the method of moments (MoM). Figure 1 depicts two different models of the EUT above a perfectly conducting ground plane with varied cable layouts as follows:

- Setup A is used for a more general purposes. The variation of the cable layout is orthogonal to the ground plane.
- Setup B represents the typical situation of an electronic device on a desk. The layout of the horizontal cable segments is varied parallel to the ground plane.

The enclosure of setup A has a square base with a length of 8 cm for each edge and a height of 14 cm. The voltage source connects the perfect conducting enclosure and the copper wire. The wire is terminated with a load of 150 Ω , which is connected to an infinite, perfect conducting ground plane. The load represents the typical resistance value of an impedance stabilization network (ISN) or the insertion loss of a common mode absorption device (CMAD). The layout of the copper



Fig. 1 Dimensioning of the setups

wire consists of a superposition of a sine wave and an exponential function. These two functions facilitate a realistic layout, in order to compare the simulation results with measurements. The wire is laid in a straight line and in the shape of the modified sine wave of two periods with amplitudes of 3, 6, 9, and 20 cm. The height of the EUT is kept constant during the variation of the cable layout.

Setup B is modeled with the same components as setup A. The wire layout consists of several exponential and sine functions. The dimensions of the enclosure of setup B are 20 cm in length, 14 cm in width, and 7 cm in height. The EUT is located 80 cm above the ground plane. First, the wire layout of each wire section is a straight line. Then, the upper wire section has the shape of a sine wave with three periods and the lower section has the shape of a sine wave with five periods. The amplitudes of the sine waves are 3, 6, 9, and 15 cm. This scenario represents an EUT on a desk at the open area test site (OATS).

3 Pattern of Electrically Large EUTs

The purpose of susceptibility measurements is to ensure that the electronic device is not disturbed by any incident field of a given strength. Thus, the detection of the incident field angle with the largest susceptibility is indispensible. Figure 2 shows the absolute value of the radiated electrical far-field pattern of the three different simulation models. The most significant electrical field lobe indicates the angle of maximum susceptibility reciprocally.

As long as the EUT can be seen as electrically small (here at 90 MHz), the radiation patterns of all setups are similar to the pattern of a monopole above a conducting ground plane. Thus, the detection of the maximum susceptibility is easy. The EUT becomes electrically large as soon as the wavelength of the disturbance signal is less than the dimensions of the EUT (including the cable). Then, the detection of the maximum susceptibility is more difficult. The radiation patterns at 750 and 3,010 MHz are different in shape and value. All lobes become narrower with higher frequencies, and it is hard to predict the angle of their appearance.

Regarding the susceptibility, this implies that it is most likely to miss the critical incident angle for high frequencies. Thus, susceptibility measurements have to be carried out with small rotation angle steps in vertical as well as in horizontal polarization. Furthermore, an antenna height scan is necessary. Moreover, slight variations of the exit cable layout cause a completely different susceptibility behavior regarding the disturbances coupled into the cable.

This qualitative study is quantified in the next section. It is pointed out how a variation of the cable layout influences the quantity of the directivity, the total radiated power, and the maximum absolute electrical field strength compared to straight layout.



Fig. 2 Absolute electrical far-field pattern of three different setups at 90, 750, and 3,010 MHz

4 Characteristic Effects Due to Exit Cable Layout

As mentioned above, the antenna reciprocity theorem is applicable for the test setups in this contribution. Thus, the variations in susceptibility can be described with simulated radiation effects of the different setups. The deviations of the directivity, the total radiated power, and the maximum absolute electrical field of setups A and B with a straight cable layout compared to a cable with varied layout are presented.



Fig. 3 Directivity ratio of different sine-shaped cable layouts to a straight laid cable for setup A (*left*) and setup B (*right*)

4.1 Directivity

Figure 3 shows the directivity ratio of the varied cable layouts normalized to the directivity of a cable laid in a straight line. The frequency range is 30 MHz–4 GHz. The course of the directivity of setup A can be divided into three sections. In the first section the directivity of the cable with sine-shaped layouts is smaller than the directivity of a cable laid in a straight line. A shift of the first resonance due to the increasing length of the cable is detectable with rising amplitudes. The directivity of the alternative cable layouts increases in the frequency range from about 250–1,200 MHz. Above 1,200 MHz the directivity decreases for moderate amplitudes up to 10 dB except the appearance of a peak value at 2,000 MHz. The directivity for the highest sine amplitude exceeds the directivity of the cable with a straight layout. The frequency value of 1,200 MHz correlates approximately with half of the cable length of the test setup.

The directivity characteristics of setup B3 are similar to setup A. First, the directivity decreases above 800 MHz for moderate amplitudes of the sine-shaped cable layout. An increase of the values is observed in the frequency range from 1,300 to 2,500 MHz depending on the size of the sine amplitude. The second directivity maximum for smaller amplitudes appears towards higher frequencies. When the amplitude is increased, the second maximum is shifted to lower frequencies. Above 2,500 MHz the directivity of the sine-shaped cables is smaller than the directivity of the straight cable. Again the frequency value of 1,300 MHz correlates approximately with half of the cable length between the EUT and the first cable bend.

The analysis shows that for high frequencies the directivity of sine-shaped cable with moderate amplitude is in general smaller than the directivity of a straight cable. The effect appears for frequencies which corresponds with half of the length between the EUT and the location of the first significant cable discontinuity or smaller. Depending on the length of the cable (the sine amplitude), a peak in the



Fig. 4 Total radiated power ratio of different sine-shaped cable layouts to a straight laid cable for setup A (*left*) and setup B (*right*)

directivity ratio is detectable. The peak shows an increase up to 5 dB. This effect does not appear for frequencies which correspond with a quarter of the distance between EUT and discontinuity or smaller.

4.2 Total Radiated Power

Figure 4 depicts the total radiated power ratio of the sine-shaped cable layouts to the straight laid cable for setups A and B3. The power deviations of the sine-shaped cable layouts are up to 55 dB in the low frequency range, but they decrease toward higher frequencies for both setups. The course of the curves is almost identical above 1,500 MHz for setup A, even for high amplitudes. The value of the amplitudes does not influence the radiated power. There is only a deviation between the straight laid cable and the varied layout.

For setup B3 the radiated power deviations for moderate amplitudes are less than 5 dB above 1,600 MHz. The setup with the high sine amplitude shows increasing radiated power in the frequency range from 3,200 to 3,800 MHz. The plot of the linear regressions for each curve is 0 dB constantly. The deviation peaks appear at the same frequencies for the different amplitudes. Thus the effect does not depend on the length of the cable but may result from a less resonant behavior of the cable with the sine-shaped layout compared to the straight laid cable.

4.3 Maximum Electrical Field

According to [2], the directivity D_{max} is defined using the total radiated power P_{rad} and the maximum radiation intensity U_{max} as follows:



Fig. 5 Maximum absolute electrical far-field ratio of different sine-shaped cable layouts to a straight laid cable for setup A (*left*) and setup B (*right*)

$$D_{\max} = \frac{4\pi U_{\max}}{P_{\text{rad}}}.$$
 (1)

As shown above, the total radiated power shows minimal variations in the high frequency range. Thus, the maximum electrical far field can be expected to have almost proportional characteristics compared to the directivity. Figure 5 shows the decrease of the maximum absolute electrical far-field ratio for the two setups.

4.4 Consequences to Susceptibility Measurements

In contrast to the huge differences of the angle of maximum susceptibility in Sect. 3, the total radiated power is not affected by varying the cable layout. A way to avoid the huge measurement effort for high frequencies, described in Sect. 3, is to illuminate the EUT with statistically uniformly distributed power and to convert the power in equivalent field strength data to determine the maximum susceptibility. For radiated emissions, this technique is well described in [3-9]. The calculation of the maximum field strength and reciprocally the maximum susceptibility is based on the estimation of the maximum directivity D_{max} [4, 10]. The input variable for the estimation is the effective electrical size only. The layout of the cable is not considered for the calculation. As shown in Fig. 3, this layout has significant influence on the pattern and the coordinates of the angle of highest radiation and susceptibility respectively. Further investigations shall improve the estimation of D_{max} for cable systems and also predict the angle of maximum susceptibility. For worst case estimations, only the directivity of the straight cable layout has to be considered because slight variations cause smaller directivities for high frequencies.

5 Conclusion

Simulations of different EUTs with attached exit cables with different layouts were carried out. Firstly, the cable was laid directly between the EUT and the ground plane. Secondly, the EUT was located on a desk and the cable layout was modified accordingly. The layout of the cable was altered with the shape of a modified sine wave of different amplitudes. The susceptibility effects of different cable layouts were evaluated using the antenna reciprocity theorem.

The simulations show that slight variations of the cable layout cause drastic differences in the location of the incident angle with maximum susceptibility for high frequencies. Moreover, this incident angle becomes narrower with higher frequencies, and the detection of the direction with maximum susceptibility is very extensive and the probability of missing this direction increases. In contrast to that, the total radiated power is not affected by variations of the exit cable layout.

Immunity tests in the high frequency range with statistically uniformly distributed power illumination are immune to slight cable layout variations. These results with distributed power can be converted to field strength by use of the maximum directivity, which can be estimated for the system.

It has been shown that, for high frequencies, moderate variations of the cable layout cause smaller values of the maximum directivity compared to a straight cable layout. Hence, for worst case susceptibility examinations and for high frequencies, only the setup with straight cable layouts has to be evaluated as long as the angle of incident wave is not important.

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Breakdown Behavior of a Wireless Communication Network Under UWB Impact

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Abstract Systems with high priority to safety and reliability such as monitoring systems on airports have to work properly. Fast information transmission, continuous access to databases, as well as the management of air traffic are most important for effective and safe operation. Sources of Intentional Electromagnetic Interference can be manufactured relatively easy using commercially available components by civilian persons with relevant expertise and can be used for sabotage or blackmail purposes. For analyzing the weak points of a system existing on airports, it is necessary to reproduce its setup. In this investigation a UHF transmitter of a wireless communication device is developed and its breakdown behavior to unipolar fast rise pulses (UWB) is determined. A breakdown is a non-permanent damage, but includes a type of upset, that requires manual reset or at least stops communications for some period of time. The transmitter consists of three main components connected by data cables: power supply, microcontroller, and loop antenna. The immunity tests are accomplished as a function of the electromagnetic field direction to the device using an open TEM waveguide.

Keywords UWB • Breakdown • Network • Susceptibility • Critical infrastructure

1 Introduction

In this chapter, the breakdown behavior of the developed UHF transmitter (EUT) depending on the amplitude and the polarization of the external electric field and the cable length between the components is investigated.

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The first part contains the measurement setup such as the EUT and the test environment, followed by the electromagnetic effects and their characteristic quantity. The last two chapters describe the immunity tests and the observed effects as well as a short conclusion of the most important facts.

2 Measurement Setup

The following subsections describe the transmitter and measurement setup.

2.1 UHF Transmitter

The UHF transmitter consists of power supply with voltage converter from 9 V to 5 V, microcontroller, and loop antenna with 433 MHz transmitter. The components are connected by data cables. Figure 1 shows the process chart and schematic structure of the transmitter.

The system differentiates between two modes: the standby and activated mode. Compared to the activated mode, the standby mode is characterized by the operating microcontroller and the nonoperating state of the UHF transmitter; hence, no electromagnetic wave is transmitted by the loop antenna into free space. Light-emitting diodes show the current state of the system observing by fiber optic cables outside the test environment. The whole system realized on printed circuit board (5 cm \times 7 cm) is shown in Fig. 2.

In standby mode, the microcontroller does not actuate the UHF transmitter device; hence, no electromagnetic wave is transmitted via loop antenna. In this



Fig. 1 UHF transmitter. (a) Process chart. (b) Schematic structure



Fig. 2 UHF transmitter



Fig. 3 Standby mode

case, one diode switches on (LED III) and lights constantly, whereas the other LEDs (I and II) are not activated.

In case of activated mode, the UHF device transmits an electromagnetic wave via loop antenna by command of the microcontroller. The state is characterized by the two blinking LEDs I and II. The LED III, which lights constantly in standby mode, is switched off. Figure 3 shows the state in case of standby.

2.2 Test Environment

For immunity tests, the open TEM waveguide [1] and the pulse generator PBG3 for generating a double-exponential pulse (EMP) are used. The electromagnetic pulse is characterized by the following equation:

$$E(t) = E_0 \left(e^{-\alpha t} - e^{-\beta t} \right). \tag{1}$$

Figure 4 shows the characteristic of a double-exponential pulse.

The parameters α and β describe the pulse form and E_0 defines the amplitude of the pulse. Figure 5a, b shows the measured disturbing signal in time and frequency domain.

For analyzing the breakdown behavior of the device, it is not only necessary to change the cable length but also the polarization of the external electric field to the equipment under test. Figure 6 shows the open TEM waveguide and its main components.

Figure 7 shows the position of the system inside the test environment with its field directions. The wave propagation carries out in the z-direction.



Fig. 5 Disturbing signal in (a) time domain and (b) frequency domain



Fig. 6 Open TEM waveguide

Figure 7 demonstrates the direction of the E and B field to the device placed in the TEM waveguide. The wave propagation carries out in the z-direction. For the immunity tests, the waveguide of the Bundeswehr Research Institute for Protective Technologies and NBC-Protection in Munster (WIS) was used shown in Fig. 8.

For generating the double-exponential pulse, the pulse generator PBG3 was used. The following (Table 1) shows the characteristic of the generator.



Table	1	PBG:	3
charac	teri	stics	[2]

Output voltage	12 kV
Rise time (10–90 %)	100 ps
Pulse width	3 ns
Max. repetition rate	500 Hz

3 Electromagnetic Effects

The effects caused by disturbances coupling into integrated electronic devices range from bit errors to destruction effects in semiconductor devices [3]. In this case, effects such as bit errors and breakdowns can be observed during the immunity tests. The susceptibility of a system can be described by the breakdown failure rate [4]. Its characterization and definition are shown in Fig. 9.



Fig. 9 Breakdown failure rate (BFR)

4 Immunity Tests

The immunity of the UHF transmitter against electromagnetic threats was measured as a function of the field direction to the system, the electric field strength, and the cable lengths. For the investigations different parameters are defined:

- Reset cable RL: connection between power supply and microcontroller
- Data cable DL: connection between microcontroller and transmitter
- RL = 10 cm and DL = 10 cm (option 1)
- RL = 20 cm and DL = 10 cm (option 2)
- RL = 10 cm and DL = 20 cm (option 3)
- RL = 20 cm and DL = 20 cm (option 4)
- Electric field parallel to cable (vertical position)
- Electric field perpendicular to cable (horizontal position)

Figure 10 shows two different positions of the equipment under test inside the test environment.

The results of this experiment indicate that the polarization of the electric field relative to the cable and its magnitude produces different effects in the system under test. Some of these observed range from flashing with the pulse repetition frequency to the disruption of the general function of the system.

Figure 11 shows the breakdown failure rate in case of changing the reset cable length RL from 10 cm to 20 cm. The result is a decreasing breakdown threshold with increasing cable length which was observed in previous investigations [2]. In case of increasing data cable length, one may observe the same breakdown behavior. Also conspicuous is the steeply rising characteristic of the breakdown failure rate. In Fig. 12 all kinds of cable configurations for the horizontal polarization are shown.

Compared to the breakdown thresholds (BT) in case of horizontal polarization, the BT for vertical polarization is considerably less based on the field direction to the EUT and its data and reset cable shown in Fig. 13.



Fig. 10 EUT in (a) parallel and (b) perpendicular position to cable



Fig. 11 Breakdown failure rate. (a) RL 10 cm and (b) RL 20 cm

Compared to the measurement, results presented before the breakdown thresholds in case of perpendicular position of the system are less than for horizontal polarization. More interesting is the fact that the BT decreases considerably with increasing data cable length. Previous investigations and measurements demonstrated that particularly reset cables are susceptible for disturbing signals by reason that the incoming pulse generates state changes [5].



Fig. 12 Breakdown thresholds horizontal polarization



Fig. 13 Breakdown thresholds vertical polarization

5 Conclusion

In this chapter a UHF transmitter was presented and its susceptibility against electromagnetic threats (UWB) was determined. The breakdown behavior was investigated depending on the field direction and the length of the reset and data cable which was defined in Sect. 4. The results are different breakdown thresholds in case of vertical and horizontal polarization. Furthermore interesting is the increasing susceptibility of the EUT in case of changing the data cable length and the variable system states in horizontal polarization.

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The Technique for Evaluating the Immunity of Digital Devices to the Influence of Ultrawideband Electromagnetic Pulses

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Abstract The procedure for evaluating the immunity of digital devices to the influence of repetitive ultrawideband electromagnetic pulses is proposed. This procedure makes it possible to choose the optimal characteristics of radiated pulses, to estimate the effects of error-performance degradation and to predict the results of influence of pulses with arbitrary parameters, etc.

Keywords Ultrawideband electromagnetic pulse • Digital device • Immunity

1 The Content of the Immunity Evaluation Procedure

The procedure for evaluating the immunity of digital devices to periodically repeating ultrawideband electromagnetic pulses (UWB EMP) includes the following stages [1]:

- Testing the device, using one or more UWB EMP sources (with varying pulse peaks and pulse repetition rates, and recording the aftereffects, for instance, the degree of data transmission rate decrease, the moment of the Denial of Service etc.).
- Calculating the probability (*P*) of incorrect transfers of data packets for each combination of pulse peak and pulse repetition rate of the influencing pulses (with the UWB EMP characteristics, the parameters of microstrip lines of the PCBs of the digital device under consideration and the characteristics of the information signal taken as the initial data for the calculation).
- Development of the relation between the parameter characterizing the aftereffects of electromagnetic influence to the digital device and the probability of incorrect transfer of data packets.

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2 Mathematical Model of Forming Upsets in the Work of Digital Devices

In the process of functioning of digital devices between their separate blocks, an exchange of information takes place, which is transmitted by data buses in the form of packets of bits. If the information signal is superimposed by the pulse disturbance, the bit error may occur.

The probability of bit error $P_{e}(z)$ can be calculated with the help of the following correlation [2–4]:

$$P_{\rm e}(z) = 0.5 \left(1 - \operatorname{erf}\left(\frac{z}{\sqrt{2}}\right)\right).$$

Here, $z = U_s/U_d\sqrt{T_s/T_d}$; U_s is the amplitude of the pulse signal (of information bit); T_s is bit duration; U_d is the amplitude of pulse disturbance; and $T_d = (U_d)^{-2} \int (U_d(t))^2 dt$ is pulse disturbance duration.

The probability of the incorrect packet data (i.e., the probability that at least one of the bits may be incorrect) is

$$P = 1 - (1 - P_{e}(z))^{f \frac{N}{R}}.$$
(1)

Here, *f* is pulse disturbance repetition rate, *N* is the number of bits in the data packet, and *R* is the data rate $(R = 1/T_s)$.

The initial data required for the calculation of the formula (1) are the characteristics of information signals, as well as the characteristics of pulse disturbance induced in wiring of the electronic devices exposed to UWB EMP. The characteristics of information signals can be found in technical descriptions of the devices, whereas to determine the characteristics of pulse disturbance, one must perform calculations. For this purpose a calculation program based on the telegraph equations was developed. To perform calculations with the help of this program, one must know the characteristics of UWB EMP and the characteristics of wiring of the device in question.

3 Initial Data for Calculating Pulse Disturbances in the Elements of Wiring

3.1 Typical Characteristics of UWB EMP

Currently a great number of UWB EMP sources have been designed. Amplitude/ time characteristics of these sources vary in great range. When solving a number of tasks in assessing the immunity of technical systems, it turned out that it is





convenient to use the typical waveform of UWB EMP shown in Fig. 1 (shown in the figure amplitude/time characteristics are given as an example).

This pulse is characterized by the amplitude (E_1) and pulse width of the first halfperiod (T_1) , as well as the relation between similar characteristics of the first and second half-periods $(E_1/E_2, T_1/T_2)$. It is essential that $|E_1/E_2| = T_2/T_1$. In addition to the amplitude/time characteristics another important parameter is pulse repetition rate (f).

Given the capabilities of modern UWB EMP sources while performing calculations, it makes sense to consider the following characteristics: $E_1 = 0.5-100$ kV/m; $T_1 = 0.1-1.0$ ns; $|E_1/E_2| = 1-4$; $T_1/T_2 = 1-0.25$; f = 0-1 MHz.

3.2 Models of Microstrip Lines in Printed Circuit Boards

Three models were chosen which reflect the principal features of real Printed Circuit Boards (PCBs).

Model 1 simulates a symmetrical two-wire line above the conductive plane of board. Model 2 simulates the asymmetric two-wire line, one wire of which is connected with the conductive plane of board. Finally, model 3 simulates a single-wire line above the conductive plane of board. These models are shown in Figs. 2, 3, and 4.

It is necessary to tell that models 1–3 do not exhaust all possible variants of interconnections in electronic devices. In the paper we have used only these models to illustrate a technique. Further we plan to expand the list of typical models and their characteristics.



Fig. 4 Model 3



As an example, with the help of the above-described calculation method and suggested models of microstrip lines in PCBs, characteristics (amplitude and energy) of electric interference were calculated. The calculations were made with the help of the following initial data on geometrical and electrical parameters of the mentioned lines, as well as the characteristics of information signal: the line's length, 10 cm; the line's width, 1 mm; the distance between the adjacent lines, 1 mm; the thickness of PCB, 2 mm; the relative dielectric constant of the material of the board, 4.7; the load resistance of signal lines, 120 Ω ; the signal amplitude, 1 V;



Fig. 2 Model 1

The Technique for Evaluating the Immunity of Digital Devices...

Initial data on characteristics of UWB EMP: $E_1 = 5$ kV/m; $T_1 = 0.5$ ns; $ E_1/E_2 = T_2/T_1 = 1$; $f = 1$ kHz			
Characteristics of disturbance	Model 1	Model 2	Model 3
$U_{\rm d},{ m V}$	~0	20	26
W _d , J	~0	2.8×10^{-9}	3.6×10^{-9}
P, %	~0	3.4	3.7

Table 1 The dependence of characteristics on the type of the microstrip line model





the data packet length, 1,000 bits; and the data rate, 10 Mbits/s. The results of calculation are shown in Table 1.

It is easy to see that pulse disturbances in model 1 practically are absent and in models 2 and 3 are approximately identical and reach 20–26 V. We will use model 3 at carrying out of the subsequent calculations.

Then, experiments have been executed. The methodology of these experiments consisted in the following (Fig. 5). Two personal computers were integrated in a local computer network. The first computer (PC1) was exposed to repetitive UWB EMP. The second computer (PC2) was placed in a shielded room and was therefore not exposed to electromagnetic pulses. The length of communication line between computers was equal to 7 m and the length of line exposed to UWB EMP radiation was equal to 1 m.

Two types of UWB EMP sources were used in the course of the experiments. The amplitude/time characteristics of pulsed fields formed by them are represented in Figs. 6 and 7 (FOM = $E \times D$, where *E*, the amplitude; *D*, the distance).

During experiments the variation of characteristics of UWB EMP that influenced on PC1 was carried out. In addition, for each combination of values E and f, the calculation of probability P was carried out.

To determine the result of the influence of UWB EMP on PC1, the measurement of the data rate (R) via the local computer network was performed. As an example, Fig. 8 shows the results of measurement of the parameter R during the irradiation of PC1 with the help of UWB EMP source No1, which has the amplitude of 2 kV/m. The pulse repetition rate varied in a range from 0.5 to 10 kHz. It is clearly seen that as the result of UWB EMP influence there is a reduction in the data rate in the

Fig. 6 Pulsed electric field generated by the source No. 1



Fig. 7 Pulsed electric field generated by the source No. 2

Fig. 8 The changing of the

data rate under the influence

of UWB EMP generated by

the source No. 1:

E = 2.0 kV/m

network up to zero, i.e., Denial of Service (DoS). The occurrence time of this event (T_{DoS}) decreases as the pulse repetition rate increases.

As a result of the experiments and calculations, basic data were obtained, which are essential for development of the dependence of parameter T_{DoS} on probability *P*. This dependence is shown in Fig. 9.



The obtained dependence combined with the method of calculation of probability P allows to estimate the effect of the influence of repetitive UWB EMP with arbitrary characteristics on digital devices.

As an example of practical application of the proposed methodological approach, let's consider the following problem.

Let's suppose we want to design a source that generates UWB EMP with the amplitude of 5 kV/m on a predetermined distance and which ensures maximum influence on the personal computer similar to the one discussed above.

Clearly, this task comes down to finding such a combination of parameters of emitted pulses, at which the maximum value of parameter P is achieved. For definiteness we assume that the emitted pulses have the form, shown in Fig. 1. In addition we must take into consideration that the increase of efficiency of the source, thanks to the increase of average power, will result in increase of size and weight. In connection with this we will introduce a restriction on the parameter T_1f . In case of fixed amplitude of emitted pulses, it is equivalent to the restriction of average power of emission. Experiments have shown that in the case of designing a compact transmitter, the value of the mentioned parameter should be equal to approximately 5×10^{-7} .

Thus, it is necessary to solve the problem of finding the maximum of the function $P(T_1,|E_1/E_2|,f)$ at following restrictions: $E_1 = 5 \text{ kV/m}$; $T_1f = 5 \times 10^{-7}$. The results are shown in Fig. 10. The figure shows that the maximum efficiency is achieved at $|E_1/E_2| = 1$. This implies that when designing UWB EMP sources, it is



necessary to seek this value of the parameter. However, this is not an easy task. In reality, the sources with parameter $|E_1/E_2| \sim 2-3$ are more common. It is clear that in this case, the maximum probability of incorrect transfer of data packets is achieved at $T_1 \sim 0.25$ ns (respectively when $f \sim 2$ kHz) and is equal to approximately 4–5%. Having looked at Fig. 9, we can easily determine that the effect of influence of repetitive UWB EMP with such characteristics will be almost immediate DoS of the computer network.

Besides, we can consider another problem, which can be solved using the proposed method. By varying the amplitude of influencing pulses, we can define its minimum value at which we observe the immediate DoS as well as the maximum value at which the normal functioning of the personal computer is maintained. The results of calculation are shown in Fig. 11. It is clearly shown that the level of DoS is about 4 kV/m and the level of computer immunity is about 2 kV/m.

5 Conclusion

We proposed the method for evaluating the immunity of digital devices to the influence of repetitive UWB EMP. This method includes a procedure with the following steps:

- Testing devices using the existing sources of UWB EMP. At this stage, the amplitude and the pulse repetition rate vary, and simultaneously the effects of their influence are registered, for example, the degree of decreasing the data rate, the occurrence time of DoS, etc.
- Development of models of microstrip lines in PCBs of the devices in question and calculating the probability of incorrect transfer of data packet for each combination of amplitude and repetition rate of UWB EMP.
- Construction of the dependence of the parameter, characterizing the effects of electromagnetic influence on the device, depending on the probability of incorrect transfer of data packet.

The result of carrying out the above procedure is an opportunity to solve the following problems: to find the combination of characteristics of UWB EMP source, which ensures the most effective mode of influence of the emitted pulses on digital devices; to predict the results of the influence on such devices of electromagnetic pulses with arbitrary characteristics; etc.

It is obvious that, as well as any technique, our technique has some limitations. These limitations are caused by application of rather simple models of interconnections in electronic devices. Other reason of the limitations is application of the simplified approach to estimating the possible aftereffects of influence of electromagnetic pulses on digital systems. This approach does not allow considering the pulse disturbances that duration exceeds duration of bit of data, the features of various methods of coding signals in digital devices, etc. We plan to eliminate these limitations in a following variant of the technique.

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Reciprocity Theorem: Practical Application in EMC Measurements

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Abstract This document reports part of the measurements developed under the project entitled "Evaluation of the reciprocity in EMC measurements." The project aims to evaluate the technical feasibility of the implementation of the reciprocity principle in EMC-EMI measurements.

Keywords Reciprocity theorem • Shielding effectiveness assessment

1 Introduction

The assessment of the shielding effectiveness (SE) of a facility is usually performed by measuring the transfer function between an external illuminating field and the measured fields or currents inside the facility under test; see, for example, [1].

In general, high power RF is injected into an antenna outside the facility, in order to produce the illuminating field. The resulting fields, currents, or voltages are measured inside the tested facility using a proper probe. However, in some cases high-amplitude illuminating fields could disperse in the vicinity, affecting surrounding equipment and infrastructures.

The principle of reciprocity could be used in order to avoid the unwanted illumination of surrounding infrastructures: the idea consists in injecting high-

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amplitude electromagnetic fields inside the tested facility and measuring the leaked fields in a given direction outside the facility, taking advantage of the fact that large RF fields are much more confined inside the facility under test.

A series of tests evaluating the practical applications of this theorem were performed. Parts of the results obtained are presented in this document.

2 Transimpedance Function

In the context of electromagnetic fields, and circuit networks, the concept of reciprocity describes [2]: "...the reversibility of an interaction, upon interchange of the source and target."

The concept can be exemplified using a 2-port network, which can be represented by its *Z* matrix in the form

$$V_1 = Z_{11}I_1 + Z_{12}I_2,$$

$$V_2 = Z_{21}I_1 + Z_{22}I_2.$$
 (1)

Reciprocity on this network implies that the transimpedance functions are equal $(Z_{12} = Z_{21})$; this means

$$Z_{12} = \frac{V_{1}^{\text{out}}}{I_{2}^{\text{in}}}\Big|_{I_{1}=0} = Z_{21} = \frac{V_{2}^{\text{out}}}{I_{1}^{\text{in}}}\Big|_{I_{2}=0},$$
(2)

where I_N^{in} and V_M^{out} mean that we consider the input of the system is a current at port N and the output is a voltage at port M. Equation (2) states that the relationship between input currents and output voltages is the same whether we are going from port 2 to port 1 or vice versa. The current and voltage transfer functions could also be reciprocal, but only if the conditions $Z_{11} = Z_{22}$ are satisfied.

3 Reciprocity Criteria in Practical Implementations

In this study we evaluate the reciprocity of the measurement based on practical criteria, rather than the ideal conditions defined by Eq. (2), as follows:

Amplitude criterion: we consider that the system is reciprocal if the difference between reverse and direct transimpedance is less than 6 dB:

$$\left\| 20 \text{Log}_{10} \left(\frac{\|Z_{12}\|}{\|Z_{21}\|} \right) \right\| \le 6 \text{ dB.}$$
(3)

Phase criterion: in terms of phase, we consider that the system is reciprocal if the absolute phase difference (APD) is less than 20° :

$$APD = |\theta_{21}(f) - \theta_{12}(f)| < 20^{\circ},$$
(4)

where θ_{12} and θ_{21} are the phases of the Z_{12} and Z_{21} , respectively.

If at a certain frequency the conditions defined by Eqs. (3) and (4) are fulfilled, it can be said that the system is reciprocal.

4 Report of the Measurements

4.1 Example 1. Two Antennas in Near-Field Range

The first example consists of two antennas in the near-field range. The system, described in Fig. 1, consists of two biconic antennas (Antenna 1, Antenna 2) separated by a distance R (2 m approximately). A unidirectional fiber optic link connects Antenna 1 to the Vector Network Analyzer (VNA). The direction of the fiber optic link can be reversed, producing the two configurations under study. In the first one, here called the *direct case*, illustrated in Fig. 1a, the signal flows from Port 1 to Port 2. In the *reciprocal case* (Fig. 1b), the signal flows from Port 2 to Port 1.

In order not to change the topology of the problem when the optical link is reversed, both fiber optic transceivers are battery powered. The measured *S*-parameters for the direct and reciprocal cases are presented in Figs. 2 and 3.

As it can be noticed in Fig. 2, there is good adaption at the entry of the fiber optic link (the electro-optic converter), $S_{11} < -29$ dB. The S_{22} on the same figure represents the reflection coefficient of the Antenna 2 that can be considered



Fig. 1 Example 1. Direct case (**a**) and reciprocal case (**b**). The system consists of a VNA, a fiber optic link, and two biconic antennas in the near-field range. The transceivers of the F.O link are battery powered, in order not to change the topology of the system when inverting the link



Fig. 2 Example 1. Magnitude of the S-parameters direct case



Fig. 3 Example 1. Magnitude of the S-parameters, reciprocal case

acceptably matched from 600 MHz up to 900 MHz. The S_{21} in this figure represents the link between Port 1 and Port 2. As it can be seen, there is not much signal coupling at low frequencies. The S_{12} represents the coupling from Port 2 to Port 1; as expected the S_{12} is near the noise floor (around -110 dB), as this is the reverse sense of the unidirectional fiber optic link. In Fig. 3 the reciprocal case is presented.



Fig. 4 Example 1. Z_{21} , direct, in *red* and Z_{12} , reciprocal, in *blue*



Fig. 5 Absolute phase difference between direct and reciprocal transimpedances, Example 1

The S-parameters were transformed to Z-parameters using the formulas presented by Pozar in [3]. The magnitude of the transimpedance for the direct and reciprocal cases is presented in Fig. 4. It can be seen that there is close agreement between these two parameters.

The calculated APD is presented in Fig. 5. It can be said that phase difference between Z_{21} direct and Z_{12} reciprocal exists at some frequencies. However, the 80 % of the frequency points plotted in Fig. 5 have an APD < 20°. It can also be observed that a high APD appears at frequencies where the amplitude of the transimpedance is small. This is illustrated in Fig. 6, where the amplitude of Z_{21} direct and Z_{12} reciprocal is presented again; the zones in blue (here called coincidence zones) correspond to the frequencies where both criteria are fulfilled. As it



Fig. 6 Direct and reciprocal transimpedances, Example 1. The zones in *blue* indicate the frequencies at which the absolute phase difference is less than 20°

can be seen, the blank zones are coincident, in most of the cases, with low values of transimpedance.

As a conclusion of this example, it can be said that in spite of the use of a unidirectional fiber optical link, reciprocity holds acceptably well. There is suitable agreement between the direct and reciprocal transimpedances along the frequency range comprised between 100 MHz < f < 1 GHz. The cases where reciprocity doesn't hold are highly correlated with low-amplitude transimpedances, which suggests contamination with noise.

4.2 Example 2: Measurements in a Faraday Cage

The topology studied on this example is presented in Figs. 7 and 8. In the direct case, the field is generated inside the cage, and the field leaked outside the cage is measured with a biconic antenna. In the reverse case, an illuminating field is produced outside the cage, and the leaked fields inside the cage are measured by a dipole antenna.

A leakage wire, traversing one of the panels of the cage, was introduced on purpose, in order to make detectable the level of signal measured inside the cage.

Figure 9 presents the direct and reciprocal transimpedance functions and the coincidence zones. There exists a high degree of similarity between the amplitudes of the direct and reciprocal transimpedances, for the frequency band starting at 100 MHz. However, numerous blank zones appear, indicating an APD $> 20^{\circ}$, as can be seen in Fig. 9.



Fig. 7 Example 2. Direct case



Fig. 8 Example 2. Reciprocal case

In order to determine the degree of similarity between amplitudes and phases of Z_{21} direct and Z_{12} , we can apply the cross-correlation operator defined as [4]

$$\operatorname{Corr}[h(f), g(f)] = \int_{f_1}^{f_2} h(f)g(f) \mathrm{d}f.$$
 (5)

When applied to the amplitude and phase of the transimpedance, we have



Fig. 9 Direct and reciprocal transimpedances, Example 2

$$\operatorname{Corr}[|Z_{21}(f)|, |Z_{12}(f)|] = \int_{f_1 = 300 \text{ KHz}}^{f_2 = 1 \text{ GHz}} |Z_{21}(f)|, |Z_{12}(f)| df = 0.99,$$

$$\operatorname{Corr}[\theta_{21}(f), \theta_{12}(f)] = 0.84.$$
(6)

As it can be seen, the correlation operator applied to the amplitudes leads to a correlation coefficient near to 1, indicating a high degree of similarity between amplitudes of the direct the reciprocal transimpedances.

On the other hand the correlation coefficient is much lower for the phase. It's difficult to give a precise explanation of this. In principle, it could be argued that noise contamination is affecting the measurement; however, we don't see how the noise is affecting the phase of the transimpedance and not the amplitude.

Another explanation, perhaps more plausible, is that the topology of the experiments was slightly changed when inverting the direction of the fiber. A different topology could have generated different paths for the signals traveling between the antennas, changing the time of arrival and the phase.

4.3 Example 3: Cable Illumination

A setup involving illumination of cables is described in Figs. 10 and 11. In the direct case an antenna connected to the Port 1 of the VNA illuminates a RG-58 coaxial cable. A 50 Ω matching resistor is added at one end of the coaxial. The signal induced by the illuminating field is sent to the Port 2 of the VNA using a fiber optic transceiver connected to the other end of the coaxial. The external conductor



Fig. 10 Example 3. Direct case



Fig. 11 Example 3. Reciprocal case



Fig. 12 Direct and reciprocal transimpedances. Example 3

of the cable was pierced on purpose in order to make it more susceptible to the external field.

In the reciprocal case the fiber optic link is reversed. The signal produced at Port 2 of the VNA is injected in the coaxial cable and the field leaked by the cable illuminates the antenna connected to Port 1.

Figure 12 presents the direct and reciprocal transimpedance functions. As it can be seen, reciprocity holds in this case, on a large band of frequencies. Again, the blank zones are correlated to low-amplitude values.

On this case it was observed that the reciprocity holds, even at frequencies lower than 50 MHz.

5 Conclusions

In this study the application of the theorem of the reciprocity in EMC-EMI was discussed from a practical point of view. The reciprocity in the transimpedance function of several systems was evaluated and measured. A criterion of evaluation consisting of 6 dB range of agreement in amplitude and 20° of difference in phase between the direct and reverse transimpedance was defined.

A well-adapted unidirectional fiber optic link relying one end of the system under tests and the VNA was used. The fiber optic link was battery powered at both ends in order to preserve the topology of the system when the direction of the fiber was changed.

In the case of the antenna tests, performed in open space, the direct and reciprocal transimpedances fulfilled the defined criteria, in most of the evaluated frequency band. The frequencies where this wasn't observed are generally correlated to low-amplitude transimpedances. This suggests that at those precise frequencies, noise affected the signal.

When one of the antennas was placed inside a Faraday cage, it was found that direct and reciprocal transimpedances still satisfied the 6 dB criteria; however, an absolute phase difference higher than 20° was found in a significant number of frequencies. We hypothesize that the topology of the experiments was slightly changed when inverting the direction of the fiber. A different topology could have generated different paths for the signals traveling between the antennas, changing the time of arrival and augmenting the phase difference between the direct and reciprocal signals.

In the case of the tests performed in cables, it was found that the criterion of reciprocity was satisfied in most of the cases. This means that the transimpedance function can be calculated illuminating the cable and measuring the induced current or voltages or by injecting current or voltages on the cable and measuring the fields radiated by the cable with an antenna.

We saw that these measurements can be affected by noise. One way of preventing this is using directional antennas that can be oriented in the direction of arrival of the signal, preventing the arrival of ambient noise from unwanted directions.

Shielding the measurement setup is highly recommended (VNA and cables). This can also reduce the noise influence on the measures.

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Coupling of Hyperband Signals with an Underground Cable

K. Sunitha, M. Joy Thomas, and D.V. Giri

Abstract This is an adaptation of Interaction Notes 612. The Impulse Radiating Antennae (IRA) find immense applications in the military and civilian domains. These antennae are of reflector-type geometry. The interference caused on underground cables due to electromagnetic fields from IRA is an interesting subject of study. The determination of the induced current and voltage on a cable when it gets illuminated by an IRA will give the intensity of the interference. Prior to the computation of induced parameters, the radiated electromagnetic fields from an IRA are determined both in the near-field and in the far-field regions.

Keywords Impulse Radiating Antenna • Electric field • Coupling • Cables • UWB

1 Introduction

Impulse Radiating Antennae have a wide variety of applications such as the detection of buried objects, ionospheric research, and target discrimination in a cluttered environment [1, 2]. The hyperband electromagnetic fields [3] coming out of the IRA can interfere with the objects in the path where it travels. This can be an object laid underground or any object on or above the ground. Since the communication cables and the buried data cables are connected to sensitive equipments at their terminals, even a slight rise in the voltage or the current at the terminals of the

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equipments can become a serious problem for the smooth operation of the system. In this aspect it is worthwhile to determine the effect of an IRA's electromagnetic field on the cables laid underground.

Treating this as an electromagnetic compatibility problem, the victim circuit becomes the cable; the source of interference being the electric field from the IRA and the air and soil forms the path of propagation of the interference signal to the victim circuit. As a first step to this computation, we need to know the exact nature of the electromagnetic field from an IRA at different distances from the antenna. Using the concept of the reflection and transmission of the field at the earth's surface, the magnitude of the field getting coupled to the cable can be found. This in turn can be used for the determination of the induced current and voltage on the cable which give an exact measure of the interference on the cable. This work deals with the coupling effect of the IRA fields on buried telephone cables.

2 Impulse Radiating Antenna

A parabolic reflector-type IRA [4] has been considered as the source of interference which is shown in Fig. 1. The reflector is fed by a transverse electromagnetic wave structure energized by a ± 60 kV HV pulser source. The detailed description of the source is available in [2, 4, 5]. For the pulser used in this work, the specifications are pulser voltage, $V_0 = 120.72$ kV, rise time = 100 ps, and maximum rate of rise of voltage = 1.2×10^{15} V/s. The peak amplitude of the voltage waveform is slightly less than V_0 . With 120.72 kV pulser voltage, the peak amplitude turns out to be 120 kV. The details of the pulser and the antenna assembly are described in detail in [6]. The voltage from the pulser is shown in Fig. 2. This voltage is used for the computation of the electric field and hence for the determination of the induced current and voltage on the cable.

The electric field from this IRA at any observation point can be found out by integrating over the aperture the equivalent magnetic currents which is explained in [7]. Using this concept, the electric field is computed along the boresight at



Fig. 1 A parabolic reflector-type IRA



Fig. 2 Output voltage of the HV pulser. (a) Frequency domain. (b) Time domain



Fig. 3 Electric field at different points along the boresight in frequency domain

distances of 5, 10, 20, 100, and 304 m from the IRA. Figure 3 gives the frequency domain waveform of the electric field at these observation points which cover the whole range from the near field through the intermediate field and to the far field. This figure shows that the spectrum becomes oscillatory at higher frequencies in the near field, but in the intermediate and the far fields the fields have less oscillation.



Fig. 4 Electric field at different points along the boresight in time domain

Figure 4 represents the time characteristics of the electric fields at the same points, i.e., 5, 10, 20, 100, and 304 m. It can be seen that whatever be the distance, the electric field has a negative portion (prepulse) up to 8 ns (=2 F/c, with F and c being the focal length of the reflector and speed of light in air, respectively), followed by its impulsive transient nature. This 8 ns accounts for the time taken by the pulse to travel to the reflector's surface and get itself felt at the source point. Till then there is a direct radiation of the feed pulse at the observation point, which is negative in nature on account of its reverse travel. The 8 ns prepulse time is common for all the observation points with far fields getting a lesser magnitude of the pulse and near fields getting a larger negative portion. The area under the negative and the positive portions of the wave has to be equal, which is reflected in the relative magnitudes of the prepulse and the impulsive parts. Hence, the magnitude of the field shoots to a maximum at around 20 m where the observation point is in the near field. Here the peak negative value of the field intensity is 996 V/m, and the positive value of the field intensity is 27 kV/m. At 5 m, the respective figures are 3.6 and 25 kV/m. In the intermediate field, i.e., at 100 m, these figures come to 284.5 V/m and 13 kV/m, respectively, and in the far field, i.e., at 304 m, these are 90 V/m and 4.951 kV/m, respectively.

3 IRA Illuminating the Cable

This section considers an underground cable getting illuminated from the electric fields due to IRA. This situation is portrayed in Fig. 5. The IRA is located at a height of 5 m above the ground. The cable is buried at a depth of 1 m below the ground and is excited by the electric field generated by the IRA. The earth plane is considered to be of infinite extent. The cable is buried such that the boresight of the IRA meets the midpoint of the cable. In this work only direct radiation from the IRA is considered. The cable is terminated at both ends by matching impedances. The conductivity of the soil is taken as 0.001 S/m. The relative dielectric constant of the soil is assumed to be 10. The cable used is a shielded communication cable. The skin effect and proximity effect can be neglected for the incident large field amplitude, as the cable is in the near field. Both the shield and the inner conductor are terminated, respectively, by their own characteristic impedances as shown in Fig. 5. It is clear that the nature of the termination is critical in the evaluation of the induced parameters. In this case, there are no reflections of the shield current from the ends and likewise no reflection of the current in the inner conductor from the ends of the cable. The cable used has a cross section that is shown in Fig. 6. The dimensions of the cable are also shown in the same figure. The outer insulation and the inner insulation used is PVC. The inner conductor is made of copper. The shielding is done using galvanized steel. The cable length used in the study is 10 m.



Fig. 5 IRA and the buried cable geometry along with the termination details



Fig. 6 Cross section of the buried cable



Fig. 7 Frequency domain waveform of the electric field along the length of the buried cable

Figure 7 shows the electric fields at different points along the length of the cable in frequency domain. x = 0 m corresponds to one end of the cable, and x = 10 m corresponds to the other end. This figure shows that for the configuration of the system considered, the entire cable gets illuminated by the UWB pulses having frequency component of up to 1 GHz, and for frequencies above this, field is concentrated mostly along the boresight. So the problem of coupling to the cable is a case of frequency-dependent distributed excitation.

4 Response of the Buried Cable to the Electric Fields from IRA

An IRA electric field coupling with the cable can be considered in two stages: coupling with the external circuit and coupling with the internal circuit. The external circuit consists of the soil, the outer layer of the shield, and the outer dielectric layer; the inner circuit consists of the inner layer of the shield, the inner insulation of the cable, and the conductor of the cable. The two circuits are coupled by means of the shield transfer impedance, Z_{t} . The value of the transfer impedance is frequency dependent and is a function of the shield thickness. The value of Z_t is computed using Schelkunoff's equation [8, 9]. The transfer impedance that couples the shield current to the inner conductor is shown plotted in Fig. 8. The value of transfer impedance varies with frequency. The transfer impedance is tens of milli-ohms per meter at low frequency. In this work a tubular shield is considered with no braiding. For such cables the transfer impedance drops with frequency. The characteristic impedance in this case is qualitatively similar to the cable considered in [10]. The incident UWB electric field reaching the cable will induce current in the shield. This can be determined by assuming the shield as a single core conductor with an outer layer of insulation. At all the points along the length of the shield, there is an excitation due to the field. This is a case of the distributed excitation, and hence the current is induced at every point on the cable at the same time. These currents propagate to both the ends and get terminated. The method of analysis should accommodate for these effects. Transmission line analysis was found to be suitable for such a case and hence is adopted in this work. This analysis uses the governing equations for the induced current and voltage as given in the following equations [11]:



Fig. 8 Transfer impedance as a function of frequency

$$I_{\rm e}(x) = \int_0^L G_{\rm i}(x, x_{\rm s}) E(x_{\rm s}, z = d) dx_{\rm s}, \tag{1}$$

$$V_{\rm e}(x) = \int_0^L G_{\rm v}(x, x_{\rm s}) E(x_{\rm s}, z = d) dx_{\rm s}.$$
 (2)

In these equations, $V_e(x)$ and $I_e(x)$ are the induced voltage and current at any point on the shield. E(x, z = d) is the transmitted component of the incident IRA field which illuminates the cable. Since this is the case of a distributed excitation, the current at any point on the shield is the sum of the incident and the reflected (from cable terminations, which is zero in this case) components of the current. This addition is taken into account in Eqs. (1) and (2) where G_i and G_v are the Green's function which automatically takes care of this addition. When the current is induced in the shield, not all of this current gets coupled to the inner circuit. Only a small portion of this shield current gets coupled to the inner circuit, which is given as $Z_t \times I_e(x)$. The induced voltage in the inner conductors can be determined from the following equation [11]:

$$V_{i}(x) = \int_{0}^{L} G_{v}(x, x_{s}) Z_{t} I_{e}(x_{s}) dx_{s}, \qquad (3)$$

where $V_i(x)$ is the induced voltage at any location x in the inner conductor and $I_i(x)$ is the induced current at that location. Z_t is the transfer impedance of the shield. The transfer admittance is very small and hence can be neglected. The expression for the Green's functions G_i and G_v was taken from [11, 12].

4.1 Induced Voltage on the Cable

Induced voltage is representative of the electrostatic coupling to the cable on account of the capacitances in several sections of the cable. The induced voltage is computed using Eqs. (2) and (3) and is shown in Fig. 9 for the cable shield. The voltage in the shield reaches a maximum value of 350 V at the midpoint of the cable where the boresight meets the cable, the nature of this being identical with the electric field at that location. This maximum value of the voltage is attributed to the initial direct radiation at that location. At other points such as at 0, 2, and 10 m, the peak voltage comes to -250, -220, and 250 V, respectively. At points equidistant from the midpoint of the cable, the voltage is equal in magnitude but opposite in polarity which is the result of the superposition of the traveling waves at any location on the cable. The traveling wave components of the induced voltage start at the point where the boresight is located, which is at 5 m away from one end of the cable, and slowly move to the ends with a finite velocity. At the ends of the cable, there are no reflected components of the voltage. So a condition arises, where the distances along the cable from x = 0 m up to the boresight get a net negative



Fig. 9 Induced voltage on the cable shield at different points. (a) x = 0 m, (b) x = 2 m, (c) x = 5 m, and (d) x = 10 m

traveling wave of voltage and hence a negative peak voltage and the remaining locations get a positive magnitude of the peak voltage. This happens for all points on the cable. Furthermore, the plots are oscillatory and the oscillations seem to increase from the boresight to the off boresight points because of the impact of the soil characteristics and the cable parameters. The time taken for the electric field from the IRA to illuminate the cable is 80 ns, and hence the voltage in the shield is established only after this time. The characteristic of the illuminating field is a prepulse which is followed by the impulse. The effects of these fields are seen in the induced voltage on the shields. The duration of the prepulse in air is 2 F/c = 8 ns. However, the prepulse has to travel an extra meter in soil, and hence $\{(2 F/c) + (1 \text{ m/velocity of light in soil})\}$ turns out to be 18.6 ns, with speed of light in soil ~ 9.4×10^7 m/s. Hence, the prepulse nature of the voltage lasts for 18.6 ns as seen in the voltage plots.

The current in the shield of the cable acts as a secondary source of excitation to the inner conductor and hence causes the induced voltage in the inner conductor. On the inner conductor, the induced voltage is obtained as a result of the current induced on the shield and the transfer impedance of the shield. The induced voltage on the conductor is computed at points 0, 2, 5, and 10 m and is plotted in Fig. 10. The peak value of the voltage is respectively 25, 16, -3.6, and -25 mV, and the time of occurrence of the peak is respectively 90, 86, 80, and 90 ns after



Fig. 10 Induced voltage on the inner conductor of the cable at different points. (a) x = 0 m, (b) x = 2 m, (c) x = 5 m, and (d) x = 10 m

the initiation of the voltage on the inner conductor which is ~500 ns. The peak value of the induced voltage in the inner conductor is negative of that in the shield at the respective points because the current flows in the opposite direction in the inner conductor as compared to that in the shield. The time of occurrence of the peak decreases from the right end of the cable to the middle and then again increases to the left, at the same time maintaining a similarity between the identical points on either side of the x = 5 m.

5 Conclusions

The electric field from an IRA antenna has been studied and is found to have prominent boresight radiation. The width of the boresight radiation depends upon whether the observation point is in the near or in the far field. Using this data as the source, the field along the cable is computed and hence the induced voltage is determined. It can be seen that induced voltage gets reduced by a factor of around 100 due to the shield properties. But as this is a communication cable, even this voltage may be large enough to cause a temporary upset of the data in the conductor. Of course, more powerful IRA can cause even higher electromagnetic interference on the communication cables and the associated connected equipments.

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Part VI SP Measurement

HPM Detector with Extended Detection Features

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Abstract The growing threat by high-power microwaves (HPM) also increases the importance of capabilities to detect high-strength electromagnetic fields. Fraunhofer INT has developed a demonstrator of an HPM detector with extended detection features for mobile and stationary use. It allows the measurement of high dynamic amplitudes and of width, repetition frequency, and number of HPM pulses. A four-channel system with four polarization-independent broadband antennas and four logarithmic amplifier/detector units permits additional direction finding to a detected HPM source.

Keywords High-power microwaves (HPM) • Detection

1 Introduction

The potential high-power microwave threat to electronic devices and systems has reached a technological level whereby customary modern electronics can be disabled over small to medium distances. In military domain, the feasibility to put electronic devices and systems out of operation using intense electromagnetic fields is investigated since a couple of years. Especially the effects far beyond the inherent immunity or the immunity enforced by EMC standards against the "common" electromagnetic environment are in focus. These efforts lead to astonishing achievements especially in close-range applications in the last years. Thus, microcontroller or computer-controlled electronics (even in vehicles) can be disabled at least temporarily in a distance of 10 m to several 100 m with suitcase-sized devices.

As these devices are realized with commercially available or easily obtainable components, the suspicion is obvious that persons and groups with criminal or

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terrorist intentions could use this method for burglaries, raids, extortions, or assaults, where electronics is responsible for the security of persons and property.

An advantage for terrorist or criminal actions as opposed to purely military applications is the ability to take up a concealed position very close to the object to be manipulated. In this way it is possible to generate considerable field strengths at the target with moderate power levels and suitable antennas. Without sufficient detection and warning systems to verify this threat, it is very easy for attackers to test the effectiveness of HPM systems on-site without being discovered. Disturbances and failures of own equipment accordingly might not be associated with electromagnetic attacks.

For the mentioned reasons, it is very important to survey critical equipment, systems, and infrastructure against these threats and have available capabilities to find and identify sources of electromagnetic threats. In previous work [1] an HPM detection system has been developed, which exceeds the capability of simple warning devices [2, 3] only detecting hazardous HPM events. It is realized with a logarithmic periodic spiral antenna with an aperture angle of 90° and a logarithmic amplifier/detector, which provides the envelope of the input signal as output signal with a dynamic range of more than 60 dB. A digital oscilloscope and a control PC allow to measure pulse amplitude, pulse duration, pulse repetition frequency, and number of pulses, respectively, and a machine user interface gives simplified information to nonspecialists/untrained users.

In the new four-channel system, four of these detector units are combined each with an antenna in 0° , 90° , 180° , and 270° position. Due to the directional diagram of a cavity-backed spiral antenna, which is approximately 90° , the entire space in the horizontal plane can be monitored with this configuration. Analyzing the amplitude ratios of the individual channels, the direction of the HPM signal can be calculated and displayed as a rough to mid-detailed angle by the analyzing software on the control PC.

Based on the results of the development of the single-channel HPM detection system [1], the following requirements for HPM detection have been considered:

- Announcement of pulse-shaped electromagnetic fields (frequency independent) with hazardous field strength (more than 1 kV/m)
- Frequency range with approximately constant sensitivity from 500 MHz up to nearly 12 GHz, capability of detection below 500 MHz with lower sensitivity
- Damage resistance against field strengths up to several 10 kV/m
- Detection of mid-range HPM sources (e.g., detection of E-fields larger than 100 V/m), frequency independent, for warning and search purposes
- Measurement dynamic range preferable larger than 60 dB
- Independence of polarization
- · Direction independent alert and indication in the horizontal plane
- Classification of detected events according to amplitude, pulse duration, pulse repetition frequency, shape, etc.

Additionally the systems has to be capable to detect the direction of an HPM signal in a rough to mid-detailed angle resolution.

2 Four-Channel HPM Detector

2.1 Antenna System and Direction Detection

Four circularly polarized logarithmic periodic cavity-backed spiral antennas are used as antenna system. Advantages of these antennas are the large bandwidth and the independence of polarity, as neither the frequency nor the polarization of the HPM source to be detected is known. The four spiral antennas are mounted in a fixture with 0° , 90° , 180° , and 270° angles in a horizontal plane (Fig. 1).

The antenna type is designed for the frequency range of 0.5 GHz up to 2 GHz with a gain of 0 dBiL (for linear polarization) and can be used in the frequency range from 0.5 GHz up to 10 GHz without significant gain variation (Fig. 2). Even below 500 MHz down to 250 MHz, HPM signal detection without precise field value determination is possible.

The radiation pattern of the antenna is quite flat and free of side lobes in the frequency range from 0.5 up to 3.5 GHz (Fig. 3, LHS). At 4 GHz the diagram starts to get more corrugated (Fig. 3, RHS), and the reverse loss is getting worse, too (at least in this test setup). Even here, there are no side lobes in the main lobe, which could affect the direction analysis significantly.

Figure 4 shows as an example the radiation pattern of the four by 90° staggered antennas at a frequency of 2.5 GHz. It is more complex than that of the single antenna due to antenna interaction. One can recognize, however, that the direction analysis of an HPM source by vector addition of the antenna signals is admissible.

Figure 5 shows the vector sum, transformed into polar coordinates, of the antenna radiation patterns again for the example frequency of 2.5 GHz. The calculated actual angle deviates on average by 8° from the target angle in this case. The standard deviation is below 20° in the whole investigated frequency range.



Fig. 1 Antenna system with logarithmic periodic cavity-backed spiral antennas



Fig. 2 Antenna gain (for linear polarization) of the spiral antenna



Fig. 3 Radiation pattern of the spiral antenna. *LHS*: 0.7 GHz (*blue*), 1 GHz (*black*), 2 GHz (*yellow*), 3 GHz (*red*), *RHS*: 4 GHz (*black*), 5 GHz (*red*), 6 GHz (*yellow*), 7 GHz (*blue*)

2.2 System Demonstrator

The four-channel system consists of four individual channels similar to the singlechannel system [1]. The block diagram is shown in Fig. 6.

An attenuator of 60 dB follows behind the antenna to reduce the expected amplitude range of the input signal to the operation range of the logarithmic


Fig. 4 Radiation pattern of the four-antenna system at a frequency of 2.5 GHz



Fig. 5 Calculated actual angle in dependence on target angle at a frequency of 2.5 GHz

amplifier/detector unit. The pulse robustness of the attenuator is a critical issue here, because pulsed input signals up to +60 dBm/1 kW are expected, corresponding to 223 V at 50 Ω . The attenuator unit has been divided into three sections with 20 dB, where the first 20 dB device is a power attenuator with high pulse rating and relatively small size as four devices had to fit into the chosen shielded housing.

The dynamic range of the amplifier/detector exceeds 0 dBm slightly, whereas the damage limit is a little bit more than 10 dBm. Therefore, the IC has to be



Fig. 6 Block diagram of the four-channel system



Fig. 7 Demonstrator of the entire four-channel system with USB oscilloscope

protected safely against higher signal levels. A 9 dBm limitation is realized with a limiter diode, and the peak power rating is above the expected signal levels. The PCB with the logarithmic amplifier/detector has been placed into a milled solid aluminum box to ensure sufficient screening against external fields.

The entire demonstrator consists of four broadband spiral antennas, the RF unit, a four-channel oscilloscope, a computer (typically a laptop) with GPIB or USB interface, and a mains or battery power supply (Fig. 7). The RF unit in the lower left

of Fig. 7 consists of another screened housing with high shielding effectiveness, containing the logarithmic amplifiers/detector modules, the input attenuators, the limiter diodes, and the power supply filtering.

The system can be operated with internal batteries alternatively to the mains or vehicle power supply. The oscilloscope either is equipped with an internal battery, which can supply the RF unit, too, or is driven with an external mains power supply or an external battery instead. The computer can also run with internal battery. Thus, the detector system can be placed very flexibly at any desired location without necessity of external power supplies. The antenna and the RF unit can be deployed outside of protected areas. The oscilloscope and the computer should be protected against the expected field values placing them in a screened housing or in the protected area. During source search mode at smaller field values, the screening efficiency of a vehicle might be sufficient.

The pulse signals of the RF unit are processed and displayed by the four-channel oscilloscope. With the built-in functions, the signals can be analyzed for numerical signal parameters like pulse amplitude, pulse width, rise times and fall times, and pulse repetition time. These parameters can be retrieved by a control computer via GPIB interface combined with the current time stamp of the occurrence. Alternatively, a USB interface is available for control and data transfer. In case it is placed in a protected area beside the control computer, it is connected directly via USB to this computer. With approximate knowledge of the transmitter frequency, the field strength at the receiving location can be derived from the pulse amplitude. Furthermore, the complete pulse oscillograms can be stored in the computer for postprocessing. The direction of the HPM signal can be calculated from the vector sum of the antenna output amplitudes.

The basic control software currently used has been developed at Fraunhofer INT. In a future development step, it might be substituted by a customer-specific script in LabView (National Instruments) environment to offer a simple graphical user interface for nonspecialists/untrained users. Furthermore, the analog-to-digital conversion and the pulse analysis can be implemented in the well-screened RF unit instead of using an external oscilloscope. For that a small-sized modular USB oscilloscope without display and control panel is planned to be used. The full control and data transfer of this oscilloscope are realized then via the USB interface and a USB-to-fiber optics converter to an also well-screened control unit with a USB-to-fiber optics converter included. With this the entire system can be used in any RF environment without restrictions.

Sensitivity and robustness tests of the detector systems have been done in the TEM waveguide and the reverberation chamber of Fraunhofer INT. We found good agreement of measured field strength with that calculated from the detector characteristic for a frequency of 1 GHz. As an example Fig. 8 shows the aggregation of the results for frequencies between 1.2 and 2 GHz. The measurements have been done with frequency steps of 50 MHz between 550 MHz and 3.1 GHz for HPM pulses of 1 μ s width and pulse repetition frequency (PRF) of 1 kHz. HPM robustness for the RF part has been verified in TEM waveguide for field strengths up to 1.5 kV/m in the above-mentioned frequency range. The detector has passed



Fig. 8 Measured (*circles*) and calculated field strengths (*black*) versus detector voltage for frequencies between 1.2 and 2 GHz together with the \pm 3 dB lines (*green*)

robustness tests in the reverberation chamber of Fraunhofer INT at average electrical field strength up to 10 kV/m in the frequency range from 500 MHz to 3.4 GHz.

3 Conclusion

The aim of the work presented here was to design HPM detection systems and to build demonstrators that offer the possibility, beyond the simple warning devices presently available, to record and to display a number of pulse parameters. These include pulse amplitude, derived threat field strength, pulse width, and pulse repetition rate or the number of pulses for low repetition rate. It is thus possible to detect and distinguish a variety of signal types ranging from continuous wave to narrow-band HPM pulses. In a second stage, a four-channel system was designed with four antennas for direction finding. In a later phase, the intention is to determine also other characteristic features such as a coarse discrimination of frequency regions.

The suitability of different detection methods for HPM signals was evaluated from own investigations as well as from the literature. Beyond the mere announcement of a signal with threat field strength, the detection system is intended also for surveillance of a certain area or for search and identification of HPM sources. The necessary dynamics can be achieved by logarithmic amplifier/detector modules with a measuring dynamics of 60 dB and bandwidths of up to 8 GHz.

The single-channel system consists of a polarization-independent broadband antenna, which covers a defined sector at constant sensitivity, and the logarithmic amplifier/detector module in an enclosure with high shielding effectiveness and with an appropriate input circuit for self-protection. This is supplemented with a signal processing system with a multichannel oscilloscope and a computer and with the necessary analysis and display software [1]. The four-channel system comprises four antennas, attenuators, limiters, and log-amplifier/detectors and a four-input trigger recognition circuit. The pulse analysis and direction finding are again done with a four-channel digital oscilloscope. The sensitivity and HPM robustness of the detectors have been tested in a TEM waveguide with field strengths up to 1.5 kV/m and in a reverberation chamber for fields up to 10 kV/m.

Both systems can be operated by mains or onboard power supply and with internal batteries, respectively, and thus can be flexibly set up in any location in both stationary and mobile settings without relying on external power supply.

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High Dynamic Range, Wide Bandwidth Electromagnetic Field Threat Detector

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Abstract Because intentional electromagnetic attacks may be difficult to distinguish from random EMI, real-time detection of threatening fields is advisable for critical facilities. We describe a low-cost approach for configuring a high-power electromagnetic field detector and central controller that can be used for detection of intentional electromagnetic interference (IEMI) and other high peak field events such as electromagnetic pulse (EMP). The information stream can be used to provide threat warning, threat tracking and prediction, threat location, security alerts, technician trouble-shooting alert, and data forensics, as well as providing a deterrent to IEMI attacks.

Keywords EMP detector • RF weapon detector • High-power electromagnetic field detector • Electromagnetic attack warning system

1 Introduction

The fundamental requirements for the RF detector class of interest derive from the need to detect and quantify an electromagnetic illumination on critical facilities from an intentional or unintentional high-power RF source [3–8]. Wide dynamic range is also a desirable characteristic, so that probing or more distant threats can be

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detected well below levels of upset or damage. The basic threat levels of interest are based on the electric field values that can (a) corrupt or alter electronic circuits and data states and (b) damage or weaken electrical and electronic components in servers, computers, SCADA controls, etc. The baseline values for these energy levels are taken from Baker et al. [1] for single pulse events:

- 1.0 µJ/cm² for circuit upset/data corruption onset
- $200 \,\mu\text{J/cm}^2$ for the onset of component damage

This translates into electric fields (E-fields) of hundreds to many tens of thousands of volts per meter. The challenge then is to select a detection antenna configuration and electronics that have the following characteristics:

- · High E-field measurement capability
- Wide bandwidth (100 MHz–10 GHz for the class of electromagnetic threats of interest)
- Wide dynamic range
- Direct and accurate E-field measurement
- · Low complexity
- · Avoid expensive integrators
- Low-cost commercial parts

The combination of these characteristics rules out most of the commercial EMI field probes, and also the high cost, high accuracy MIL-STD B-dot and D-dot devices. The electro-optic (Pockel cell) devices are also unsuitable at this time due to their development status and high cost. Recent testing has led to a proprietary "inferential" approach to high electromagnetic field detection. Inferential refers to the fact that we are using the far-field measurement of the *magnetic* field of the EMP/IEMI wave, which is directly proportional to the high electric field of interest in air. The high electric field is "shielded" from the measurement, resulting in antenna output voltages on the order of a few volts instead of hundreds or thousands of volts with associated arcing, connector problems, noise problems, and attenuation requirements. The major benefit is that the "inferential" antenna output can be interfaced directly into high-frequency analog RF-integrated circuit electronics. This greatly reduces the cost of the entire detector system. Figure 1 shows an early production EM InferentialDetectorTM (Patents Pending).

2 Block Diagram

The system contains three measurement channels. Each channel is connected to one split-shield loop antenna. By using three antennas, one on each orthogonal axes, the magnitude of the incoming electromagnetic field can be measured regardless of its angle of arrival or polarization. Figure 2 shows the system block diagram, including one of the three orthogonal channels.





Fig. 2 Simplified block diagram

The performance parameters for the system are:

- Frequency range: 100 MHz–10 GHz
- Field strength: 100 V/m-100 kV/m
- Minimum pulse width: 10 ns
- Rise time: <10 ns
- Measurement interval: every 1 ms

Two main challenges were posed in achieving these parameters. The first was to maintain a flat frequency response over this broad range, so that accuracy requirements could be achieved. The second was the need to make these measurements on very fast and narrow RF pulses, requiring good high-speed front-end electronics design discipline.

3 Antenna Design

Space limitations made it impractical to build an antenna with a flat frequency response down to 100 MHz. Instead of building a flat antenna, we chose to build a simple antenna and equalize its frequency response using high-speed analog RF electronics.

The antenna selected is a split-shielded loop [2]. This antenna is sensitive to magnetic rather than electric fields. Since the electric and magnetic fields in far-field locations are related to each other by the scalar impedance of free space, 377 Ω , it is a simple matter to calculate the electric field from the measured magnetic field. The split shield around the loop shields the loop from electric fields, so that only the magnetic field is measured.

We use a small-shielded loop antenna because our experience has shown that B-field measurements are usually more accurate and repeatable than E-field measurements. The loop, since it is balanced, is less affected by the presence of other conductive material in its environment because the common-mode impedance to ground does not affect its sensitivity.

What is perhaps somewhat surprising is that a balun is not required to convert the balanced loop to a single-ended signal. Extensive experience with these types of antennas, such as the Beehive Electronics model 100A, has shown that very low surface currents are present on the outside of the shield; the common-mode impedance of the outside of the shield is high enough to reduce these currents to a very low level.

These types of antennas are typically built from a length of coaxial cable. In our case, they are built more precisely using RF-printed circuit technology. The "coaxial" section is made using a stripline transmission line. Ground continuity between the two ground planes is created by using closely spaced ($d \ll \lambda$) plated-through holes (Fig. 3).

The entire antenna array is shown in Fig. 4. The loop antennas are embedded in the tips of the printed circuit board. The dielectric antenna mount is designed to align the three antennas, so that the axis of each antenna is orthogonal to the other two.

4 Equalizer

A small loop offers a very predictable frequency response at low frequencies; it is effectively a high-pass filter with a 20 dB/decade slope. This response is well-behaved until the loop reaches its self-resonant frequency. We compensate for this frequency response by cascading it with a low-pass filter. The filter's -20 dB/decade slope, in theory, perfectly cancels the antenna's opposite slope and results in a flat frequency response. Given our desire for a flat frequency response up to 10 GHz, careful tuning of the filter is required, since we are pushing the limits of surface mount technology.



Fig. 3 Transmission line cross section





5 Pulse Response

The measured RF pulses can be as narrow as 10 ns. At the output of the logarithmic demodulator, these are converted into longer DC pulses with the same peak characteristics. The DC pulses are sampled at a 1 kHz rate by the microprocessor and displayed on a PC using a 1-s peak sample basis.

Another approach would have been to use a high speed analog-to-digital converter (ADC) running at several hundred megasamples per second. In that case, digital hardware could search the data stream for the peaks of the incoming values. This approach is viable but is expensive, power inefficient, and complex. For these reasons, we chose to sample at a much slower rate. This allowed us to use the on-board ADCs on the microprocessor itself to digitize the data. At this slower rate, all the data manipulation can be performed in firmware instead of requiring a field programmable gate array.

6 Peak Detector

Our approach requires analog peak detectors to capture the peak value of the demodulated DC pulses and hold them long enough for the ADCs to capture them. One of these peak detectors is shown in Fig. 5. When designing peak detectors, there is always a trade-off between the fast rise time required to capture fast pulses and the long hold time required for the ADC to measure the output. Fast rise time requires a small sampling capacitor to hold the peak voltage; larger capacitors cannot be charged quickly enough to capture a short pulse. However, a small capacitor can only hold a small amount of charge. As a result the capture voltage rapidly droops due to leakage current in the capacitor, Schottky diode, or buffer. Our solution was to use a two-stage peak detector. The first peak detector is high speed—it uses a 68 pF capacitor. This is small enough that the peak detector has a rise time of approximately 5 ns. This peak detector droops quickly as a result of the small capacitance value. Following this peak detector is a second, slower, peak detector that has a long hold time. This peak detector uses a 0.033 µF hold capacitor. The second peak detector would never be fast enough to catch our 10 ns pulses, but the first peak detector holds the pulses long enough for the second peak detector to capture them. The unique combination of two sampling stages gives the fast rise time required, yet holds the peak long enough for the ADC to measure its output.

7 Reset

It is a common practice to use a reset switch in a peak detector, viz., after a measurement is complete, the switch is closed and the hold capacitor is discharged. While the switch is closed, the peak detector cannot capture pulses and is effectively blind. If a large, short-duration RF burst arrives while the peak detector switch is closed, the peak detector would miss the burst, i.e., the system would not be aware that the burst occurred. Our solution is to use a 10 M Ω bleeder resistor as shown in Fig. 5, instead of a typical switching circuit. The resistor discharges the



Fig. 5 Two-stage peak detector

capacitor, but very slowly. Over the course of the 1 ms measurement interval, the droop is insignificant.

8 Detector Packaging

The above discussion describes the antennas and electronics for the Inferential-DetectorTM. These have been packaged as illustrated in Fig. 6.

The housing is machined out of a solid block of T-6061 aluminum. The walls of the RF-tight cavity (on the left in Fig. 5) are maintained at $\frac{1}{4}$ inch or greater, including around blind-tapped bolt holes to attain high (>90 dB) shielding for the electronics and copper-to-fiber conversion. The fiber optic output is fed through a waveguide beyond cutoff to a junction box which also contains the EMP power filter for the incoming 24 V dc power to eliminate conducted susceptibility to high external fields. This detail is shown in Fig. 7.

All housing pieces are black anodized before final assembly, with appropriate masking to provide good conductivity to the gasketing material. The RF-tight cavity is sealed with RF gasketing material, and both the RF cavity and the J-box are environmentally sealed, including the dielectric antenna dome.

9 System Implementation

The above discussion of the detector design describes a single device that measures electric fields as high as 100 kV/m, including very fast rise time transients associated with EMP and IEMI weapons, and provides continuous output information in the form of the peak field magnitude seen during each one second sample period. The serial data output is provided over a single fiber capable of providing simultaneous input and output using different wavelengths (1,310/1,550 nm). This fiber data transmission makes the data flowing into and out of the RF-protected volume impervious to high electromagnetic fields in the sensor vicinity. The single string data flow is shown in Fig. 8.

The system architecture is designed to monitor multiple sites for abnormal electromagnetic field activity, with the capability to monitor multiple detectors at each site. During each 1 s sample period, the peak detector output is compared with a preselected threshold. If the threshold is exceeded, a warning alarm appears on the output dashboard, shown in Fig. 9.

All site detectors can be individually monitored and examined for status or detail in the event of an abnormality to examine detail in its real-time behavior, as shown in Fig. 10.

In an integrated security system with electromagnetic sensing and warning capability, the EM detection feature can be combined with additional traditional security features, such as visible and infrared cameras. Based on vector information



Fig. 6 InferentialDetector[™] Packaging



Fig. 7 InferentialDetector[™] J-Box Detail

from the InferentialDetectorsTM, cameras may be pointed in the direction of a sensed electromagnetic threat. The cameras can provide useful information about the threat even if they are destroyed by the high EM field, since their stored data can be examined in a secure and EM-protected facility right up to the initiation time of the RF attack. Such a security system might look conceptually like Fig. 11.

The present InferentialDetectorTM architecture contains only a scalar computation of the maximum EM field level at each sample interval, since the three orthogonal axes of information are root-sum-square combined to obtain the E-field amplitude. However, a more advanced version is being developed where the three individual axes are recorded separately to enable determination of the Poynting vector S, whose azimuth indicates the direction to the EM source, regardless of its antenna characteristics, using data from multiple detectors. This architecture is shown in Figs. 12 and 13.







Fig. 9 Detector monitoring and display dashboard



Fig. 10 Individual detector dashboard display

Each detector can then determine a unit normal n which is parallel to the magnetic field vector B measured at the detector. This unit normal defines a plane containing the E-field vector and the Poynting vector at that detector. The intersection of the planes from two detectors determines the line containing the Poynting vectors from the detectors to the EM source. The Poynting vector plane measured by a third detector enables triangulation to determine the exact location of the source.



Fig. 11 Total integrated EM and video security system



If a contour map of the terrain surrounding a protected site exists, and if it is assumed that the electromagnetic threat is ground-based, the Poynting vector plane's intersection can be computed as shown in Fig. 13 from the $n_1 \times n_2$ data from two detectors and that line's intersection with the ground topography defines the source location. If no such ground knowledge exists, the source location can still be computed from the data from three or more detectors, on a best fit basis.

Fig. 13 Non-isotropic source determination of planes containing E-field vector and Poynting vector and their intersection



10 Summary

A cost-effective electromagnetic threat detection approach has been developed with desirable features:

- Very high EM field capability—100,000+ V/m
- Wide dynamic range up to 10 GHz
- Useful for IEMI and EMP detection
- Information for threat warning, threat tracking and prediction, and post-attack forensics
- Data for derivation of threat location
- Data for security alerts
- · Low bandwidth output

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Resistive Sensor for High-Power Microwave Pulse Measurement in Double-Ridged Waveguide

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Abstract A resistive sensor, the performance of which is based on electron heating effect in semiconductors, for the measurement of high-power microwave pulse in a double-ridged waveguide has been developed. Finite-difference time-domain method was used for the calculation of electric field in the waveguide section with a sample made from *n*-type Si. Cross-sectional dimensions and specific resistance of the sensing element have been chosen to optimise the performance of the sensor. It was implemented in the waveguide WRD250 covering frequency range 2.60–7.80 GHz. Output signal dependence on pulse power and frequency response of the sensor have been measured.

Keywords Resistive sensor • Microwave pulse measurement • Finite-difference time-domain method • Double-ridged waveguide

1 Introduction

A resistive sensor (RS), the performance of which is based on electron heating effect in semiconductors, consists of a sensing element (SE) made from n-type Si [1]. The SE is actually a bulk resistor with two Ohmic contacts placed in a waveguide. When the electric field of microwave pulse heats electrons in the SE, its resistance increases, and by measuring this resistance change, pulse power in the transmission line is determined. The RS has applications for high-power microwave (HPM) pulse measurement. Some advantages can be mentioned when comparing the RS with a semiconductor diode, which is also sometimes used for HPM pulse measurement. The RS that measures HPM pulses directly is overload resistant and demonstrates perfect long-term stability [1].

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A few types of the RS have been developed to cover intermediate and very high microwave pulse power range. For an intermediate power range, a crosswaveguide-type RS with the SE covering the waveguide's window was used. Cross-sectional dimensions and specific resistance of the SE was chosen to get flat frequency response within pass band of the waveguide [2]. For very high microwave pulse power range, the RS with a sensing element placed under a thin metal diaphragm stretched in parallel to a wide wall of the waveguide [3, 4], as well as the layout with two sensing elements, the height of which is much less than the narrow wall of the waveguide [5, 6], was considered, and a flat frequency response has been engineered by choosing the optimal electrophysical parameters of the SE. Pulse power level measured by these sensors is practically limited by the breakdown in the waveguide. Unfortunately, the RS for a particular waveguide covers frequency range that is limited by a pass band of it. From this point of view, a double-ridged waveguide shown in Fig. 1 is very attractive, since its pass band is much wider than the rectangular one. As one can see from the figure, the doubleridged waveguide is a traditional rectangular waveguide containing two metal inserts or ridges. The region of the waveguide between metal ridges is usually named as a gap region, whereas side regions are named as troughs.

2 Resistive Sensor in Double-Ridged Waveguide

Cross-sectional view of the double-ridged waveguide is shown in Fig. 1. In the present paper, we used WRD250 waveguide [7], the parameters of which are collected in Table 1. In the table characteristic dimensions of the waveguide, lowest and highest passing frequencies, cutoff frequency and maximum pulse power are presented. It is seen that the ratio $f_{\text{max}}/f_{\text{min}} = 3$ is a characteristic of the double-ridged waveguide, whereas for the rectangular waveguide, this ratio is roughly 1.5. Thus, mounting the SE in a double-ridged waveguide, the frequency range, where the RS operates, can be significantly widened. The smaller distance between metals in the gap region is a reason why the smaller pulse power can be transmitted through the waveguide, but according to the last column of the table it is still significant.



 Table 1
 Parameters of waveguide WRD250

a, mm	b, mm	s, mm	d, mm	f_{\min} - f_{\max} , GHz	$f_{\rm c}, {\rm GHz}$	$P_{\rm max}$, kW
42.0	18.2	11.2	3.8	2.6-7.8	1.932	120

Closed form approximation for the critical wavelength λ_c of the dominant TE₁₀ mode and formulas describing power flow *P* in a double-ridged waveguide can be found in [8]. Description of electromagnetic field components in a double-ridge waveguide [8] is based on retaining a single mode in a gap region with the components E_y , H_x and H_z , the amplitudes of which depend on coordinate *x* in a similar way as mode TE₁₀ components in the ordinary rectangular waveguide. In a trough region, the amplitudes E_x , E_y , H_x , H_y and H_z are expanded in infinite series of modes. It is obvious that for mode TE₁₀ $E_z = 0$.

The SE of the RS is schematically shown in Fig. 1. It is seen that a bar-shaped n-Si sample with contacts on its ends is inserted between the ridges. Its height *h* corresponds to the span between ridges *d*, the width is denoted as *w* and the length in a wave propagation direction is *l*. General requirements for the n-Si sample that can serve as the SE can be formulated as follows: the SE should not cause considerable reflections in the waveguide; hence, the value of a voltage standing wave ratio (VSWR) has been set less than 1.5; the DC resistance of the RS should not exceed 1 k Ω enabling measurement of microsecond duration microwave pulses; and the frequency response of the RS in the waveguide's frequency band should be as flat as possible.

Since the performance of the RS is based on a resistance change in the electric field of the microwave pulse, it is common to define a sensitivity of it as a ratio of relative resistance change $\Delta R/R$ to the power transmitted through the waveguide

$$\zeta = \frac{\Delta R/R}{P} = \frac{\Delta R/R}{P_{\infty}\sqrt{1 - (\lambda/\lambda_c)^2}},\tag{1}$$

where P_{∞} is power flow at infinite frequency, λ is a wavelength of microwaves and square root appearing in the denominator of the expression accounts for dispersion of the wave in a waveguide. In waveguide's WRD250 frequency range, electron heating inertia does not manifest itself; therefore, relative resistance change, in a so-called warm electron region [1], can be expressed as

$$\frac{\Delta R}{R} = \frac{3}{2}\beta \langle E \rangle^2,\tag{2}$$

where β is a warm electron coefficient that defines deviation of the current–voltage characteristic from Ohm's law and $\langle E \rangle$ is an average of the electric field amplitude in the SE. Inserting Eq. (2) into Eq. (1), one can get the final expression describing the sensitivity of the RS in the waveguide:

Ž. Kancleris et al.

$$\zeta = \frac{3\beta E_0^2}{2P_\infty} \times \frac{1}{\sqrt{1 - (\lambda/\lambda_c)^2}} \times \frac{\langle E \rangle^2}{E_0^2},\tag{3}$$

where E_0 is an amplitude of electric field in the centre of empty waveguide. In Eq. (3) three multipliers are discriminated. The first one is independent of frequency. The second multiplier in the obtained expression accounts for the fact that due to wave dispersion in the waveguide even at the same power level transmitted through the waveguide, the electric field strength in it decreases with frequency leading to the decrease of the sensitivity. The third multiplier is a square of average of the electric field in the SE normalised to the amplitude of the electric field in the centre of the empty ridged waveguide. It is easy to calculate (refer to Table 1) that in a pass band of the waveguide WRD250, the second multiplier decreases by a factor 1.50. Therefore, the increase of the average electric field in the SE in the same frequency range by a factor 1.22 will compensate the decrease of the sensitivity. Therefore, the main task of our theoretical consideration was to choose electrophysical parameters of the SE, enabling the increase of the average electric field amplitude in it with frequency compensating the decrease of the sensitivity due to dispersion in the waveguide.

We used a finite-difference time-domain (FDTD) method for the calculation of the electromagnetic field components in the structure. Using this method, all volume of interest is filled with points where components of electric and magnetic fields are calculated. The section of a double-ridged waveguide with the SE inside has been modelled. Non-reflecting boundary conditions at both ends of the section are satisfied. At a few grid points from the left side of the structure, the regular TE_{10} wave is excited. Amplitudes of electric and magnetic fields were calculated during each period. The values of E_{ν} calculated in the current period have been compared with the ones calculated one period earlier in order to determine whether the computation can be terminated. Calculations were terminated when the difference between amplitudes of E_v was less than 3 %. In our calculations we accept that the relative dielectric constant of Si is $\varepsilon = 11.9$. The values of warm electron coefficient β for different specific resistance n-Si samples at room temperature are known [9]. Therefore, by calculating average electric field in the SE and making use of expression (3), the sensitivity of the RS is determined. As follows from Table 1, dimensions of the waveguide WRD250 expressed in millimetres have tenths of mm; therefore, for exact modelling of the waveguide window, sufficiently small step should be used. Details of the modelling procedure and calculation results for cruder model a = 42 mm, b = 18 mm, s = 11 mm, d = 4 mm can be found elsewhere [10].

For practical realisation, we chose two types of the SE, the parameters of which are collected in Table 2. In the table dimensions of the SE, their specific resistance, resistance of the SE, calculated value of a voltage standing wave ratio (VSWR) and sensitivity variation within waveguide's WRD250 pass band are presented. Sensors

Table 2 Parameters of SE	<i>h</i> , mm	w, mm	<i>l</i> , mm	$\rho, \Omega \mathrm{cm}$	<i>R</i> , Ω	VSWR	$\Delta \zeta / \zeta$
tested experimentary	3.8	1.0	2.0	50.0	950	<1.3	±12.5 %
	3.8	1.0	1.0	10.0	380	<1.35	±12.3 %

have been manufactured and mounted in the double-ridged waveguide. Three sensors from each group were tested.

3 Measurement Results and Discussion

For the measurement of the output signal dependence on a pulse power, we employed magnetron generators in S (2.75 GHz) and C (5.7 GHz) frequency bands. The measurement setup is shown in Fig. 2a. It is seen that the RS is connected directly to the magnetron generator via transition from the rectangular to the double-ridged waveguide. Pulse duration was 4 µs and repetition rate— 25 Hz. The reference RS was connected to the main port via directional coupler. It controls pulse power in the main waveguide. By changing damping of the precise attenuator A, the pulse power in the main waveguide is set, enabling measurement of the output signal dependence on pulse power. The sizes of rectangular waveguides used in experiments were $72 \times 34 \text{ mm}^2$ for S band (WR284) and $48 \times 22 \text{ mm}^2$ (WR187) for C band. A DC current source was used to feed the RS. Voltage drop 10 V on the SE was kept. The output signal from the RS can be measured with or without additional amplifier. The amplification of it was K = 10. Since the input resistance of the amplifier or oscilloscope is much larger than the resistance of the SE, the dependence of the relative resistance change on the pulse power can be easily determined from the measurement results. It is well established [1] that $\Delta R/R$ dependence on the pulse power in a wide range of P can be approximated as the second-order polynomial in the following way:

$$A\frac{\Delta R}{R} + B\left(\frac{\Delta R}{R}\right)^2 = P,\tag{4}$$

where coefficients *A* and *B* are determined by fitting experimentally measured dependence of $\Delta R/R$ on *P* with Eq. (4). Comparing Eq. (4) with Eq. (1), one can see that $\zeta = A^{-1}$.

For the measurement of the dependence of the sensitivity of the RS on frequency, we used wide band low-power microwave generator producing roughly 200 mW. To increase the measurement sensitivity, so-called meander modulation of the microwave signal (microwave pulse repetition period is twice longer than the duration of the pulse) and lock-in amplifier technique has been used [9]. The measurement setup is shown in Fig. 2b. It is seen that microwave generator is connected to the main waveguide via coaxial to a rectangular waveguide adaptor. In turn, the RS under the test is connected to the rectangular waveguide using



Fig. 2 Output signal measurement setup using high-power magnetron generator (**a**) and wide band low-power generator (**b**) with a DC current supply for the RS feeding. Precise attenuator is denoted as A, T is a transition from WR to WRD250 waveguide or transition from coaxial line to WR waveguide, DC denotes directional coupler and RS indicates the sensor under test. (**a**) Amp is DC pulse amplifier, calibrated RS in the rectangular waveguide serves as reference power meter and output signal is measured with oscilloscope. (**b**) An average power meter from Rohde & Schwarz is used as a reference power sensor, output signal is measured with lock-in amplifier and measurement results are collected by PC

transition from the rectangular to the double-ridged waveguide. Pulse power in the main waveguide was controlled by the average power sensor from Rohde & Schwarz. To avoid lattice heating, the modulation frequency has been chosen high enough $f_m = 10$ kHz [9]. At such condition, the sensitivity of the RS can be expressed in the following way [9]:

$$\zeta = \frac{\pi U_{\rm m}}{\sqrt{2}U_0 P},\tag{5}$$

where $U_{\rm m}$ is the output signal measured with the lock-in amplifier and U_0 is the DC voltage drop on the SE.



Fig. 3 Measured dependences of the relative resistance change of the RS on pulse power in the waveguide at 2.75 GHz (a) and 5.7 GHz (b). Points show measurement results for two different specific resistance sensors with (K = 10) and without (K = 1) amplifier, respectively. Solid line demonstrates two terms fitting Eq. (4). Parameters of the SE: $\rho = 10 \Omega$ cm, $h \times w \times l = 3.8 \times 1 \times 1$ mm³; $\rho = 50 \Omega$ cm, $h \times w \times l = 3.8 \times 1 \times 2$ mm³

Measured dependences of $\Delta R/R$ on pulse power at 2.75 and 5.7 GHz are shown by points in Fig. 3. It is seen that using DC pulse amplifier the relative resistance change of the order of 10^{-4} can be measured. In a high power region, measurements were done up to approximately twofold increase of the resistance of the RS. Measured power ranges from Watts up to tens of kW. It is seen that the RS made from higher specific resistance material is more sensitive in comparison with the lower specific resistance sensor. The same feature has been predicted by our calculations. Solid lines in the figure show two terms fitting using Eq. (4). It is seen that used approximation fits well on experimentally measured results. From the fitting procedure, the coefficients A and B were determined Eq. (4). The sensitivity values determined from the coefficient A are also shown in Fig. 4 by open triangles.

Measured sensitivity dependences on frequency using meander modulated low-power microwave signal are shown in Fig. 4. Three samples of each SE, the parameters of which are collected in Table 2, have been measured and the averaged results with error bars are shown. As it is seen from the figure, sufficiently small error bars are characteristic of the measured values of the sensitivity; therefore, the SEs with repeatable characteristics have been manufactured. In the figure calculated values of the sensitivity are also shown by a solid line. Although exact values of waveguide window up to tenth of mm was used in the calculation, only reasonable agreement between calculated and experimentally measured values of the sensitivity was obtained. The largest discrepancy appears at the lowest and highest frequencies. It is seen that sensitivity measured using HPM pulses reasonably agrees with the low-power measurement results. Measured sensitivity variation is roughly two times larger as predicted by our calculation (ref to Table 2). It might be caused by the fact mentioned in Sect. 2 that different number of modes is taken into account when calculating electromagnetic field components in the gap and trough regions of the waveguide.



Fig. 4 The dependences of the sensitivity if the RS on frequency: (a) $\rho = 10 \ \Omega \ cm$, $h \times w \times l = 3.8 \times 1 \times 1 \ mm^3$, (b) $\rho = 50 \ \Omega \ cm$, $h \times w \times l = 3.8 \times 1 \times 2 \ mm^3$. *Points* demonstrate measurement results—the average for three samples; *solid line* corresponds to the calculations. *Triangles* show the values of the sensitivity calculated from the experimental results shown in Fig. 3

4 Conclusions

The RS in a double-ridged waveguide was realised and tested experimentally. Cross-waveguide-type sensing element fixed between waveguide ridges was used. Output signal dependencies on pulse power have been measured using high-power microwave pulses. It was found that the RS measures pulse power in the range from Watts up to a few tens of kilowatts. Sensors frequency response was investigated using low-power meander modulated microwave pulses, and a reasonable agreement between measured and calculated results was obtained. The frequency range of the RS has been increased significantly in comparison with the sensor in the standard rectangular waveguide. Experimentally measured variation of the sensitivity ± 25 % is larger than expected from our calculations but this value is good enough having in mind wide frequency range of the RS. It seems that the sensor can find its main application working together with a double-ridge horn antenna for the measurement of the HPM electric field strength in free space.

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Characteristic HPEM Signals for the Detection of IEMI Threats

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Abstract HPEM sources can be threats to electronic equipment and might be used with criminal intent. Therefore, these sources have to be detected. To realize a detection system that is independent of known sources and their parameters, these parameters have to be described in an analytical way. In this work, a sophisticated detection system is briefly presented. For the algorithm of the detection system, this work shows that analytical functions are adequate to describe real HPEM signals.

Keywords Intentional electromagnetic interference (IEMI) • High-power electromagnetics (HPEM) • Detection • Sensor network

1 Introduction

In the last years, the importance of electronic equipment has increased more and more. Due to this change, the vulnerability of the society has increased as well, because of the equipment's possibility to fail. This can occur due to electromagnetic interferences [1, 2], which can be produced by natural sources, such as lightning strikes, or by man-made sources. The man-made sources are divided into two subcategories: *intentional* and *unintentional* interferences, where the intentional disturbances are referred to as intentional electromagnetic interferences (IEMI) [3, 4]. Such interferences can be produced by high-power electromagnetic (HPEM) sources [5–7].

In the scope of this work, a system for the detection of intentional interferences has to be realized. Previous works on the detection of IEMI are based on a detector

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diode [8] or a demodulating logarithmic amplifier [9], loosing important information such as the fundamental frequency of interference signals. That is why this detection system is based on the fast sampling in time domain and a detection with software algorithms.

The overall topology of this sophisticated detection system will be presented in the first part of this work. One part of this detection procedure is the application of the detection algorithms. These algorithms have to be independent of the parameters of known sources such as rise times, fall times, and fundamental frequencies. Therefore, in the ongoing part of this work, the characteristics of HPEM signals are collected and described in an analytical way. Afterwards, it will be shown that these analytical expressions are adequate to realize a parametric detection algorithm.

2 Architecture of the Detection System

The architecture of the detection system is shown in Fig. 1 [10]. The core of the network is the central operating unit (COU). In this part, the main detection algorithms are implemented. In order to observe the electromagnetic field, several intelligent sensors (IS) are connected to the central operating unit. These sensors buffer the monitored electromagnetic field in a continuous mode.

To keep the costs for the overall system as low as possible, the processor capacity of the ISs is limited. Thus, the ISs are not capable of running a sophisticated detection algorithm. Instead, a simple predetection algorithm is implemented in the ISs. Once an IS has a positive match of its predetection, this information is send to the COU. The COU triggers all ISs to send the buffered data and runs a sophisticated detection algorithm on the multiple data streams. As a result, the COU can ascertain what kind of IEMI signal has been found.

Furthermore, the detection algorithm extracts the time of arrival (TOA) of the IEMI signal for each sensor. With these values of TOAs, it is possible to calculate the location of the IEMI source by applying a time difference of arrival (TDOA) algorithm.

3 Sources for the Intentional Electromagnetic Interference

The focus of this work is the detection of intentional electromagnetic interferences. To create such interferences, high field strength values are needed, which couple into the victim system. Sources, capable of producing such field strength values, are, for example, high-power electromagnetic (HPEM) sources. Examples of such sources are shown in Figs. 2, 3, 4, and 5.



Fig. 1 Architecture of the detection system



Fig. 2 PBG3 Pulser by Kentech



Fig. 3 HPEM source DS110 by Diehl



Fig. 4 HPM source SUPRA at the WIS in Munster, Germany



Fig. 5 HPM source made out of a commercial off-the-shelf microwave oven

A very simple way to realize an HPEM source can be the combination of typical test equipment. Figure 2 shows the pulse generator PBG3 by Kentech, which can be combined with any kind of antenna. The generator produces a double exponential pulse (DEP) with an amplitude of 13 kV, a rise time of under 100 ps, and a pulse width of 3 ns [11].

A second kind of source is shown in Fig. 3. This is an HPEM source, industrially manufactured for testing purposes, by the company Diehl. The DS110 produces a damped sine (DS) with a center frequency of 350 MHz and has a pulse duration of 4 ns [12].

The third kind of source has a pulse-modulated cosine (PMC) waveform. HPEM sources realizing this waveform are shown in Figs. 4 and 5. The first example shows the SUPRA of the Bundeswehr Research Institute for Protective Technologies and NBC Protection (WIS) in Munster, Germany. This is a high-power microwave (HPM) source working with center frequencies between 0.675 and 3 GHz. From Fig. 4 it can be seen that this is a very immobile source, technically very sophisticated to build, and thus it is unrealistic that this kind of source is used for an IEMI attack [13]. Nonetheless, this kind of waveform has to be taken into account, especially as a commercial off-the-shelf microwave oven can be converted into an HPM source (cf. Fig. 5) [14].

All these exemplarily shown HPEM sources are potential IEMI sources and thus have to be detected by the detection system presented earlier in this work. However, from the nominal characteristics of these sources, it can already be seen that their behaviors in time and frequency domain differ from each other. This will be examined in the next part.

3.1 Comparison of Real HPEM Signals to Analytical Waveforms

In the previous part, examples of IEMI sources were presented. However, the shown characteristics are examples only, whereas the requirement for the detection system is the identification of *all* threatening signals. Thus, it is practicable to work with analytical functions for HPEM characteristic waveforms. Hence, it will be shown that the waveforms of the exemplary sources correlate with the analytical functions. Furthermore, the correlation coefficients will be presented, serving as a possible indication for the presence of an HPEM source.

The waveform representing the PBG3 is a double exponential pulse (adapted from [15]):

$$g_{\rm dep} = g_0 \cdot k_{\rm dep} \cdot \left(e^{-t/T_{\rm f}} - e^{-t/T_{\rm r}} \right) \quad \forall \quad t > 0, \tag{1}$$

where g_0 is the amplitude, k_{dep} is a correction factor for the amplitude, T_r is the rise time, and T_f is the fall time. The analytical time domain signal for a DEP, representing the PBG3, is shown in Fig. 6, where the analytical signal was optimized with respect to the correlation with the measured signal. In this example, the correlation coefficient between the measured and the analytical signal is 0.95. The fast Fourier transformations (FFT) of the signals are shown in Fig. 7. It can be seen that the signals match in frequency domain as well.



Fig. 6 Double exponential pulse in time domain



Fig. 7 Double exponential pulse in frequency domain

The nominal waveform of the DS110 is a damped sine (adapted from [14]):

$$g_{\rm ds} = g_0 \cdot k_{\rm ds} \cdot \sin\left(2\pi f_0 t\right) \cdot e^{-t/T_{\rm f}} \quad \forall \quad t > 0, \tag{2}$$

where g_0 is the amplitude, k_{ds} is a correction factor for the amplitude, f_0 is the center frequency, and T_f is the fall time of the envelope. The nominal and the measured signals are presented in Figs. 8 and 9. With this realization, the correlation coefficient in time domain is 0.85.

Furthermore, Fig. 8 shows that the measurement of a real DS does not have its peak field strength value at the first half wave. Instead, the envelope of the sine is first increasing and then decreasing again. This is why a DEP is used as the envelope of the sine for a better accordance with the measured signals:

$$g_{\text{deps}} = g_0 \cdot k_{\text{deps}} \cdot \sin\left(2\pi f_0 t\right) \cdot \left(e^{-t/T_{\text{f}}} - e^{-t/T_{\text{f}}}\right) \quad \forall \quad t > 0,$$
(3)

where g_0 is the amplitude, k_{deps} is a correction factor for the amplitude, f_0 is the center frequency, T_r is the rise time, and T_f is the fall time of the envelope. With this modification the correlation coefficient increases to 0.88. The results in time and frequency domain are shown in Figs. 10 and 11.

The third waveform, a pulse-modulated cosine, is defined by the function

$$g_{\rm pmc} = g_0 \cdot \cos(2\pi f_0 t) \cdot e^{-(2t/\alpha)^2},$$
 (4)



Fig. 8 Damped sine in time domain



Fig. 9 Damped sine in frequency domain


Fig. 10 Double exponential pulse-modulated sine in time domain



Fig. 11 Double exponential pulse-modulated sine in frequency domain



Fig. 12 Pulse-modulated cosine in time domain

where g_0 is the amplitude, f_0 is the center frequency, and α controls the width of the envelope (adapted from [15]). The comparison of the measured (SUPRA) and the analytical waveforms is shown in Figs. 12 and 13.

3.2 Comparison of Signals of an EME to Analytical Waveforms

In general, the detection of IEMI threats takes place in electromagnetic environments (EME). This means that in the progress of the detection, the analytical signals are compared to the signals of the EME as well. In this work, the IEMI attacks have to be detected at an airport. The correlations of the analytical waveforms of HPEM sources with the signals from two different kinds of EMEs are exemplarily shown in Figs. 14, 15, 16, and 17. The difference between the general EME and the airport EME is that the general EME does not contain any radar signals. Thus, it can be regarded as an EME, which can also be found outside of an airport.

The results show that especially the radar signals from the airport EME have a high correlation with the sine and cosine HPEM signals. This can be explained by the dominant behavior of the radar signal, which mainly consists of a single continuous wave signal. Therefore, a detection based only on the correlation coefficient is not sufficient in such an EME and would lead to erroneous detections.



Fig. 13 Pulse-modulated cosine in frequency domain



Fig. 14 Correlation of a DEP-modulated sine with a general EME is 0.61



Fig. 15 Correlation of a DEP-modulated sine with an airport EME is 0.95



Fig. 16 Correlation of a pulse-modulated cosine with a general EME is 0.61



Fig. 17 Correlation of a pulse-modulated cosine with an airport EME is 0.98

4 Conclusion

In this work, a sophisticated detection system for HPEM signals is presented. It consists of distributed sensors for the observation of the electromagnetic field and a central operating unit, which executed a detection algorithm. This algorithm is dependent on the analytical waveforms of HPEM signals. In the second part of this work, it is shown that the analytical signals correlate with exemplarily measured HPEM signals, where correlation coefficient of over 0.85 could be achieved. However, the correlation coefficient as the only measure is not enough indication for the presence of HPEM sources, because signals from typical EMEs correlate with the analytical HPEM waveforms as well.

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High-Frequency Impedance Measurement of Electronic Devices Using a De-embedding Technique

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Abstract Many studies have shown the sensitivity of systems to conducted Intentional Electromagnetic Interferences (IEMI) (Giri and Tesche, IEEE Trans Electromagn Compat 46:322-328, 2004; Sabath, Threat of electromagnetic terrorism: lessons learned from documented IEMI attacks. Proceedings of EUROEM 2012, Toulouse, France, 2012) on power grids. In another context, the propagation of spurious compromising emanations in a power grid (Vuagnoux, An improved technique to discover compromising electromagnetic emanations. EMC symposium, Fort Lauderdale, USA, 2010) may result in the leakage of sensitive information. The modelling of the low-voltage network is therefore of fundamental interest since the propagation of electromagnetic waves is highly influenced by the channel behaviour which is related to the cable parameters, connected appliances, etc. First, in this chapter, a de-embedding method is proposed for characterising appliances connected to the power grid by measuring their input impedance from 1 to 100 MHz. Next, a Monte-Carlo simulation is used for analysing the sensitivity of this method to measurement errors. A statistical approach is then proposed for deriving simulated loads from measured ones. Finally, loads are integrated into the CRIPTE Code (Parmantier, Degauque, Topology based modeling of very large systems, Oxford University Press, 1996) developed by the French Aerospace Lab (ONERA) in order to estimate the effect induced by loads on the conducted propagation of a spurious compromising signal that emanates from a computer.

Keywords De-embedding technique • Electromagnetic interferences • High frequency

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1 Input Impedance Measurement of Appliances

1.1 De-embedding Technique Applied for Loads Characterisation

The study presented in this chapter aims at characterising devices fitted with a three-port male plug. In order to connect appliance under test to the network analyser, an adaptor between an N-type connector and a female socket needs to be designed (depicted in Fig. 1).

In our design, we chose to build a three-wire adaptor compliant with the TT ("Terra-Terra") earthing scheme, for which neutral and ground cables are directly connected (this indeed enables to get a framework in which the impedance of appliances can be measured as would be seen in a realistic network).

For performing measurements, thanks to the network analyser, a possible method is to design a dedicated calibration toolkit for moving the plane of reference at the input of the chained system composed by the network analyser and the adaptor. This method is already used in the car industry [1] but requires time to implement as it needs new design for each type of new connectors.

Alternatively the plane of reference can be kept at the input of the network analyser by using the calibration toolkit dedicated to the instrumentation. This is the method proposed in this chapter. Its efficiency essentially relies on the quality of the software algorithm involved to correct the measurements. To specify such an algorithm, we suggest here to use the scattering parameters of the adaptor referenced on 50 Ω . As those parameters are a priori unknown, we designed a two-port system by chaining the *N-type connector female outlet adaptor* (called subsystem *A*, depicted in Fig. 1) to be characterised with an *N-type connector male plug adaptor* (called subsystem *B*). By measuring the scattering parameters S_{ij} of the resulting two-port system, the chained system has been decomposed analytically as the combination of the unknown scattering parameters S_{ij}^A and S_{ij}^B of each subsystem *A* and *B*.



High-Frequency Impedance Measurement...

$$\begin{cases} S_{11} = S_{11}^{A} + \frac{S_{11}^{B} \cdot S_{12}^{A}}{1 - S_{11}^{B} \cdot S_{22}^{A}} \\ S_{12} = \frac{S_{12}^{A} \cdot S_{12}^{B}}{1 - S_{11}^{B} \cdot S_{22}^{A}} = S_{21} \\ S_{22} = S_{12}^{B} + \frac{S_{12}^{B^{2}} \cdot S_{22}^{A}}{1 - S_{11}^{B} \cdot S_{22}^{A}} \end{cases}$$
(1)

The complex system [Eq. (1)] of three equations and six unknowns needs to be completed with at least three additional equations. Those three new equations are obtained by measuring the reflection coefficients S_{11}^{Asc} , S_{11}^{Aoc} and S_{11}^{Bsc} of the open-circuited (oc) and short-circuited (sc) *A* and *B* adaptors. The measurements of the reflection coefficients of the open- and short-circuited adaptors are performed, which corresponds to the theoretical equations below:

$$\begin{cases} S_{11}^{Aoc} = S_{11}^{A} + \frac{S_{12}^{A^2}}{1 - S_{22}^{A}} \\ S_{11}^{Asc} = S_{11}^{A} - \frac{S_{12}^{A^2}}{1 + S_{22}^{A}} \\ S_{11}^{Boc} = S_{11}^{B} + \frac{S_{12}^{B^2}}{1 - S_{22}^{B}} \end{cases}$$
(2)

By solving the 6×6 -dimension system, we get the following values for the female adaptor:

$$S_{11}^{A} = \begin{cases} 2 \cdot S_{11}^{Asc} \cdot S_{11}^{Boc} \cdot S_{11}^{Boc} \cdot (-1 + S_{22}) \\ + \begin{pmatrix} S_{12}^{Asc} - S_{11} \cdot (-1 + S_{22}) \times \\ (S_{11}^{Asc} - S_{11}^{Aoc} + (S_{11}^{Asc} + S_{11}^{Aoc}) \cdot S_{11}^{Boc}) \end{pmatrix} \end{cases} \\ \times 1 / \begin{cases} S_{11}^{Asc} \cdot (-1 + S_{22}) \cdot (-1 + S_{11}^{Boc}) \\ + 2 \cdot S_{11}^{Boc} \cdot (S_{11} + S_{12}^{2} - S_{11} \cdot S_{22}) \\ + S_{11}^{Aoc} \cdot (-1 + S_{22}) \cdot (1 + S_{11}^{Boc}) \end{cases} \end{cases},$$
(3)

$$S_{22}^{A} = \left\{ \begin{array}{c} 2 \cdot S_{12}^{2} + (-1 + S_{22}) \times \\ (-2 \cdot S_{11} + S_{11}^{Asc} + S_{11}^{Aoc} + S_{11}^{Boc} \cdot (-S_{11}^{Asc} + S_{11}^{Aoc})) \end{array} \right\} \\ \times \left. 1 \right/ \left\{ \begin{array}{c} S_{11}^{Asc} \cdot (-1 + S_{22}) \cdot (-1 + S_{11}^{Boc}) \\ +2 \cdot S_{11} \cdot S_{11}^{Boc} \cdot (S_{12}^{2} - S_{22}) \\ +S_{11}^{Aoc} \cdot (-1 + S_{22}) \cdot (1 + S_{11}^{Boc}) \end{array} \right\},$$
(4)



Fig. 2 Characterisation of a 75 Ω reference load

$$S_{12}^{A} = \begin{cases} -\left(i \cdot \sqrt{2} \cdot \sqrt{S_{11}^{Boc^{2}} - 1} \cdot \sqrt{S_{11}^{Asc} - S_{11}^{Aoc}}\right) \times \\ \sqrt{\left(-(S_{11} \cdot S_{12}) + S_{11} + S_{12}^{2} + (S_{22} - 1)\right)^{2} \cdot S_{11}^{Aoc}} \cdot S_{11}^{Asc}} \\ \times 1 / \sqrt{\left(\frac{2 \cdot S_{11}^{Boc} \cdot \left(-(S_{11} \cdot S_{22}) + S_{11} + S_{12}^{2}\right) \\ +(S_{22} - 1) \cdot S_{11}^{Asc} \cdot \left(S_{11}^{Boc} - 1\right) \\ +(S_{22} - 1) \cdot S_{11}^{Aoc} \cdot \left(S_{11}^{Boc} + 1\right)}\right)^{2}}$$
(5)

We checked the soundness of this method for measuring the *S*-parameters of a 75 Ω resistance. First we confirmed by experimentation that the introduction of the female adaptor implies a correction. Secondly we showed how the adaptor scattering parameters can be used to perform this correction. We carried out measurements in the following two test configurations:

- (Reference test) The impedance is retrieved by measuring its reflection coefficient using the network analyser alone.
- (Measured test) The same procedure is performed by using the chained system composed of the female adaptor and the network analyser.

As expected, it can be observed in Fig. 2 that the adaptor has a little effect below 35 MHz since the induced error remains below 10 % (difference between the measured and reference curves). On the other hand, above this frequency a correction is required. This was done by applying a de-embedding technique. It consists in measuring the impedance in a test fixture and in removing the effects of the female adaptor (subsystem A). This leads to the following correction formula:



Fig. 3 Statistical analysis of the de-embedding technique

$$S_{11}^{\text{De-embedded}} = \frac{S_{11}^{\text{Measured}} - S_{11}^{A}}{S_{22}^{A} \cdot (S_{11}^{\text{Measured}} - S_{11}^{A}) + S_{12}^{A}2}.$$
 (6)

By applying the previous formula on the measured data, it can now be verified that the error has been significantly reduced (difference between the reference and de-embedded curves) in Fig. 2. Moreover, it seems that it tends towards the reference value at high frequency. Additionally we performed the experiment for several measurements, and we observed that the resulting mean has the same behaviour as plotted in Fig. 3. This allows us to conclude that the remaining error is independent of the experiment set-up.

1.2 Sensitivity of the Proposed Method to Errors in Measurement

When solving the non-linear system, both errors in phase and magnitude measurements are propagating along the complex computation of the scattering parameters of the female adaptor. In order to estimate the robustness of the de-embedding procedure in presence of measurement noise, a Monte-Carlo approach has been applied. Error, documented by the network analyser manufacturer, for phase measurement is given at \pm 5 ° and the error in dB level is given at \pm 10 % around the measured parameter. This leads us to define the magnitude error as $|S_{ij}^{\text{Simulated}}| = |S_{ij}^{\text{Measured}} \cdot (1 + \epsilon)|$ with ϵ following a normal law with a zero mean and a 0.1 standard deviation. On the other hand, the phase error was simulated in degree as

 $\operatorname{Arg}(S_{ij}^{\operatorname{Simulated}}) = \operatorname{Arg}(S_{ij}^{\operatorname{Measured}}) + \beta$ with β following a uniform distribution over $[-5^\circ; 5^\circ]$. Several instances of the random variables ε and β have been generated to design a set of simulated measurements at each frequency regarding the *type B contributions* [2].

When compared with the direct reflection coefficient measurements (reference and measured curves), the measurement errors on the resulting simulated impedances only induce a variation of 10 % around the mean values. However, the resulting de-embedded impedance (corrected) is sensitive to the propagation of errors during the 6×6 system solving. This is particularly true below 35 MHz, where the variance around the mean value can reach a maximum of 1,000 %. This discrepancy is essentially due to phase error measurements. The mean μ and standard deviation σ were computed and plotted in Fig. 3. It can be observed that both reference and de-embedded curves overlap quite well.

We deduce that non-corrected measurement from 1 to 35 MHz can be combined with the de-embedded impedance results to extend the frequency band of characterisation up to 100 MHz.

1.3 Derivation of Loads from Measured Impedances

The resulting mean and standard deviation in Fig. 3 suggest that the random impedance variables follow different distribution for each frequency. In order to infer those distributions, we estimated the four first-order moments. The final purpose was to derive a relationship between the characteristics of these distributions and the frequency. With such relationship in hand, we should be able to derive theoretical distribution of loads and hence to reduce the number of appliances to be measured. Loads (computers, printers, lights, etc.) were derived by using the appropriate inferred distributions at each frequency, and the generated appliances have been inserted into an electromagnetic software simulator. Finally, we succeeded in analysing the effects induced by the loads on the electromagnetic wave propagation [3, 4] in a simulated power-grid network.

2 Effects Induced by Loads on the Propagation of Spurious Compromising Emanations

2.1 The CRIPTE Code Developed by ONERA

The electromagnetic topology formalism [5, 6] allows the decomposition of a non-uniform complex network into a set of connected uniform subnetworks by means of tubes and junctions. This principle is here applied for a three-metre-long test network for which the transmission lines are the tubes and the connecting



points are the junctions as well as the input and output points in the network, as depicted in Fig. 4.

2.2 Cable Modelling, Multiconductor Transmission Line Method

The low-voltage network is based on a three copper wire harness generally named Phase (P), Neutral (N) and Ground (G). It is assumed that those copper wires have a 1.5 mm² cross-section. The Multiconductor Transmission Line Theory [7] method can be used, assuming that the height of the cables above the ground is smaller than the shortest wavelength, the dielectric losses are negligible and a quasi-TEM mode is propagated. By using the CRIPTE code, the per-unit length parameters for the cross-sections have been obtained.

The per-unit length parameters can be computed for random cable cross-sections assuming that the raceway (not handled in the electromagnetic model) is defined at 3.5 cm above the ground. The test network is a part of a larger distribution network. The reference of the MTL model is the Ground cable.

To simulate the connection of this larger network at the J6 junction, an equivalent input impedance of a power-grid network has been measured separately and allocated to the J6 junction.

2.3 Effect of Loads on EM Wave Propagation Along Cables

The junction J1 is characterised using a measured impedance of a computer for analysing the propagation of a 1 V spurious compromising [8] signal level (V_{computer} , Z_{computer}).

Appliances have been randomly picked up from the database generated according to the method detailed in Sect. 1. Assuming a fixed network, the variation



Fig. 5 Mean (μ) and standard deviation (σ) of the computed output current level



Fig. 6 Skewness and kurtosis of the computed output current levels

of the current induced by those appliances (connected at junctions J4 and J5) has been computed at the J6 junction. The mean value μ and the standard deviation σ can also be computed at each frequency step and are depicted in Fig. 5. When load impedances are low, the cable per-unit length parameters may have a notable effect on the current level. The distribution of random values at each frequency step can be inferred.

As a result, it has been observed in Fig. 5 that the mean and standard deviation are function of frequency. The third moment (also called skewness) and the fourth moment (also called kurtosis) can also be used for describing the shape of distributions. The analysis of those moments [9] shows that random values follow different types of distribution at each frequency.

The skewness is used for describing the asymmetry of distribution. When it is negative or positive, it can be assumed that respectively the random values are skewed to the left or to the right of the mean value. The analysis of kurtosis aims at measuring the "peakedness" and the tail heaviness of a distribution. When the resulting skewness is close to 0 and the kurtosis level close to 3, as it is the case at high frequencies in Fig. 6, a Kolmogorov-Smirnov good-of-fitness test has confirmed that the distribution is a Normal one.

3 Conclusion

Propagation of waves is directly affected by connected appliances and by the network topology. In this chapter, the influence of load variability has been highlighted for a fixed architecture. It appears that it has a significant effect on the propagation of electromagnetic waves along cables. Hence, it is crucial that modelled networks integrate measured or realistic simulated values of appliances. In this chapter, a method based on a de-embedding technique for characterising the input impedance of appliances from 1 to 100 MHz has been proposed. Modelled low-voltage distribution network stressed by a conducted high-power electromagnetic pulse or by a spurious compromising source can be simulated. The signal level induced by those sources can be computed at each socket. By analysing the impedance of low-voltage networks, the transferred energy between sources, cable networks and loads can be estimated. This is, for instance, of particular interest to study the propagation of electromagnetic interferences in power grids.

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Automated and Adaptive RF Effects Testing

E.G. Farr, L.H. Bowen, W.S. Bigelow, R.L. Gardner, and P. Finlay

Abstract Testing electronics for vulnerability to radio frequency (RF) radiation is time consuming, due to the large number of source variables of interest, including center frequency, pulse width, pulse repetition frequency, number of pulses, and bandwidth. One must intelligently select the source parameters most likely to expose the greatest vulnerability. We do so here using standard techniques from minimization theory. Within a space of two or more variables, we search for the combination that upsets the system at the lowest power or field level. We investigated the vulnerability of media converters to pulsed RF fields, by pinging a remote computer.

Keywords Radio frequency (RF) effects • Electric field upset threshold • Automated testing • Minimization algorithm • Media converter (MC)

1 Introduction

The vulnerability of electronics to radio frequency (RF) fields has been well documented, for example, in [1, 2]. This has led to a major effort to test electronics to find the minimum field or power at which an effect is observed. However, such testing is time consuming, due to the large number of source variables of interest. One typically searches for the minimum electric field that causes upset, as a function of center frequency, pulse width, pulse repetition frequency, number of

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pulses, and bandwidth. It is impossible to test all combinations of all the variables, so one must intelligently select the source parameters most likely to expose the greatest vulnerability.

To select source parameters, we propose using standard techniques from minimization theory. Within a space of two or more variables, we search for the combination that upsets the system at the lowest power or field level. We begin by measuring the vulnerability levels on a coarse grid and then fit a surface to the measured data. We then find the minimum of the surface and measure the vulnerability at the minimum. With the new data, the process repeats itself iteratively until it converges.

Ideally, the entire process can be automated. The source variables can all be controlled electronically. In addition, one can determine automatically whether the test object has been upset and send a reset command if necessary. This leads to a completely automated system that intelligently selects the test parameters, monitors the status of the device, and converges on a minimum upset threshold. During this first implementation, some manual operations were required; however, these can be automated at a later date.

In this project, we investigated the vulnerability of media converters (MCs) to pulsed RF fields. MCs are network devices that convert signal on Cat 5 Ethernet cable to optical fiber and are known to be vulnerable. We tested these devices by pinging a remote computer and observing the field levels at which the pings failed to return. We searched a space of source variables and converged on a minimum upset threshold. Most of the operations were carried out automatically. Note that this paper is a condensed version of [3].

2 Experimental Setup

The MC we tested was the IMC model TP-TX/FX-MM850-ST, operating at 850 nm. This was selected because of its low cost and easy availability. A photo of this MC is shown in Fig. 1. When configured for testing, the MC requires connections for two optical cables: an Ethernet cable and a power cable. The two optical cables are necessary in order to communicate in both directions.

We tested the MCs in the configuration shown in Figs. 2 and 3. The main computer, running LabVIEW code, pings a remote computer through four MCs, two lengths of fiber optic cable, and three lengths of Cat 5 network cable. The computer controls the parameters of the synthesizer, which drives the amplifier that feeds into the TEM cell. Software running on the main computer then pings the remote computer, listens for the return signal, and detects a failure to respond.

This configuration, which tests two MCs concurrently, was chosen in order to limit RF leakage from the TEM cell. The penetrations of the TEM cell, shown in Fig. 4, were either fiber optic cables or filtered DC power cables, both of which



Fig. 1 The IMC model TP-TX/FX-MM850-ST media converter



Fig. 2 Experimental setup to test degradation of MC performance when exposed to RF

could be configured to maintain the integrity of the RF shield. Previously, we tested a configuration with a single MC in the TEM cell; however, the metallic Cat 5 cable that penetrated the wall of the TEM cell caused excessive RF leakage. Attempts to limit this leakage by filtering the Cat 5 cable at the wall of the TEM cell resulted in loss of network signal.

The interior of the TEM cell is shown in Fig. 5. Note that the interior is somewhat cluttered with cables. Changing their exact position might affect the outcome of the measurements.

It is necessary to relate the power out of the amplifier to the field in the TEM cell. To do so, we observe that power at the input is converted to voltage as



Fig. 3 Photo of the experimental setup





$$P_{\rm p} = \frac{V_{\rm p}^2}{2 \times 50 \ \Omega}, \quad V_{\rm p} = \sqrt{P_{\rm p} \times 100 \ \Omega}, \tag{1}$$

where P_p is the peak power and V_p is the peak voltage in the sine wave on the 50- Ω feed line. The peak field in the TEM cell is related to the voltage at the input as $E_p \approx V_p/h$, where h = 0.22 m, the plate separation in the test volume. Finally, the





average power is $P_{\text{avg}} = P_{\text{p}} \times \text{DF}$, where DF is the duty factor, DF = PW × PRF, and PRF is the pulse repetition frequency. The duty factor is the fraction of time the square pulse of CW energy is turned on.

The software consists of two pieces of code: a threshold detector and a minimization routine. The threshold detector determines automatically the minimum field level required for upset for a given set of source parameters. It sets the frequency, pulse width (PW), and pulse repetition frequency (PRF) of the synthesizer. The power level is initially set to a low value and is gradually incremented. At each increment, the remote computer is pinged 20 times. When the power is high enough to yield 3 failures out of 20 pings, it is considered an upset condition, and that level is the upset threshold. This entire procedure is carried out in code that was written in LabVIEW.

The minimization routine guides the selection of parameters to test, in order to iterate to find a minimum upset threshold. We choose two variables over which to search, pulse width and either duty factor or frequency. The procedure begins by taking data at nine points in the data space, the minimum, center, and maximum of each variable. A surface is fitted to this initial set of data using the "fminsearch" function in MATLAB, which then finds the minimum of the surface. This minimum is then used as the next point to test. The new results are added to the previous data, a new surface is fitted to the data, and a new minimum is found. The process repeats until the result converges.

3 Experimental Data

A key goal was to locate a minimum in the middle of a vulnerability test space. In a number of early experiments, we found minima at less interesting locations—at either a corner or edge of the test space. However, finding a minimum in the





middle of the test space demonstrates the usefulness of our minimization algorithm. We do so here.

We tested the vulnerability of two media converters in our TEM cell, using the configuration shown in Fig. 2. We tested with a pulse width (PW) of 10 μ s, with frequency (*f*) ranging over 700–900 MHz and duty factor (DF) ranging over 0.1–10 %. The original nine points in the space are shown in Fig. 6 (top) and a fit to the surface is shown below. Data are plotted in terms of peak power units. We have left the power units arbitrary intentionally. We then iterated three more times to find the minimum, and the result is shown in Fig. 7. The minimum converges to a frequency (*f*) of 805.1 MHz and duty factor (DF) of 7.8 %, where we found a peak power (*P*_p) of 0.010, based on the curve fit. We then measured a point very close to the minimum, at *f* = 800 MHz and DF = 8 %, where we found *P*_p = 0.018.

Thus, we observed $P_p = 0.010$ in the curve fit, and we measured $P_p = 0.018$ very nearby. This is a little less accurate than we would like to see, but it is still very good. We should be able to improve the fit of the surface to the data by giving greater weight to data that is closer to the minimum. This is appropriate, since it is that portion of the surface in which we are most interested. We can do so, for example, by fitting to the inverse square or inverse cube of the vulnerability function.



4 Future Work

We outline here a number of areas that would benefit from further work. First, we would fully integrate the software into a single program. Currently, the software exists in two separate programs, which leads to manual operations. These programs have to be integrated in order to realize a fully automated system.

Second, we would add an automatic power characterization to each measurement, using a directional coupler and oscilloscope. This would involve having the software talk to the oscilloscope, downloading the voltage waveform, and converting the measured voltage to peak power. Ideally, one would prefer having an amplifier whose power is described by its dial settings, however, that seems to be difficult to realize in practice.

Third, we would investigate a number of variations on our minimization algorithm. For example, we would investigate alternative surface functions to fit to our data. In this paper, we used a function of the form

$$z = (ax^{2} + bx + c)(dy^{2} + ey + f),$$
(2)

where x and y are the two variables over which we are minimizing and a, \ldots, f are the unknown coefficients that are chosen to give the best-fit surface to the measured data. However, many other functional forms are possible.

Fourth, we would investigate methods of giving greater weight to the function value near its minimum. The current method simply implements a least-mean-square fit to the measured data. However, the data close to the minima are of greater interest, so it is more important to reduce the fitting error in that region. To emphasize the minima, one might fit a surface not to the data itself, but to its inverse square or inverse cube. By this method, errors near the minima carry more weight and therefore are reduced in the fitted function.

Fifth, note that in this project we searched a two-dimensional space for the minimum upset threshold. However, this technique should apply equally well to searches in higher-order spaces, and this should be examined.

Finally, we would test a variety of other devices, which might include cell phones, iPods, and/or network routers. The idea here would be to incorporate alternative upset modes and reset mechanisms into the programming. One could detect an upset by listening (electronically) for the music on a telephone or iPod to stop. One could also detect when a screen goes dark with a photodetector. One could reboot a system after upset by electronically toggling a power switch. One could use a servomotor to twist a knob on a source.

5 Conclusion

We have automated the testing of media converters for vulnerability to RF effects. Our testing involved pinging a remote computer and listening electronically for missing return signals. To do this, we used software written in LabVIEW and MATLAB. The most important result is that we have successfully observed a minimum in the middle of a test space. This is the first nontrivial use of the minimization algorithm, so it is a significant milestone.

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Pockels' Effect-Based Probe for UWB and HPEM Measurements

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Abstract The aim of this paper is to give a qualitative and quantitative comparison between the commonly used transducer dedicated to electric (E) field measurement. A review of the several kind of sensor (antenna, electro-optics, bolometer, IR thermography,...) is firstly proposed and respective diagrams of merit are given. These later ones include parameters such as sensitivity, dynamics, spatial and temporal resolutions, frequency bandwidth, selectivity to one vector component of the E field and invasiveness of the measurement. This comparison finally demonstrates the suitability of electro-optic (EO) sensors to perform fully vectorial characterization of the E-field. The theoretical principles of EO technique are then reminded, from the E-field induced birefringence modifications of a non-centrosymetric crystal, to the exploited modulation to finally extract the proper information about the E-field vector. The last section deals with experimental realizations of EO probe as well as to a pulsed E-field characterization.

Keywords Electro-optics • HPEM measurements • Vectorial characterization • Optical fiber • Electric-field sensor

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1 Introduction

Although antennas currently constitute the reference for E-field characterization, they still involve many drawbacks linked to their conductive composition. Indeed, antennas exhibit high sensitivities to the detriment of high induced perturbations and relatively low bandwidths. Requirements for HPEM and UWB characterization are perfectly fulfilled by pigtailed EO sensors (fully dielectric probes with adapted sensitivity). Indeed, the intrinsic bandwidth of such transducers reaches more than 10 GHz and the associated dynamics exceeds 100 dB. Their spatial resolution belongs to the submillimetric range. Moreover, pigtailed EO sensor realizations developed from the beginning of the century have demonstrated non-invasive remote measurement up to more than 10 m. The chapter is divided in three main sections. We initially propose a comparison between the most usual tools used to perform E-field characterizations. This review includes linear sensor as well as quadratic ones. The next sections are dedicated to a specific insight on EO probes, from principle and design to their experimental use to perform transient E-field measurement.

2 E-Field Sensor Comparison

Here is given a comparative analysis based on an exhaustive bibliography study. The E-field measurement techniques taken into account are the following ones:

- Infrared Thermography which is based on the E-field-induced heating of a photothermic film. This technique gives a direct 2D mapping representing the transverse distribution of the EM energy;
- Bolometer which characterizes the EM-associated power via a temperature measurement of an absorptive material (*e.g. layer of metal*);
- · Passive antenna leading to a vectorial measurement of the E-field;
- Franz–Keldish effect which is based on the modification of a semi-conductor band diagram and leading to a quadratic response relatively to the E-field;
- EO probe, based on the Pockels' effect, performing a vectorial measurement of the E-field vector components.

In order to give an objective analysis of the sensors characteristics, the comparison criteria are defined in Table 1.

In the case of a quadratic transducer (*e.g.* bolometer), the response α (V/W) has to be taken into account instead of the undimensioned linear response \Box^2 for the calculus of the spatial and temporal resolutions.

Two tables summarize this analysis, the first one presenting the references of the comparison (Table 2) and the second one dedicated to the experimentally achieved characteristics of the sensors (Table 3). The chosen criteria illustrating advantages and drawbacks of the transducers are sensitivity, dynamics, E-field vector

Characteristic	Symbol (unit)	Definition	
Sensitivity	$E_{\rm min}~({\rm V~m}^{-1}~{\rm Hz}^{-1})$	$\overline{V}_{\rm eff}(E_{\rm min}) = \sigma(V_{\rm eff}(E_{\rm min}))$	
Dynamics	ΔE (dB)	$20 \log_{10} \left(\frac{E_{\max}}{E_{\min}} \right)$	
Bandwidth	$\Delta f_{-3 \text{ dB}}$ (Hz)	$f_h - f_l$	
Temporal resolution	t_{\min} (s)	$\sqrt{\frac{\int_{-\infty}^{+\infty} \Re(E,t)^2 \cdot (t-\overline{t})^2 dt}{\int_{-\infty}^{+\infty} \Re(E,t)^2 dt}}$	
		with $\overline{t} = rac{\int_{-\infty}^{+\infty} \Re(E,t)^2 \cdot t dt}{\int_{-\infty}^{+\infty} \Re(E,t)^2 dt}$	
Spatial resolution (Cartesian)	$x_{\min}, y_{\min}, z_{\min}$ (m)	$x_{\min} = \sqrt{\frac{\int_{-\infty}^{+\infty} \Re(A_{\mathrm{eff}}, x)^2 \cdot (x - \overline{x})^2 dx}{\int_{-\infty}^{+\infty} \Re(A_{\mathrm{eff}}, x)^2 dx}},$	
		idem for y_{\min} and z_{\min}	
Spatial resolution (cylindric)	$ \rho_{\min}\left(\mathbf{m}\right) $	$\sqrt{\frac{\int_{-\infty}^{+\infty} 2\pi\rho \Re (A_{\rm eff},\rho)^2 \cdot \rho^2 d\rho}{\int_{-\infty}^{+\infty} 2\pi\rho \Re (A_{\rm eff},\rho)^2 d\rho}}$	
Selectivity	$S_{E_{ }}$ (dB)	$20\log_{10}\left(\min\left[rac{\Re(E_{\parallel})}{\Re(E_{\perp})} ight] ight)$	
		with $E_{ }\perp E_{\perp}$	

Table 1 Definition of the characteristics taken into account for the sensor comparison

Table 2 Intrinsic properties of the studied techniques and associated references	Technique	E measurement	References
	IR thermography	Quadratic	[1–14]
	Deported bolometer	Quadratic	[15-23]
			[24–33]
	Passive antenna	Linear and	[34-41]
		vectorial	[42, 43]
	Franz-Keldysh probe	Quadratic	[44-47]
	EO probe	Linear and	[48–58]
		vectorial	[59-66]

component selectivity, invasiveness, spatial resolution, temporal resolution and bandwidth.

The scale of the different characteristics corresponds to the use of the sensor in the frequency domain (*i.e.* CW regime). The time domain behaviour of the transducers implies a wider resolution bandwidth and hence to an increasing value of the minimum detectable field and a decline of both the dynamics and the selectivity.

Except concerning the sensitivity, EO-based probes present homogeneous and rather good performances. However, in case of high power and pulsed signals, involving high-intensity E-field, this rather poor sensitivity of the EO sensor does not constitute a drawback anymore.



 Table 3
 Performance diagrams of the studied sensors related to Table 2

 S_{Emin} is the sensitivity, S_{Emax}/S_{Emin} is the dynamics, $S_{//}/S_{\perp}$ is the selectivity, Δ_{Eact} is the invasiveness, ρ_{min} is the spatial resolution, τ_{min} is the temporal resolution and B W is the bandwidthas the ratio between f_h and f_l , the cut off frequencies. *Black lines* correspond to a linear behaviour of the transducer, while greyones correspond to a quadratic response

3 Electro-Optic Principles

The Pockels' effect, also called linear EO effect, traduces the proportional variation of the refractive indices of non-centrosymetric crystals with the E-field. Duvillaret et al. introduced the sensitivity vector **K**, inherent to any EO crystal and linked to the orientation of a laser probe beam. **K** gives the canonical and vectorial link between the E-field **E** to be measured and the induced modification of an eigen refractive index n_{\pm} :

$$\delta n_{\pm} = \mathbf{K} \cdot \mathbf{E} = \nabla n_{\pm}(\mathbf{E})|_{\mathbf{E}=0} \cdot \mathbf{E}$$
(1)

The exhaustive description of **K** can be found in [65]. As an example, $|\mathbf{K}|$ takes the value of 120 pm/V for a LiTaO₃ crystal probed along it *Y* eigen axis.

The E-field-induced modulation of the refractive indices is probed by a laser beam (wavelength $\lambda = 1.55 \,\mu$ m). This latter one sees its polarization state modified during its travel forth and back within the EO crystal. The polarization state modulation is then analysed; thanks to waveplates and polarisers and the final conversion into an analog electric signal is ensured by a high speed photodiode. The obtained modulated photocurrent \tilde{I} is linear with the E-field and is written:

$$\tilde{I} = \eta \bar{P} \frac{2\pi L_{\text{crystal}}}{\lambda} \mathbf{K} \cdot \mathbf{E}$$
⁽²⁾

 \overline{P} is the mean optical power (1–10 mW), $L_{crystal}$ is the physical length of the crystal (10 mm) and η is the photodiode responsivity (0.8 A/W). Let us notice that the typical values given for \overline{P} guarantee no additional optical noise contributions (Relative Intensity Noise and shot noise) to the intrinsic noise level of visualization device such as a fast oscilloscope or a spectrum analyser [69].

4 Experimental Realization

4.1 Pockels' Effect-Based Probe and Associated Measurement Setup

The EO sensor is made of stoechiometric LiTaO₃ monocrystal (*Y*-cut), this latter one constituting the E-field transducer. Its refractive indices are linearly modified by the E-field; thanks to the Pockels' effect. This modification is probed with a circularly polarized laser beam ($\lambda = 1,550$ nm). This circular polarization state is obtained with an inner probe quarter waveplate. A gradient index (GRIN) lens shapes the intra-crystal beam geometry. The sensor is pigtailed with a polarization maintaining (PM) fibre. The fully dielectric composition drastically limits its invasiveness and also guarantees its immunity to a potential air breakdown (see Fig. 1).



Fig. 1 Schematic of the EO probe (a) and photography a dielectric coated sensor (b)

This sensor is linked to a servo-controlled optoelectronic setup (Kapteos[®] system). An accurate wavelength tuning (3 nm around 1,550 nm) allows a realtime compensation of the fibre dephasing. The whole EO setup exhibits a temperature-dependent-free measurement, a bandwidth ranging from 10 kHz up to 18 GHz, a dynamics exceeding 100 dB and a rejection of orthogonal E-field components higher than 35 dB. Moreover, last developments allow simultaneous measurement of two E-field orthogonal components; thanks to a unique isotropic EO crystal (ZnTe, BSO, CdTE, ...) and a single laser probe [67, 68].

4.2 Short Pulse Measurement

In order to demonstrate the potentialities of EO sensors, we here propose a short pulse comparative measurement. A pulsed generator delivers a peak voltage of 4 kV with a pulse duration of a few nanoseconds. The E-field propagates along a GTEM (Gigahertz Transverse ElectroMagnetic cell) cell (see inset of Fig. 2b). An asymptotic Conical dipole (ACD) constitutes the reference sensor and is located on a predefined position within the GTEM cell (onto the lower base of the structure). As this antenna is derivative, the signal, shown in Fig. 2, has been integrated to be compared with the EO sensor's one.

A good agreement is observed between the two measurements. Nevertheless, a delayed signal is noticed close to the end of the time span. This is due to a reflection of the field inside the GTEM cell. This parasitic reflection is measured at the different positions of the two sensors.

The measured rise time is slightly steeper with the EO probe. The Fourier transform of the two measured signals is shown in Fig. 2b and confirms the agreement between the two measures.



Fig. 2 Measurement of ns-pulsed E-field in a GTEM Cell fed by a Kentech[®] generator (4 kV) with an ACD sensor (*blue line*) and with the pigtailed (30 m) EO sensor (*red line*). Transient evolution of the pulse (**a**) and associated Fast Fourier transforms (**b**)

5 Conclusion

The proposed comparison between the available sensors dedicated to E-field measurement leads a first conclusion: antennas and EO probes are the unique tools providing a vectorial measurement of the field. While the use of metallic antennas remains the most widespread technique, EO sensors propose greater performances, especially in terms of dynamics, bandwidth and invasiveness. EO measurement is becoming mature technique since EO sensors are actually pigtailed and temperature-dependent free. Their intrinsic wide bandwidth (ranging from

quasi DC to several tens of GHz) and their dynamics (exceeding 100 dB) are fully suitable for HPEM and UWB measurements. Moreover, as they are fully dielectric and deported, thanks to an optical fibre, the induced disturbance on the field to be measured is very low comparatively to metallic antennas. This latter properties can be essential for some application such as antenna radiation pattern, on chip diagnostic or even disruptive field evaluation. Some recent improvements of EO demonstrate the real time and rigorous polarimetric measurement of a single-shot intense pulses.

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Part VII UWB Sensing

Estimating Magnetic Polarizability Tensor of Buried Metallic Targets for Land Mine Clearance

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Abstract This chapter addresses the problem of identifying metallic objects in buried land mines and discriminating them from clutter using low-frequency electromagnetic induction (EMI) techniques. From dipolar fields, the magnetic polarizability tensor extracted from the target response can be used as a basis for identification. Here, a deterministic nonlinear optimization method is presented to estimate target polarizability matrix and location by fitting a dipole model to EMI data collected above target in a least squares sense. Using finite element simulated data with added synthetic low-frequency noise (10 dB SNR), results show initial guess misestimating target position with few centimeters in the transversal (x, y) plane can be corrected very close to the true location. The method is also able to estimate the polarizability tensor to within 12 % error of the true tensor.

Keywords Component • Electromagnetic induction • Land mines • UXO • Magnetic polarizability • Nonlinear inverse problems

1 Introduction

Recently there has been research interest in improving metal detectors for identifying buried metallic targets such as land mines. This research has been mainly supported by humanitarian agencies and defense organizations for the purpose of clearing land mines from contaminated environmental fields. Conventional metal detectors can typically detect metal components within land mines, even low-metal antipersonnel (AP) devices, but cannot discriminate them from harmless clutter.

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This leads to higher rates of false alarms and hence, needless excavations, making the clearance procedure less effective and more costly. The challenge within this research initiative is to develop metal detectors with novel electromagnetic induction (EMI) techniques able to provide target identification capabilities.

When the target distance to the sensor is much larger than its characteristic size, typical in land mines detection, the excitation field is approximately uniform in the target region. In this case, dipolar fields are assumed to adequately describe the target response. Based on this assumption, researchers have proposed various approaches using the EMI data in attempt to identify buried land mines. One approach is based on measuring the spectral response at a single sampling location above the target over a broadband bandwidth [1]. This response is then compared with a library of stored spectra from known land mine types. Although it may aid in the discrimination process, this approach has the limitation that the obtained spectra is not unique to the probed target; that is, one can find a target of different shape and EM properties that can produce the same spectral response. Another approach employs magnetic singularity identification method originally developed by Baum [2]. In this procedure, EMI data can be used to extract target's singularities in the form of negative real poles or equivalently exponential decay time constants for time-domain systems. These parameters are dependent on target's size, shape, conductivity, and permeability. A related technique derives the magnetic polarizability tensor **M** by magnetically polarizing the target from different orientations. This matrix is unique to the target and similarly holds information about its geometry and material composition. In this chapter, we have adopted this latter approach as a tool to be used for target identification.

A few studies have previously evaluated the multi-axis magnetic polarizability generally by fitting a dipole model to the target response using least squares methods [3, 4]. Others have used these methods to solve directly for the electrical parameters, size, and depth of the target [5]. Various optimization methods have been developed to solve least squares problem. Here, we apply a nonlinear search algorithm that can calculate the magnetic polarizability tensor of the target and simultaneously yields a correction to target location in the ground.

2 Theory

When a conductive target is interrogated with a time-harmonic sinusoidal primary magnetic field, eddy currents are induced inside the target. These currents can be characterized by a series of multipole moments. The associated secondary field can in turn similarly be represented as a sum of corresponding multipole terms. For a magnetic multipole expansion of order *n*, the terms will have field strengths falling off as $1/r^{n+2}$, at distance *r* away from the currents. The lowest order term in this series is called the dipole term. If the target if far enough from the sensor, the secondary field can be approximated by an induced dipole moment. For a sensor

with colocated excitation and receiver coils, the response measured by the sensor due to the induced dipole can be expressed using reciprocity principles as follows:

$$\mathbf{Z}(\mathbf{r},\omega) = \mathbf{H}^{\mathrm{R}}(\mathbf{r}) \bullet \mathbf{M}(\omega) \bullet \mathbf{H}^{\mathrm{T}}(\mathbf{r}), \tag{1}$$

where $\mathbf{Z}(\mathbf{r},\omega)$ is the complex mutual impedance induced in the coil and \mathbf{H}^{T} and \mathbf{H}^{R} are the field intensity vectors from the excitation and the receiver coils, respectively, at the target location, \mathbf{r} . In Eq. (1), \mathbf{M} is a 3 × 3 complex valued, symmetric matrix that is frequency dependent, which denotes the magnetic polarizability tensor characterizing the target. The fields \mathbf{H}^{T} and \mathbf{H}^{R} , however, can be regarded as frequency invariant since the skin depth in the medium in which the target is buried is usually long compared to the target's depth. This should be valid for typical EMI frequencies ($\leq 100 \text{ kHz}$), soil conductivities ($\leq 1 \text{ S m}^{-1}$), and typical AP land mine depths (up to 20 cm). Therefore, using the dipole model, the frequency and target positional dependences in the EMI response are separated where **M** is invariant with target position but contains its spectral behavior, and the fields are only a function of target position. This has the advantage to help in the estimation of the target's tensor and location.

We propose to use a nonlinear iterative algorithm to evaluate the magnetic polarizability tensor of the target and optimize estimation of its location. An initial guess of location can usually be inferred from target response pattern or some other technique of measurement. The tensor is also assigned an initial estimate, for example, a complex unity matrix. The algorithm starts with these two guesses and fits the dipole model to the EMI data to reconstruct an estimate of the polarizability tensor. This latter is then fed into the model but this time to produce a new estimate of target location. Having these first reconstructions of the target's tensor and location, the algorithm proceeds with an iterative refinement process of these parameters. In each iteration, where an update of \mathbf{M} is generated, the algorithm uses this estimate in a nested iterative process to search out for an optimum update of target location that fits well the model to the data.

Before we proceed to describe the outline of the formulations used by the optimization method in solving for the tensor and location of the target, let us reformulate the dipole model into a simpler expression involving the product of two vectors, namely a tensor and a field vector. Expanding Eq. (1) yields

$$\mathbf{Z} = \begin{bmatrix} \mathbf{H}_{x}^{\mathrm{R}}, \mathbf{H}_{y}^{\mathrm{R}}, \mathbf{H}_{z}^{\mathrm{R}} \end{bmatrix}^{\mathrm{T}} \cdot \begin{bmatrix} M_{11} & M_{12} & M_{13} \\ M_{21} & M_{22} & M_{23} \\ M_{31} & M_{32} & M_{33} \end{bmatrix} \cdot \begin{bmatrix} \mathbf{H}_{x}^{\mathrm{T}}, \mathbf{H}_{y}^{\mathrm{T}}, \mathbf{H}_{z}^{\mathrm{T}} \end{bmatrix}^{\mathrm{T}}$$

$$= M_{11} \underbrace{\left(\mathbf{H}_{x}^{\mathrm{T}}\mathbf{H}_{x}^{\mathrm{R}}\right)}_{\text{hxx}} + M_{12} \underbrace{\left(\mathbf{H}_{x}^{\mathrm{T}}\mathbf{H}_{y}^{\mathrm{R}} + \mathbf{H}_{x}^{\mathrm{R}}\mathbf{H}_{y}^{\mathrm{T}}\right)}_{\text{hxy}} + M_{13} \underbrace{\left(\mathbf{H}_{x}^{\mathrm{T}}\mathbf{H}_{z}^{\mathrm{R}} + \mathbf{H}_{x}^{\mathrm{R}}\mathbf{H}_{z}^{\mathrm{T}}\right)}_{\text{hxz}} + M_{22} \underbrace{\left(\mathbf{H}_{y}^{\mathrm{T}}\mathbf{H}_{y}^{\mathrm{R}}\right)}_{\text{hyz}} + M_{23} \underbrace{\left(\mathbf{H}_{y}^{\mathrm{T}}\mathbf{H}_{z}^{\mathrm{R}} + \mathbf{H}_{y}^{\mathrm{R}}\mathbf{H}_{z}^{\mathrm{T}}\right)}_{\text{hyz}} + M_{33} \underbrace{\left(\mathbf{H}_{z}^{\mathrm{T}}\mathbf{H}_{z}^{\mathrm{R}}\right)}_{\text{hzz}}$$

$$(2)$$

= $[\mathbf{hxx} \ \mathbf{hxy} \ \mathbf{hxz} \ \mathbf{hyy} \ \mathbf{hyz} \ \mathbf{hzz}] \cdot [M_{11} \ M_{12} \ M_{13} \ M_{22} \ M_{23} \ M_{33}]^{1}$.

Given the data vector **D** to consist of *m* measured data containing in-phase and quadrature components at a given frequency, we can define the unknown parameter vector to be the magnetic polarizability tensor **M** comprising *n* real and imaginary coefficients (n = 12 and corresponds to six complex elements) as illustrated by Eq. (2) and the dipole model as $\mathbf{F}(\cdot)$: $\mathbf{M} \to \mathbf{Z}$ used to calculate the forward data. In this chapter, we employ the damped Gauss Newton method to solve for **M** by minimizing the Euclidean L2 norm written in the form

$$E(\mathbf{M}) = \frac{1}{2} \|\mathbf{F}(\mathbf{M}) - \mathbf{D}\|^2.$$
 (3)

The method is an iterative procedure that generates a series of tensor updates $\{\mathbf{M}_k | k = 1, 2, 3, ...\}$ which desirably should converge towards a global solution \mathbf{M}^* that best minimizes [Eq. (3)]. The tensor is updated using the recurrence relation given by

$$\Delta \mathbf{M} = \left(\mathbf{J}_{\mathbf{M}_{k}}^{\mathrm{T}} \mathbf{J}_{\mathbf{M}_{k}} + \lambda_{k} \mathbf{R}_{k}\right)^{-1} \mathbf{J}_{\mathbf{M}_{k}}^{\mathrm{T}} (\mathbf{D} - \mathbf{F}(\mathbf{M}_{k})),$$
(4)

$$\mathbf{M}_{k+1} = \mathbf{M}_k + \Delta \mathbf{M},\tag{5}$$

where \mathbf{M}_{k+1} and \mathbf{M}_k are the new and the old updates of the tensor. $\mathbf{R}_k \in \mathbb{R}^{n \times n}$ is the regularization matrix used to stabilize the inversion and chosen to be a diagonal matrix with $[\mathbf{R}_k]_{ii} = \text{diag}(\mathbf{J}_{\mathbf{M}_k}^T \mathbf{J}_{\mathbf{M}_k})$, and λ_k is the regularization parameter which controls the weight of regularization on the solution. This parameter is adapted in the iterative process using a damping mechanism similar to the Levenberg–Marquardt method to achieve optimum reduction to Eq. (3). $\mathbf{J}_{\mathbf{M}} \in \mathbb{R}^{m \times n}$ is the Jacobian matrix defined as the partial derivative of the model **F** with respect to **M**:

$$\mathbf{J}_{\mathbf{M}} = \frac{\partial \mathbf{F}}{\partial \mathbf{M}}.$$
 (6)

The Jacobian matrix can be directly defined by referring to the formula in Eq. (2):

$$\mathbf{J}_{\mathbf{M}} = \begin{bmatrix} hxx_1 & hxy_1 & hxz_1 & hyy_1 & hyz_1 & hzz_1 \\ hxx_2 & hxy_2 & hxz_2 & hyy_2 & hyz_2 & hzz_2 \\ \vdots & \vdots & \vdots & \vdots & \vdots & \vdots \\ hxx_m & hxy_m & hxz_m & hyy_m & hyz_m & hzz_m \end{bmatrix}.$$
 (7)

The same procedure described above can be applied to evaluate the location of the target. In this case, the unknown parameter vector is $\mathbf{r} = [\mathbf{r}_x, \mathbf{r}_y, \mathbf{r}_z]$ relative the coordinate system of the sensor. In order to evaluate the corresponding Jacobian,

the partial derivative of the dipole model with respect to \mathbf{r} is calculated using the perturbation method. Using this technique $\mathbf{J}_{\mathbf{r}}$ can be calculated as

$$\mathbf{J}_{\mathbf{r}} = \frac{\partial \mathbf{F}}{\partial \mathbf{r}} = \frac{\mathbf{F}(\mathbf{r} + \Delta \mathbf{r}) - \mathbf{F}(\mathbf{r})}{\Delta \mathbf{r}}.$$
(8)

3 Estimating Target's Magnetic Polarizability Tensor and Location

3.1 Trial Axial Helmholtz Coil System

The Helmholtz coil is a sensor configuration that produces a nearly uniform field in the region between a pair of identical coils that are coaxial and excited in the same direction. If a target is projected sequentially with a uniform field by three Helmholtz coils placed so as their axes are mutually orthogonal, and the target response is evaluated between the different coil combinations, then we will accumulate six independent complex measurements producing an exactly determined system to solve for the polarizability tensor of that particular target. This so-called triaxial Helmholtz coil system has been simulated in COMSOL (a finite element software package) as depicted in Fig. 1. The coils were modeled as squares and placed at distance larger enough than the target size to enhance the homogeneity of the field across the target. Usually, there are a few types of land mines to search for in a given exploration field, in which case this coil arrangement can be efficient to build a library of tensors for the known metal components in such land mines. Identification procedure can then be performed by comparing the inverted tensors acquired from a test-field metal detector to the set of library tensors.



Fig. 1 Simulated FEM setup for (a) triaxial Helmholtz coil and (b) EMI planar sensor above type 72 analogue target that is buried in a hemispherical ground



Fig. 2 Simulated targets: (a) sphere, (b) Elsie mine ring, and (c) type 72 analogue

3.2 Test Targets

For testing we have considered three targets of different size and shape as shown in Fig. 2. To start with, a sphere was chosen to verify the accuracy of tensor computation by the triaxial Helmholtz coil since it is a simple shape for which analytical solution of the inductive EM response is available. A second target is a model of aluminum detection ring from an "Elsie" AP mine, and the last test piece model is an aluminum analogue of the detonator component inside a low-metal type 72 AP mine, which is considered to be difficult to detect.

3.3 Electromagnetic Induction Metal Detector

We have tested the inversion algorithm on simulated EM data obtained from the planar EMI sensor shown in Fig. 1b. It consists of an excitation coil and two differential receivers arranged in two layers with crossovers normal to each other. This gradiometer configuration is not untypical of the type of coil geometry in some commercial devices. In order to determine the type of scan that will provide the desired multi-axis polarization of the target, we have exploited the pattern of the magnetic field distribution generated by the sensor. Right beneath the sensor, the field is straight down, whereas to the edges, the field tends to be more horizontal. Thus, a raster scan consisting of sweeping the target in horizontal lines parallel to each other is a suitable scanning approach. A 17×17 sample grid with 2 cm spacing has been used to collect 289×2 complex data points above the target. For simulations, the sphere and the aluminum ring have been placed at 10 and 20 cm below the sensor, respectively, and the inversion algorithm was used to estimate their polarizability tensor, assuming their location was known. For the last example, the type 72 analogue target was buried in a hemispherical ground with a lightly magnetic soil ($\mu_r = 0.01$ H) at 10 cm depth, and its tensor was determined, assuming its location on the transversal plane is misestimated by 2 cm in the x and y directions, respectively.

4 Results and Discussion

Results of polarizability tensors evaluated using the triaxial Helmholtz coil system are shown in Table 1 in the top row of each target. The sphere polarizability is almost diagonal with homogenous coefficients. For the same frequency, material, and size, the analytical formulation yields an isotropic tensor with coefficient $\alpha = (-2.92 - 0.28i) \times 10^{-07}$. The relative error is virtually zero indicating this uniform field forming coil is an efficient numerical tool for tensor calculation. The calculated tensors for the Elsie mine ring and the type 72 analogue are approximately diagonal with two coefficients more or less equal to each other in the *x* and *y* directions. This conforms to the geometrical nature of these targets exhibiting bodies of revolution with the Cartesian *z* axis as the axis of rotation. Other than geometrical features of these targets, the real and imaginary of dominant tensor elements are of the same sign which correlates to conductive nonmagnetic targets.

The second row for each target shows the tensors obtained by inverting the spatial EMI data collected above the target using the damped Gauss Newton. For all targets the solutions converged quickly in no more than 10 iterations. The relative mismatch RMS error was calculated between these tensors and those obtained using the Helmholtz system. The aluminum ring produced the largest error 31 %, and this may be attributed to its furthest distance away from the sensor, whereas the sphere at short distance 5 cm yielded the smallest error.

As an example illustration, Fig. 3a shows the inductive response on the complex plane from the type 72 analogue for all sampled data. This response exhibits a constant phase shift as expected and is symmetrical about the origin reflecting the differential nature of the receivers. A plot of the optimization of the corresponding tensor and location is shown in Fig. 3b, where the initial distance of the target relative to the true location has reduced after 5 iterations to about 1 cm. In parallel, the tensor converged to within 12 % deviation from the actual tensor.

	M_{11}	M_{12}	M_{13}	M ₂₂	<i>M</i> ₂₃	M ₃₃	Error
Sphere at	-2.92e-07	2.29e-11	4.16e-11	-2.92e-07	-1.70e-12	-2.92-07	5 %
1.58 kHz	-2.79e-07i	-2.42e-13i	-5.86e-11i	-2.79e-07i	+3.21e-13i	-2.79e-07i	
	-2.95e-07	9.05e-09	-1.57e-09	-2.70e-07	5.09e-09	-2.95e-07	
	-2.82e-07i	9.21e-09i	-2.20e-09i	-2.60e-07i	+5.92e-09i	-2.81e-07i	
Al ring at	-7.05e-07	-2.55e-09	-3.72e-09	-7.04e-07	-8.00e-10	-2.13e-05	31 %
1.58 kHz	-2.28e-06i	+1.99e-10i	+6.75e-11i	-2.28e-06i	-1.03e-10i	-2.60e-05i	
	-6.11e-06i	5.28e-07	4.46e-07	-2.90e-06	8.14e-07	-2.44e-05	
	-8.96e-06i	9.06e-07i	+5.61e-07	-5.04e-06	9.25e-07i	-2.94e-05i	
Al capsule at	-1.48e-07	9.31e-11	1.18e-10	-1.48e-07	-1.55e-11	-1.55e-07	12 %
6.3 kHz	-1.69e-07i	+3.46e-11i	-2.03e-11i	+1.69e-07i	-8.70e-12i	-1.57e-07i	
	-1.71e-07	-4.72e-09	1.48e-08	-1.39e-07	2.52e-08	-1.56e-07	
	-1.85e-07i	+2.95e-10i	+1.64e-08	-1.58e-07	2.47e-08	-1.53e-07i	

Table 1 Inverted polarizability tensors from Helmholtz and EMI planar sensor



Fig. 3 (a) Inductive response on the impedance plane and (b) reconstructed tensor and location for the 72 APM capsule

5 Conclusion

The chapter presented a procedure for evaluating the magnetic polarizability tensor of buried metallic targets. The Helmholtz coil proved to be an efficient tool for numerically evaluating the tensor of arbitrary-shaped targets with known geometrical and material properties. The damped Gauss Newton method was able to successfully invert spatial EMI data above target to yield an estimation of its magnetic polarizability tensor and a correction of its location. Properties of inverted tensor in relation of target geometrical and material properties have been discussed. Future work will focus on tensor spectroscopy for target identification.

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UWB Short-Pulse Radar: Combining Trilateration and Back Projection for Through-the-Wall Radar Imaging

O.B. Daho, J. Khamlichi, M. Ménard, and A. Gaugue

Abstract In this chapter, we propose a novel way to combine back projection and trilateration algorithms for through-the-wall imaging using an ultra-wideband (UWB) short-pulse radar system. The combination of the two algorithms increases the detection-localization performance. To accomplish this improvement, the multi-target localization problem of trilateration is addressed by the calculation of the root-mean-square error with regard to the estimated position and those of all possible target positions. The radar system's entire processing pipeline is described, with a focus on the imaging block. The data were acquired using a multistatic radar system with a 3.2 GHz bandwidth. Simulations and experiments indicate that our combined method outperforms other methods. Simulation and experimental results are shown, compared, and discussed.

Keywords UWB pulse radar • Through-the-wall radar imaging • Back projection • Trilateration • CFAR

1 Introduction

Through-the-wall surveillance (TWS) has become a strategic research topic because of its outstanding applications in major fields, such as antiterrorism and urban security. TWS radar systems are used as a tool to aid decision making when establishing the best intervention strategies that maximize the effectiveness of the intervention while minimizing human casualties. UWB radar systems are commonly used to detect, locate, and track human beings behind dielectric opaque obstacles (walls) such as concrete, cinder block, or wood. In recent years, several

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products have been developed like Prism200 [1], Xaver800 [2], and Radar Vision [3]. Nowadays, these systems are used by task forces in real field operations.

For any TWS system, the imaging module is the most important component in its processing pipeline. Hence, different methods have been developed. Back projection and its different variants [4], widely used because of their simplicity, can provide a map of the scene under surveillance. The intensity of each pixel of a generated image is the result of combining the energy of all the received signals. Therefore, back projection is well adapted to multiple-target detection and gives good results for low signal-to-noise ratios (SNR). Nevertheless, signals have to be well calibrated; otherwise targets will be missed even if their echoes are detected. This problem does not arise in the case of trilateration [5], which is used in several systems of different applications. Trilateration determines the position of the target by computing the point of intersection of N spheres whose centers are the receivers and the rays are the target-receiver distances. However, trilateration algorithm is not suited to multiple-target detection and performs badly at low SNR values. Thus, the two algorithms have a degree of complementariness. Consequently, their combination was investigated in order to develop a radar imaging method with high detection-localization performance. However, even when endowed with an effective imaging method, the radar system has to be supplemented by other functionalities. The latters are aimed at extracting more relevant information, e.g., the characteristic pattern of a target and its trajectory. For that purpose, several post-imaging techniques were investigated such as image enhancement, constant false alarm rate (CFAR) detection [6], and multiple-target tracking. The adoption of a CFAR algorithm in short-range imaging is judicious. Instead of using a single detection threshold for the whole image, the use of CFAR can validate a distant target with a low intensity signature in order to extract it from the noisy background. This capability is possible because the data are observed through a window of analysis to determine the local threshold of detection, which depends on the positioning with reference to the radar system. The radar system under discussion does not use the CFRAR processor in this way. Its 2D version is used to process the produced radar images directly. In this work, we propose a combined back projection-trilateration radar imaging method. We demonstrate how the multitarget issue of trilateration can be addressed and show how the combination of the two algorithms increases the detection rate. At this stage, the combined method has a better performance at the price of a slightly higher false-positive (FP) detection rate, which is reduced using a modified CFAR algorithm.

The remainder of the chapter is organized as follows. A succinct description of our radar prototype is given in Sect. 2. CFAR algorithm is presented in Sect. 3. In Sect. 4, we show how to combine the back projection and trilateration algorithms. Section 5 presents simulation and experimental results. A summary of the results and some extensions to the current work are discussed in Sect. 6.

2 Description of the Radar Processing Pipeline

Our multistatic radar employs three UWB receivers with an aperture of $\pm 45^{\circ}$ and an UWB omnidirectional antenna as a transmitter. The transmitted signal is modeled by an amplified Gaussian pulse that modulates a sinusoidal function. The center frequency is 4.6 GHz with a bandwidth of 3.2 GHz.

A block diagram of the processing steps [7] is shown in Fig. 1. The acquired signals are preprocessed using the usual techniques. Envelope detection with the Hilbert transform is used for demodulation. Time zero is determined knowing the positions of the receiving antennas and the time of arrival of the corresponding cross talk. The latter is eliminated using a priori cross talk information. Then, the signals are de-noised to enhance their SNR. The radar imaging algorithm is performed via a combined back projection-trilateration algorithm. A modified CFAR detection algorithm is also used in order to decrease the FP detection rate. Afterward, the obtained time series of the radar image is split into two parts. The first part contains the moving targets, and the second part contains the stationary ones. This partition is done using a moving object segmentation algorithm. This operation aids the application of specific processing to each part. Multiple-target multi-hypotheses tracking is applied to the first part. The highlighting of interior walls using a Radon transform-based method is applied to the second part. The final result is the display of several images showing different kinds of information: notably moving targets, steady targets, and the architectural plan of the scene. In the following section, we focus on the radar imaging method and how to enhance its detection performance using a CFAR detection algorithm.



Fig. 1 Block diagram of the radar processing pipeline

3 CFAR Algorithm

Traditionally, the images are scanned with a sliding 2D CFAR window. The scan is done pixel by pixel throughout the entire image, but the edges undergo a special processing. In Fig. 2, we show the three different cells as presented in Sect. 1. The Cell Under Test (CUT) is the inner (red) part. It should cover the target signature. The Reference Cell (RC) is the outer (green) part. It should contain pixels from the background. The Guard Cell (GC) is the third (yellow) part. It separates the two first parts and prevents the use of pixels with high intensity belonging to the target's sidelobes [8]. The sizes of these cells depend heavily on the radar system features such as bandwidth, angular resolution, and the signal processing performed. For a given image I, the sum of the power of the CUT (μ_{CUT}) and the RC (μ_{RC}) are first computed. Then, these two powers are compared. The inner cell to outer cell power ratio is defined in Eq. (1). According to the value of the ratio R, the center of the CUT will either be considered or not as a target. Therefore, RC energy should not be close to zero in order to avoid increasing FP detections and to prevent division by zero problems. Equation (2) defines the CFAR test which produces a binary image of detection φ .

$$R = \frac{\mu_{\text{CUT}}}{\mu_{\text{RC}}} = \frac{\sum_{(i,j)\in\text{CUT}} \left[I(i,j)\right]^2}{\sum_{(m,n)\in\text{RC}} \left[I(m,n)\right]^2},\tag{1}$$

$$\varphi(x,y) = \begin{cases} 1 & R > T \\ 0 & \text{else} \end{cases},$$
(2)

where, T is a suitable threshold chosen according to the image's characteristics.

The traditional CFAR detection was implemented. However, for real-time and detection performance considerations, several significant modifications were made to the CFAR window and the CFAR test to improve performance. The modified CFAR algorithm outperforms the traditional one.



Fig. 2 Traditional CFAR sliding windows. (a) 1D CFAR window, (b) 2D CFAR window

4 The Combined Radar Imaging Method

Our imaging module is mainly based on time-domain back projection and trilateration algorithms, which are well adapted to our pulse radar system. The use of only one imaging method is not obligatory. Indeed, the imaging methods can be combined to outperform their classical versions. Thus, back projection method can be combined with other techniques, e.g., trilateration method, to get more accurate localization or just to confirm the information given by back projection. Such a combination can be observed while operating Radar Vision [3]. To combine back projection and trilateration algorithms, the latter is modified to be adapted to multiple-target localization. To address the multi-target localization issue of the trilateration, an association rule is established. Figure 3 shows two cases of association. In the case of a good combination (a), the points of intersection between the ellipses are close to the estimated location of the target. Otherwise, the target is at least far from one ellipse (b). The quantification of this distance is done through the root-mean-squared error of the position, expressed by Eq. (3). The association factor $Q_{\text{association}}$ measures the gap between estimated position of the targets and the ellipses of detection. The greater the gap, the higher its $Q_{\text{association}}$ value, because the likelihood that the considered echoes originate from a real target is low. Thus, after detection of the target echoes in the received signals, all possible locations are computed using the traditional trilateration algorithm [6]. Among these locations, the false are eliminated according to their association factor value. The choice of an appropriate threshold leads to good localization performance.

$$Q_{\text{associatin}} = \sqrt{\frac{1}{N} \sum_{i=1}^{N} \left[c \cdot \text{ToA}_i - \left(d(\text{Tx}, P) + d(P, \text{Rx}_i) \right) \right]^2}$$
(3)



Fig. 3 Echoes association of the multiple-target trilateration. (a) Good association, (b) bad association

where *N* is the number of receiving antennas, *c* is the speed of light, ToA_i is the time of arrival of the investigated target on the *i*th receiver, d(Tx,P) is the transmitterpixel distance, and $d(P,\text{Rx}_i)$ is the pixel—*i*th receiver distance.

As mentioned before, the two algorithms seem to be complementary, which is why we considered combining them. If one naively summed the results of back projection and trilateration (blind combination), then the detection performance would not be good. First, the combination is done by matching the localization results of trilateration method with those of CFAR detection on the back-projected image. Thereafter, isolated targets detected only by back projection are added depending on their clutter strength. The ones detected only by trilateration are added depending on the value of $Q_{\text{association}}$. The false detections are then suppressed using image processing based on the fast hybrid geodesic reconstruction algorithm described in [9]. At this stage, the results are in the form of two layers. One layer contains back-projected targets and the other contains the trilaterated ones. Despite the fact that the two layers comprise all the targets detected, one layer does not contain all of them. The two layers are just complementary. In order to adapt the imaging bloc results to the downstream processing steps, each layer must be completed relative to the other in two steps. The coordinates of the centers of gravity of the isolated targets in the back-projected layer are added to the trilaterated layer. The received signals are amplified by the times corresponding to the positions of the isolated targets in the trilaterated layer. These parts of the signals are then back projected, and the results are added to the back-projected layer. Thus, the results are adapted to the processing pipeline since the backprojected layer is used by the segmentation bloc and the trilaterated layer is used by the tracking bloc.

5 Experiments and Results

In this section, simulation and experimental results are shown, compared, and discussed. We have developed multistatic pulse radar system. This prototype employs a laptop on which the entire processing pipeline is loaded. With the aim of assessing the method proposed, we created many complex scenarios; however, we limit ourselves to one real scenario and one simulated scenario in this chapter. Both sets of data were processed with the algorithms described above. The results of combining back projection and trilateration, with and without incorporating a CFAR algorithm, are now discussed. The methods are evaluated according to the probability of detection (PD) and the average number of FP detections at different SNR values. In the following account, the unit of the coordinates is a meter, and the transmitter is the origin of the coordinate system. The scenarios contain fixed and moving targets in order to make close to real experiments. For the simulated experiments, one target moved from point (-2, 1.5) to point (2, 4), and the other traveled from point (2, 1) to point (-2, 4.5). These two targets had the same



Fig. 4 The used scenarios for evaluating the proposed method. (a) simulated scenario : 3 targets, (b) real scenario : 2 targets

 Table 1
 Detection performance of the presented methods (101 frames, simulations) (two moving targets + one immobile target as a back wall)

_	SNR (dB)	0	3	10	15
Back projection	PD ^a (%)	79.54	83.17	84.48	87.13
	ANFA ^b	1.45	0.81	0.72	0.62
Multi-target trilateration	PD (%)	47.85	57.09	73.27	76.24
	ANFA	0.63	0.26	0.32	0.25
Blind combination	PD (%)	85.15	88.45	93.73	95.38
	ANFA	2.09	1.07	1.04	0.87
Combination without CFAR	PD (%)	71.95	81.52	90.76	94.72
	ANFA	1.15	0.71	0.94	0.79
Combination using CFAR	PD (%)	71.95	81.52	90.76	94.72
	ANFA	0.67	0.46	0.34	0.37

^a*PD* probability of detection

^bANFA average number of false detections

The bold values in the table represent the best joint compromise between the PD and the ANFA

reflectivity coefficient. A third target remained immobile at (0, 5) as a back wall. For real experiments, the scene under surveillance was a (4×3.5) rectangular room where a metallic target was moved according to a zigzag trajectory. For both cases, targets are detected through a 7 cm plasterboard wall.

Table 1 (respectively, Fig. 5) shows the detection rates of the methods at different noise levels in a simulated (respectively, reel experiments) high-complexity scenario. The ANFA is the average number of FP detected targets over the number of the detection frames. The results depend heavily on the number of the targets and how they move in the scene.

Naively summing (blind combination) the results of the two methods does not give good results due to the high rate of false detection. To ameliorate this problem, the modified CFAR detection algorithm was incorporated into the combined imaging method. It helped to enhance the detection performance further by reducing the



Fig. 5 Graph representation of the proposed method performance (31 frames, real experiments) (one moving target + back wall). (a) Average number of false detection, (b) probability of detection

FP detection rate. Our method demonstrated the best joint compromise of increased PD and reduced FP detections, especially at low SNR and in the case of complex multiple-target scenarios where it is of most benefit. Furthermore, the imaging bloc has the two imaging/localization layers as output. The back-projected layer is used for the extraction of moving targets from the background. The trilaterated layer is used for moving target tracking. Consequently, the imaging bloc is well adapted to the processing pipeline.

6 Conclusion and Future Works

The multiple-target issue of the trilateration method was addressed, in order to combine it with back projection. The combined imaging method has been presented. The results show that the combination of these two methods eliminates their limitations and yields an accurate imaging method with regard to the false-negative detection rate. However, the values obtained for false detection rate using this method do not fit with our aim of detection effectiveness. Therefore, a CFAR detection algorithm was incorporated into the imaging block to decrease the false detections. The presented imaging method outperforms both back projection and trilateration with regard to false-positive and false-negative detection. The quality of images produced by this method increases the effectiveness of the subsequent processing steps, especially for the low SNR values.

In future work, different types of walls will be tested. Also, information obtained from the subsequent processing steps, e.g., the segmentation of moving targets and the multiple-target tracking, will be used to improve further the imaging method presented and the adaptation of the whole of processing pipeline.

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Toward Integrated µNetwork Analyzer

M. Kmec, M. Helbig, R. Herrmann, P. Rauschenbach, J. Sachs, and K. Schilling

Abstract The article deals with recent development steps toward monolithically integrated micro-Network Analyzer (µNA). The device will deploy M-Sequencebased single-chip transceivers with a built-in ultra-wideband wave separation unit in the receiver chains. The introduced on-chip wideband wave separation is realized using an optimized resistive directional coupler combined with a customized differential LNA as detector. The wave separation works almost down to DC, and its upper frequency limit is determined by the performance of the implemented technology (i.e., bridge resistors, transistors, etc.), the selected circuit topology, and the wirings of particular coupler components but also by the IC packaging itself. Even though the upper limit is designed to be compatible with the analog input bandwidth of the receiver circuit [which is about 18 GHz for naked die (Kmec et al., M-Sequence based single chip UWB-radar sensor. ANTEM/AMEREM 2010 Conference, Ottawa, 2010)], the packaged IC is intended for use up to 8 GHz. Finally, the discussed transceiver is a further development of the mother SiGe System-on-Chip (SoC) presented in the work cited above.

Keywords Integrated UWB sensor • M-Sequence • Integrated directional coupler • System-on-Chip (SoC)

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1 Introduction

It is undisputable that the last decades in electronics are stamped with exceeding innovative achievements resulting in rapid performance upgrowth. For example, the first silicon transistor was produced in 1954 by Texas Instruments [1] and several decades later, the physical gate length of elementary MOS device reaches only several tens of nanometer or the enhanced vertical compositions of HBT devices allows application in nearly THz range [2, 3]. Thus, the miniaturization is commonly considered as a key factor to performance increase in the electronics. Nevertheless, the achievements in the field of instrumentation and measurements plaved and still play also an important role in this evolution. As the applications, devices, and systems have penetrated the RF and microwave ranges, the Network Analyzers became standard measurement equipment for measurement of scattering parameters (S-parameters). Now, it is about 50 years ago as the first Network Analyzers started to be introduced. The earliest devices of this type were the Z-g-Diagraph from Rohde and Schwarz and the Elliot automatic swept-frequency impedance meter (A.S.F.I.M) developed by Elliott Brothers Ltd. The Z-g-Diagraph was the first instrument to directly indicate complex S-parameters without the need for additional calculation. The A.S.F.I.M has measured either the reflection or transmission properties of a device or component under test, as summarized in [4] or [5].

Today's Network Analyzers became very complex systems which offer a large S-parameter measurement flexibility as well as a number of added capabilities such as for harmonics, modulations, mixer, and noise measurements. Despite this multitasking capability, in many applications, as, for example, specific microwave sensing tasks [6] or certain device and component tests in production chains, the measurement objectives are simply straightforward or predefined and fixed. In order to solve such kind of tasks, one can partially abstain from the large flexibility of the classical analyzer and use simpler, smaller, and cheaper devices tailored for one or another implementation. Especially here, the equipment using pseudo-noise (pn) sensing approach which utilizes a unique combination of high-frequency analogue electronics and digitally realized correlation in one device offers an interesting alternative to the classical approaches and opens new instrumentation and measurement possibilities as extensively discussed in [7] and references therein. The chapter concentrates on the recent steps toward integrated micro-Network Analyzer (μ NVA) which profit from this unique approach. The first section describes briefly the latest multichip-based M-Sequence sensing devices whose primary topology is considered as paradigm for the work toward *pn*-based µNVA. The adjacent main section deals with S-parameter test set specifics. A single-chip head with integrated wave separation is discussed in the final section.

2 Recent General-Purpose M-Sequence-Based Devices

So far there are various achievements that show the practicability of pn sensing devices which use M-Sequences as stimuli for short-range applications, as described, for instance, in [6–11].

Recent standard devices of such kind (example is depicted in Fig. 1) include commonly one transmitter and two receiver ultra-wideband channels as well as complex electronics and software engine for device control, preprocessing, and data transfer and storage over USB (or LAN). Moreover, the modular structure of the sensor heads provides many options to individually adapt the device operation to application-specific requirements. So, for example, the standard sensing devices can be cascaded to complex MiMo-sensor arrays or equipped with special frontends which lead to new measurement concepts [7, 9, 11].

3 Network Analysis Using the M-Sequence Approach

As dimensionless expressions of the behavior of objects under test, scattering parameters turn out to be particularly convenient in many electric network, radar, or material investigations. If we consider a generic network with N ports, the S-parameter matrix $(N \times N)$ can be expressed through the b = Sa relationship, where b and a are column vectors of the length N which comprise a set of waves $(a_i \text{ and } b_i)$ incident on and emanated from the *i*th port of the network [12]. For conveniently used reference, impedance Z_0 can be particular elements of a and b symbolically rewritten in the following forms:

$$a_i = \frac{V_{\text{incident},i}}{\sqrt{Z_0}} \quad \text{and} \quad b_i = \frac{V_{\text{emanated},i}}{\sqrt{Z_0}}.$$
 (1)

and a	std. operation band	DC up to 8GHz	
	measurements speed	up to 70 000 IRF/s	
A	specific dynamic range ¹	114 dBs	
Manna mar	equiv. impulse widths	down to 77ps	
an areas	time samples per IRF	511 or 4095	
ALAN TRANSPORT	unambiguity ranges	from 4 up to 123 m	
	easy configurable for ECC and FCC conform apps		

¹Specific dynamic range given here represents a dynamic range of impulse compressed receiving data for a recording time of one second

Fig. 1 Portable general-purpose M-Sequence-based sensing device with one transmitter and two receivers and its selected parameters—for parameter description, see [7]



Fig. 2 Conceptual equipment for measuring of reflection coefficient S_{11} using of a standard M-Sequence sensing unit—generalized principle of M-Sequence-based reflectometer

Thus, in a given environment Z_0 , the voltage waves $\underline{V}_{\text{incident},i}$ and $\underline{V}_{\text{emanated},i}$ are essential quantities to calculate the scattering parameters. In order to measure these waves, let us consider the equipment which utilizes the M-Sequence-based approach—i.e., a standard M-Sequence sensing device.

For the sake of simplicity, we initially regard measurement of the reflection coefficient of a 1-port test medium (i.e., S_{11} -parameter) given by b_1/a_1 , as illustrated in Fig. 2.

The conceptual reflectometer depicted in Fig. 2 is realized by connecting the receivers #1 and #2 to two wave separators, i.e., directional couplers. This arrangement provides the receivers #1 and #2 with information about the waves (ideally) incident on and reflected from the device under test (DUT). While the receiver's outputs of the standard M-Sequence devices are proportional to the received voltage, the measured reflection coefficient is proportional to $V_{\text{reflected},1}/V_{\text{incident},1}$. However, the imperfections in the wave separation block and reflected from the receiver's directivity of the couplers C1 und C2, so that a fraction of the reflected wave leaks to the receiver #1 and a portion of the incident wave transpires into the receiver #2. The directivity is therefore one of the most important parameters in the wave separation block.

Note, in order to measure more usable, full *S*-parameter set of generic two-port devices under test (i.e., set of S_{11} , S_{21} , S_{12} , S_{22}), one way is to deploy two M-Sequence-based reflectometers working synchronously, each connected to a different port of the tested device and with alternately activated stimuli. Finally, such cascading of several basic units leads to an *N*-port Network Analyzer without the need of RF switches.

One of the most interesting directional couplers for monolithic integration employs a customized Wheatstone bridge topology as illustrated in Fig. 3.

The illustration in Fig. 3 displays a redrawn variant of the bridge for use in configuration with the M-Sequence device receiver topology (see Fig. 4) [13]. The circuit uses an optimized differential amplifier as detector and interface buffer,



Fig. 3 Directional coupler employing a Wheatstone bridge with a differential amplifier as detector



Fig. 4 Conceptual diagram and IC microphotograph of discussed coupler connected with an ultrawideband sampling circuit

whereupon this amplifier is kept as close as possible to the bridge to avoid long transmission lines. In Fig. 3, the dimensioning formulas for particular bridge elements derived from the balancing condition when the bridge is loaded with the Z_0 are also depicted. Since the factor α can be arbitrary selected, there are infinity different layouts which differ from each other only by transmission factors (S_{31}, S_{32} , etc.) as shown, for instance, in [7] or [14]. This is an interesting feature of the circuit even for a case of monolithic integration—it allows to precisely trim the bridge characteristics. Moreover, the use of miniature resistors on chip, carefully shaped interconnections, and exact $Z_{in} \sim Z_0$ dimensioning of detector amplifier over the entire intended frequency band easily enables high bandwidth with excellent directivity.

Figure 4 shows the conceptual diagram and IC microphotograph of the discussed coupler with an ultra-wideband sampling circuit, which is an essential part of the receivers in UWB M-Sequence-based sensing systems. The circuit was realized using IHP's high-performance process with peak f_T of 110 GHz ([3] and IHP technology reference therein). The bridge components (Z_1 , Z_2 , Z_3) were built



Fig. 5 Measured performance characteristics of the sampling circuit with the directional coupler at the input (*left*); close-up photograph of the packaged IC and test board example (*right*)

using polysilicon resistors. Note, the "poly" resistors work very well with respect of high-frequency performance since they are separated from the silicon substrate by the field oxide, leading to small parasitic shunt capacitance. Despite the fact that the bridge can easily operate within the frequency range from DC up to far beyond 20 GHz, our currently implemented active circuits are optimized for frequencies up to about 18 GHz [13] or even less (see specifications in Fig. 1). Thus, the performance characteristic measurements of the final IC were performed over the frequency range which is consistent with the standard device operation band (up to 8 GHz) as depicted in the Fig. 5.

Keeping in mind that only housed chips can be used, the characterization was done with both naked die using a wafer prober (solid lines) and with the IC housed in standard plastic QFN (quad-flat no-leads) package and mounted on a test carrier (dashed lines). The test carrier and close-up photograph of packaged chips from Fig. 4 are shown in Fig. 5 on the right side. Examining the curves in Fig. 5, we can roughly conclude that in the case of a naked die, the isolation is better than -23 dB and the test port mismatch is better than -19 dB over the entire band. The insertion loss is about 16 dB as desired due to the restricted LNA 1 dB compression point ($P_{1dB_{IN}}$, see Fig. 4) to avoid amplifier saturation when open or short circuit is connected to the test port. The coupling factor over the complete signal chain (i.e., from the "test port" to the "output"—see Fig. 4) is positive due to the active nature of the circuitry, and it amounts about 9 dB over the band of interest. Unlike the naked die, with the housed IC, the influence of non-optimized packaging becomes prominent as expected at higher frequencies. Nevertheless, this impact is still



Fig. 6 High-frequency board with implemented integrated directional coupler (left) and portable UWB reflectometer deploying this board

acceptable, even though the test port mismatch slightly exceeds the 10 dB margin already at about 6.5 GHz.

Lastly, Fig. 6 depicts an experimental high-frequency board assembled with the discussed coupler in a receiver chain and device implementation example, i.e., ultra-wideband *pn*-based reflectometer intended for applications up to 8 GHz. The specific dynamic range of the device after the common open, short, match calibration is approximately 100 dB.

4 Single-Chip System Integration

As presented in [13], it is in some aspects beneficial to integrate all high-frequency functional blocks of the M-Sequence device into a single chip. Figure 7 shows the simplified block topology of realized M-Sequence SoC (System-on-Chip) transceiver with two integrated directional couplers. The implemented SoC configuration is thus consistent with the conceptual measurement equipment depicted in Fig. 2, whereas the directional couplers are embedded on the chip internally. The particular functional blocks of the SoC except the coupler unit have been extensively discussed in [13] or [7]. The directional bridges utilize the circuit topology depicted in Fig. 3. The parameters of the individual bridge components are trimmed by means of factor α to operate the receivers in linear region over the full span of the reflection coefficients.

The photograph of the related transceiver die and selected performance characteristics of the bridge-receiver chain measured on the die are shown in Fig. 8. In spite of the implementation of several high-swing digital and sensitive analog circuits that must be decoupled from each other, the SoC die features a compact size of $2 \times 1 \text{ mm}^2$ and moderate power consumption of around 1 W at -3 V supply.

Comparing the individual curves in Figs. 5 and 8, we can conclude that the directivity of the SoC bridge-receiver chain over the entire frequency band of interest is slightly lower in comparison to the previous bridge variant. This is due



Fig. 7 Implemented single-chip architecture with two embedded directional couplers



Fig. 8 Single-chip transceiver with integrated directional couplers and selected performance characteristics of bridge-receiver chain

to component parameter variations on chip and corresponds to the results of corner simulations for design functionality margins.

5 Conclusions

In this chapter we have shown our recent development steps toward μ NVA which opens new measurement possibilities. The small form factor of the final device will offer the ability to mount the measurement unit very close to the observed medium

or in a direct contact with the subject of interest (in situ measurement). This will avoid feeding cables including their complicacies and facilitates the handling of low or high impedance DUTs. The future work will focus on more desirable SoC packaging and system implementation as well as redesign and performance optimization of the directional bridges in the SoC transceiver.

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M-Sequence-Based Single-Chip UWB-Radar Sensor

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Abstract The article deals with a fully monolithically integrated single-chip M-sequence-based UWB-radar sensor, its architecture, selected design aspects and first measurement results performed on wafer and with packaged IC modules. The discussed chip is equipped with one transmitter and two receivers. The IC was designed and manufactured in commercially available high-performance 0.25 µm SiGe BiCMOS technology ($f_t = 110$ GHz). Due to the combination of fast digital and broadband analogue system blocks in one chip, special emphasis has been placed on the electrical isolation of these functional structures. The manufactured IC is enclosed in a low-cost QFN (quad flat-pack no-leads) package and mounted on a PCB permitting the creation of MIMO-sensor arrays by cascading a number of modules. In spite of its relatively high complexity, the sensor head features a compact design (chip size of $2 \times 1 \text{ mm}^2$, QFN package size $5 \times 5 \text{ mm}^2$) and moderate power consumption (below 1 W at -3 V supply). The assembled transceiver chip can handle signals in the frequency range from near DC up to 18 GHz. This leads to an impulse response (IRF) of FWHD ≈ 50 ps (full width at half duration).

Keywords UWB sensor • System-on-Chip (SoC)

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1 Introduction

Steady growing areas of UWB applications push the requests for new enhanced UWB sensor systems. A promising alternative for UWB sensor realisations is based on the so-called M-sequence approach—a concept which combines high-frequency electronics and associated signal capturing with appropriate data processing [1]. One of its most important benefits lies in the relatively low signal levels which have to be handled by the circuitry. When looking at the currently available semiconductor technologies and their roadmaps [2], the implementation in the SiGe BiCMOS branch seems to provide the best balance between performance and costs [3, 4]. Furthermore, it supports future development perspective for intended fully integrated M-sequence-based UWB sensor architectures due to its compatibility with the mainstream Si CMOS processing.

In recent time, advancements in the investigation of various UWB M-sequence device architectures have been achieved. They had been based on multi-chip structures equipped with separate customised ICs [5–7]. The gained experience, the need for UWB systems of increasing complexity (e.g. for MIMO UWB sensor arrays) and the cost considerations have motivated the first monolithic integration of the complete RF-part of the M-sequence UWB electronics into a single silicon die.

2 Single-Chip Architecture

Figure 1 shows a simplified block topology of the realised M-sequence SoC (System-on-Chip) and the periphery for signal digitisation and preprocessing.

According to the figure, the chip topology includes fully differential transmitter and receiver channels. Its architecture is based on the original approach presented in [1] covering the M-sequence generator and a subsampling receiver. But it was modified by further I/O ports and a second receiver channel in order to get the option to select the operational frequency band, to create MIMO-sensor arrays or to permit the device calibration. This consequently supports the utilisations of such integrated IC in a higher number of applications. For instance, the multichannel sensor topology is crucial for high-resolution imaging [7] or localisation [8], and the calibration capability is beneficial for stand-alone UWB sensors implemented, e.g. in material testing or characterisation systems.

2.1 SoC Stimulus Generator

The stimulus generator core consists of a chain of nine digital microcells (flip-flops) with two feedback taps (see Fig. 2). It realises the characteristic polynomial providing a periodical M-sequence of 511 elementary states—the so-called signal chips. This signal is transferred via output buffers to the periphery (see also [5]).



Fig. 1 Simplified architecture of the designed UWB M-sequence-based SoC



Fig. 2 SoC generator architecture with ninth-order shift register core and modulator

The generator core is designed to be driven with an arbitrary RF-clock within the frequency range $f_c \in \langle 0.5; 20 \rangle$ [GHz]. Despite the critical timing schema, it may be started or stopped by a simple TTL-signal originating from digital standard components (e.g. FPGA). Possible timing instabilities provoked by slow TTL-edges are compensated by the trigger control unit. As shown in Fig. 2, the generator is additionally equipped with a wideband modulator. It allows an optional increase of the overall bandwidth [5] and/or to shift the centre frequency of the operational frequency band [6, 9].

Figure 3 depicts two examples of different spectral allocations—baseband spectrum (left) and shifted spectrum (right). For illustration purposes, we used a clock rate of only $f_c = 1.6$ GHz and $f_{mod} = 6$ GHz so that the screen of the



Fig. 3 SoC transmitter output spectrum examples for $f_c = 1.6$ GHz and $f_{mod} = 6$ GHz

available spectrum analyser was able to give a complete overview of the signal spectrum.

The transmitter output buffers are trimmed for interfacing both 50 Ω singleended and 100 Ω differential ports. The 3 dB-bandwidth is 15 GHz for the baseband output and 20 GHz for the modulator output. The output power is 0 dBm each in single-ended mode. The SoC input buffers for f_c and f_{mod} represent active BALUNs to convert the external single-ended signals to the internally used differential signalling. Their bandwidth is 20 GHz. A level of 15 dBm for the clock and carrier signal is sufficient to operate the circuit.

2.2 The Synchronisation Unit

The synchronisation unit is responsible to provide a stable clock signal for the signal capturing and conversion. This involves a wideband T&H stage (internally routed), a commercial video ADC (externally connected) and preprocessing like synchronous averaging and pulse compression [1].

Basically, the synchronisation unit core consists of a nine-stage binary counter supported by input/output buffers and supplemental synchronisation flip-flops at the end of the divider chain (D_s , see Fig. 5) which removes countertiming ambiguities. Thus, the IC sampling clock synchronisation error with respect to f_c lies in the sub-picosecond region. In order to increase the flexibility of the system configuration [5, 6] without the obligation of circuit redesign, the synchronisation unit is



Fig. 4 Simplified schematic of the synchronisation unit architecture (ports legend corresponds to that in Fig. 2)



Fig. 5 Simplified schematic of the Rx architecture (ports legend corresponds to that of Fig. 2)

additionally equipped with an optional (switchable) shunt sampling clock path, which allows direct clocking supply from chip periphery, thereby enabling user selectable clock rates or enhanced signal-capturing schemes [5, 6] (Fig. 4).

2.3 The Ultra-wideband Receiver

The SoC receiver chains act as broadband sampling unit, which provide the measurement signals at a lower rate (typically 20–100 M samples/s) for subsequent digitalisation and data processing with commercially available low-cost components. The sampling unit works in the T&H mode (track and hold), whereas the actual sampling gate is embedded between a low-noise preamplifier (LNA) and an output buffer of high input impedance (Fig. 5).

The LNA is a common emitter differential amplifier with miller capacitance neutralisation. Its key parameters are 50/100 Ω single-ended/differential input impedance; 12 dB gain; noise figure better than 6 dB; and operational frequency band from 100 MHz to 18 GHz. The T&H stage is based on switched emitter followers which drive the charge storage capacitors. Thus, the T&H network represents a time variable capacitive load of the LNA leading to transient spikes while switching. This perturbs the measurement signal. These perturbations are

eliminated by an input interstage differential buffer which reduces the retroaction of the T&H onto the LNA by more than 40 dB.

Due the subsampling mode, the voltage droop of the charge storage capacitor during hold phase is an essential parameter of the T&H core, which calls for very low leakages in the switch, in the hold capacitor and mainly the input impedance of the output buffer. The achieved droop rate of the T&H circuit is better than 10 % FS/ms (FS—full scale). The output buffer represents a differential amplifier with emitter degeneration and a cutoff frequency of about 3 GHz. It serves as an ADC driver and provides a differential voltage swing of 1 V at the 1 dB compression point.

3 Single-Chip Design Philosophy

The chip architecture presented in Fig. 1 envisages very fast switching cells as well as sensitive analogue blocks integrated on the same chip substrate. Hence signal coupling and crosstalk will degrade the performance of the sensitive analogue parts of the receive circuitry, affecting finally the whole system. Therefore, special emphasis was placed on the isolation of the SoC subcomponents. Basically, this can be done by three ways: reduce the creation of the switching noise, reduce the noise susceptibility of critical circuits and avoid cross-coupling. Therefore, the first step in the SoC design was to identify the main noise aggressors and victims.

The strongest perturbations are provided by the flip-flops of the transmitter and synchronisation blocks which toggle at the RF-clock rate having a transition time of about 20 ps. The receivers are the victims. We have chosen the balanced current steering topology (based on differential amplifiers) as the best suited structure to reduce the generation of switching noise. Concerning the receivers, the benefit of a differential topology is to be seen in its ability of common mode suppression reducing the influence of the stray fields generated by fast switching cells in the neighbourhood of the analogue circuits. Additionally, the power supply rejection ratio (PSRR) is maximised in order to avoid perturbations by power supply impurities. The remaining noise susceptibility strongly depends on the path symmetry in the layout; thus very precise layouting is mandatory.

The usually applied method to reduce noise coupling propagation is a careful definition of the floor plan and supply power domains on the chip with respect to the architecture and later application needs. Figure 6 shows the floor planning adapted to our chip architecture.

The goal is to spatially separate the functional blocks based on their "sensitive/ destructive" order, i.e. their digital and analogue nature as well as the handled frequency and voltage level. The sources and victims should be separated as wide as possible. Additionally, each block is surrounded by a guard ring that completely encloses a given region as shown in Fig. 6. The guarding structure provides a low impedance contact from the defined supply voltage (e.g. GND) to the chip substrate. It is one of the most commonly used isolation schemes and seems to be



Fig. 6 Realised SoC floor plan to minimise on chip noise coupling adapted to M-sequence SoC



Fig. 7 (a) Microphotograph of the SiGe UWB SoC substrate with depicted functional blocks and isolation structures. Chip size is about $2 \times 1 \text{ mm}^2$. (b) Photograph of a packaged SoC mounted on the prototype test board. *Inset*: photograph of first SoC prototype in a QFN package

best suited for reducing crosstalk at high frequencies. This is due to voltage fixing of the substrate around the guarded area by absorbing the substrate potential fluctuations. It serves as a low impedance path to a "quiet" potential. Furthermore, the guarded regions should ideally have their own separated power supplies to avoid functional block interactions in the SoC.

4 Realised Chip and Packaged Test Prototype

The final SoC IC has been manufactured using IHP's 0.25 μ m SiGe BiCMOS technology SG25H3 [4]. We choose the high-performance bipolar section, which offers npn-HBTs with $f_t = 110$ GHz. The collector-emitter breakdown voltage of the transistors is BV_{CEO} = 2.3 V. The microphotograph of the manufactured chip with marked functional blocks is shown in Fig. 7a.

Although the internal speed in SoC is basically very high, RF signals exchanged with the external world are prone to problems due to the capacitance and inductance



Fig. 8 *Left*: SoC test assembly with the packaged module, $f_c = 9$ GHz. *Right*: Normalised system IRF with TX and RX1 connected via a DC block plus a 20 dB attenuator

of bonding wires and pads. To reduce the impact of these parasitics, octagon-shaped bond pads with 80 μ m diameter are implemented for high-frequency I/O ports. The rectangular pads are used for power supply, grounding and also for low-frequency control signals.

5 First Measurement Results

The chip characterisations were performed on two levels: on wafer and with the packaged module in order to evaluate the performance also under real conditions. For that purpose, some ICs were housed using low-cost 24 pin $5 \times 5 \text{ mm}^2$ QFN packages and mounted at a multilayer carrier made from Rogers 4003CTM laminate (Fig. 7b). The results of the on-wafer measurements were in good accordance with the expectations of the simulations. The SoC is able to operate at an arbitrary clock rate between 0.5 and 20 GHz.

The mutual decoupling between the different subcomponents was better than 60 dB within the whole frequency band. The packaged module was tested in a configuration covering the whole periphery as ADC and FPGA, as depicted in Fig. 8. The internal crosstalk degraded by about 20 dB due to QFN-housing which is however often less critical since in practical sensor application antenna coupling will usually determine the overall crosstalk. The excellent timing stability of the SoC allows removing this crosstalk from actual measurements by an appropriate calibration. This is demonstrated by the measured IRF in Fig. 8. The chip wiring slightly reduces the maximum clock rate to 18 GHz. The Tx power is about 0 dBm and the input related 1 dB compression point of the Rx circuit is 12 dBm.

6 Conclusions

Fully integrated, fully differential M-sequence radar SoC for short-range applications was presented. The realised IC shows very compact dimensions and good system performance such as broad frequency range of operation, reasonable on chip isolation, excellent timing stability, good linearity in the receivers and moderate power consumption in spite of its system complexity.

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UWB Antennas for CW Terahertz Imaging: Geometry Choice Criteria

I. Türer, A.F. Dégardin, and A.J. Kreisler

Abstract There is a strong need for wideband and sensitive receivers in the terahertz (THz) region. This chapter focuses on the detection of continuous waves (CW) in the THz, using innovative bolometric sensors working as mixers for heterodyne reception with down conversion to the GHz range, as needed for radio astronomy, remote sensing, or passive imaging. In this case, the coupling to the incident THz radiation is accomplished by means of multi-octave planar micro-antennas. After selecting the antenna shape as angular, self-complementary, and/or self-similar, various antenna geometries are studied, namely, bow tie, Sierpinski fractal, sinuous, and log-periodic. Delivered power, directivity, and radiation pattern are discussed. According to seven criteria, the log-periodic antenna is preferred.

Keywords Self-complementary antenna • Multi-octave antenna • THz imaging

1 Introduction

After four decades of applications in the space sciences, the terahertz field (THz, from 500 GHz to 10,000 GHz, typically) is covering a growing range of applications: atmospheric and environment studies, security, health, transport, etc. [1]. This growth has been accompanied by the introduction into the market of imaging systems based on (sub)THz transients triggered by infrared (IR) brief (fs) pulses. These systems inherently exhibit, however, some limitations in terms of sensitivity, resolution, and image acquisition speed, for instance.

The approach considered in this chapter deals with the detection of continuous waves (CW) in the THz, using innovative bolometric sensors working as mixers for

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heterodyne reception (with down conversion to the GHz range), as needed for radio astronomy, remote sensing, or passive imaging.

The superconducting hot electron bolometer (HEB) offers a competitive choice to other mixer technologies in the THz range [2]. An HEB consists of an ultrathin (a few 10 nm thick) superconducting micro-bridge, that is, of submicron lateral dimensions, connected to a planar antenna because the superconductor is totally reflective to the THz radiation (Fig. 1). Besides, by using a high- T_C material such as YBaCuO [3], the cooling requirements are moderate (60–80 K, typically).

Our targeted 1–7 THz range, significantly larger than the center frequency, requires ultra-wideband (UWB) antenna technology, so to ensure a correct impedance match between antenna and HEB through the band. This approach contrasts with the also used resonant antenna scheme [4]. To reach the match, the antenna impedance should be close to the resistance of the HEB at mid-transition, i.e., $50-150 \ \Omega$ (Fig. 2). The antenna function is also to couple the THz radiation propagating in free space to a sensing device much smaller than the wavelength.

The UWB antenna selection criteria are enunciated in Sect. 2. Section 3 is devoted to the description of the four selected antennas, whereas Sect. 4 is devoted to establish comparisons and select a final antenna type.

2 Selection Principles

We are considering here integrated planar antennas because of their advantages, as compared to waveguide coupling [5], such as cheaper fabrication, better accuracy, and robustness, particularly at higher frequencies. Planar antennas are suitable for array implementation, as well [6]. To design a UWB antenna, three leading principles were adopted: (1) Angular constant (*A*) structures (e.g., bow tie or spiral) are frequency independent, the bandwidth being in practice limited by the smallest and largest features. (2) Self-complementarity (SC) is the property of invariance as


the metal part is replaced by the dielectric part, and vice versa. As derived from Mushihake's results [7], the input impedance of a self-complementary antenna at an air/dielectric (of constant ε_r) interface is $Z_{ant} = (\eta/2)[(\varepsilon_r + 1)/2]^{-1/2}$, where η is the characteristic impedance of free space [6]. (3) Self-similarity (SS) is the property of invariance that the electromagnetic behavior can be similar at the multiple scales where the shape reproduces itself.

According to these principles, four antennas were selected as bow tie (A + SC), Sierpinski fractal (SS), sinuous (SC), and log-periodic (SC).

3 Selected Antennas: Modeling and Description

3.1 Methodology

A number of constraints are linked to the application. First, the antenna substrate should be compatible with the superconducting material YBaCuO. MgO single crystal (dielectric constant $\varepsilon_r = 10$) was chosen for that purpose, the thickness of which was fixed to 250 µm for manufacturing reasons. A large number (some 50) of high-order modes can be excited in this electrically thick substrate. Second, the substrate being mounted in a metal housing (the mixer block of the heterodyne receiver), a ground plane is present. Although preventing backside radiation, standing waves and the associated resonances will tend to degrade the antenna return loss. For a self-complementary geometry on this substrate, the antenna resistance is ≈80 Ω , i.e., well matched to the HEB device resistance (see Fig. 2).

The simulations were performed with the CST Microwave Studio[®] software. Ludwig-3 coordinates were used, with *z*-axis pointing at the boresight direction (elevation angle $\theta = 0$), *xz* and *yz* defining the *E*- (azimuthal angle $\varphi = 0$) and



Fig. 3 Photographs of large-scale models. (a) Quarter-pie bow tie—outer diameter: 52.2 mm. (b) Sinuous—outer diameter: 52.4 mm. (c) Log-periodic—outer diameter: 46.4 mm

H ($\varphi = 90^{\circ}$)-planes, respectively. In order to cross-correlate simulation with experiment, large-scale microwave models were built with a size factor of ×400 by wavelength; the actual micro-antenna bandwidth of 1–7 THz so became 2.5–17.5 GHz. We kept $\varepsilon_r = 10$ for the scaled model substrate materials. The antennas were manufactured on Rogers RT/duroid[®] 6010LM copper-laminated substrates of thickness 1.27 mm; those were glued to an ECCOSTOCK[®] HiK500F slab of thickness 10 cm and sides 30 × 30 cm², for further testing in an anechoic chamber. Some of these models are shown in Fig. 3.

3.2 Bow-Tie Antenna

Bow tie is a relatively historic antenna and one among the best documented in UWB literature [8]. Later on, in different domains, this antenna has been exploited extensively such as in UWB microwave radar or cancer detection. The classical bow-tie antenna is shown in Fig. 4a (with angle $\alpha = 90^{\circ}$ for a self-complementary geometry). To avoid sharp edges or corners that degrade the return loss, the triangular shape can be transformed into a smoother quarter-pie structure (Fig. 4b), with an improvement at lower frequencies. The upper and lower bandwidth limits are fixed by the inner (R_{in}) and outer (R_{out}) radii, respectively.

3.3 Sierpinski Fractal Antenna

The application of fractal shapes to the electromagnetic interaction domain has been signaled since two decades [9]. To exploit fractal structures for a UWB antenna, we chose the planar Sierpinski gasket for this purpose.

According to the Sierpinski gasket design, the original gasket is constructed by subtracting a central inverted triangle from a main triangle shape. After the sub-traction, three equal triangles remain on the structure, each being half of the size of



Fig. 4 Sketches of the designed THz bow-tie antennas: (a) triangular and (b) quarter pie. The critical dimensions are $R_{in} = 7.5 \ \mu m$ and $R_{out} = 58.1 \ \mu m$ (bandwidth 0.9–7 THz)



the original one. By iterating the same subtraction procedure on the remaining triangles, the fractal Sierpinski gasket is obtained, which exhibits a self-similar structure [10]. The iteration steps and resulting shapes are shown in Fig. 5a–e. Choosing this antenna allows to seek for improvements and compare with a bow-tie antenna exhibiting the same dimensions, approximately (Fig. 6).

3.4 Sinuous Antenna

Among the frequency-independent antennas, the sinuous antenna is a good candidate because of its broadband and dual-polarized nature. Still seeking for a selfcomplementary antenna, the genesis of such a planar sinuous scheme could be derived from the original conical geometry [11]. This geometry led us to develop a sinuous curve using the equation $\phi = (-1)^p \alpha_p \sin[\pi \ln(r/R_p)/\ln(\tau_p)] \pm \delta$, where *r* and ϕ are the polar coordinates, α_p the angular width of the *p*th arc (*p* = 1 to *n*), Fig. 6 The THz fractal antenna ($l_{min} = 10 \ \mu m$, $l_{max} = 114 \ \mu m$)

Imax

Fig. 7 The designed THz sinuous antenna, with parameters n = 5 arms, scale factor $\tau_p = 1.33$, $R_{\rm in} = 4.8 \ \mu m$, and $R_{\rm out} = 64 \ \mu m$

 R_p the *p*th radius, $\tau_p = R_{p+1}/R_p$ the scale factor, and δ the angular parameter for self-complementary condition.

The self-complementary THz antenna was designed by defining the bandwidth limits from the active region of radius R_p as $2 R_p(\alpha_p + \delta) \approx \lambda_{\text{eff}}/2$ with $\alpha_p = \pi/4$ and $\delta = \pi/2$. The effective dielectric constant of the air/substrate interface is taken into account in λ_{eff} . The final design is shown in Fig. 7 for a 1–7 THz bandwidth.

3.5 Log-Periodic Antenna

The last geometry we consider is the log-periodic antenna. The idea at its early introduction was to add teeth/arms to the bow-tie scheme and let the currents travel on those arms [12]. The geometrical parameters of a planar log-periodic antenna are detailed in Fig. 8. The key parameter is the ratio of sequential radii or scale factor







 $\tau = R_{p+1}/R_p$ (p = 1 to 2n). In addition, a leading principle in UWB antenna design is self-complementarity, which is ensured, provided the sum of the flare angle α and arm angle β is $\pi/2$. Finally, empirical considerations lead to that one arm should be considered as active only when its length $L = \lambda_{eff}/4$, where the effective wavelength $\lambda_{eff} = \lambda_0/\varepsilon_{eff}/2$; $\varepsilon_{eff} = (\varepsilon_r + 1)/2$ is the effective permittivity of the antenna medium (see Sect. 2) and λ_0 the free space wavelength.

From the targeted bandwidth limits (1–7 THz), one can then determine the minimum and maximum teeth lengths that are related to the inner and outer radii as $L_{\min} = R_{in}\beta$ and $L_{\max} = R_{out}\beta$, respectively, as detailed in [13]. Initial tests with an HEB detector at 2.5 THz (n = 15, 0.3-10 THz bandwidth) were reported in [3].

4 Selected Antennas: Comparison

4.1 Delivered Power

By normalizing to 1 the power coupled to free space, the delivered power from the antenna to the HEB is $1 - |S_{11}|^2$, where S_{11} is the reflection coefficient referenced to 100 Ω (the HEB average resistance; see Fig. 2). The results are plotted in Fig. 9. At the center frequency of 2.5 THz, the best antenna is the fractal by 84 %. The other antennas are more or less the same (around 60 %). Although the log-periodic antenna has sharp resonances in comparison to the bow tie, due to well-matched bands, it is better than the bow tie in some bands, but its coupling is lower than the bow tie at the other frequencies. Sinuous is well matched at low frequencies.

4.2 Directivity and Radiation Pattern

The directivities are plotted in Fig. 10 at the central frequency of 2.5 THz. The most directive in the boresight direction are the bow tie and sinuous by 10–12 dB. For the bow tie, there are peaks at $\theta = 15^{\circ}$ and $\theta = 50^{\circ}$ (*E*-plane) and at $\theta = -50^{\circ}$







Fig. 10 For the selected

elevation angle

antennas, directivity at 2.5 THz as a function of

(*H*-plane). The log-periodic main lobe is the widest. On the whole, the Sierpinski fractal shows weak main lobe characteristics. The in-plane directivity ($\theta = \pm 90^{\circ}$) is important for cross talk issues in antenna imaging arrays. There is a low in-plane cross talk value ($\theta = \pm 90^{\circ}$) for the sinuous and the fractal. For the log-periodic, the cross talk value is, however, the lowest in both *E*- and *H*-planes. Those results were confirmed by measurements on large-scale models (Fig. 3) [13].

Criterion	Bow tie	Fractal	Sinuous	Log-per
Chieffon	Bow the	1 Iuctui	Sindous	Log-per.
Delivered power	4	1	2	3
Gain/boresight	1	4	2	3
Gain/nondirective	3	4	2	1
Cross talk (in-plane)	2	4	3	1
Polarization sensitive	4	3	1	2
Manufacturability	1	4	3	2
Output coupling to IF	4	3	2	1
Final score	19→3rd	$23 \rightarrow 4$ th	$15 \rightarrow 2nd$	$13 \rightarrow 1st$

Table 1 For the four selected antennas, ranking according to various criteria

4.3 Antenna Selection Criteria Summary

To rank the antennas, we have chosen seven selection criteria, as listed in Table 1. Delivered power, gain at boresight, nondirective gain (for passive THz imaging, a broad main lobe is preferred), and low in-plane cross talk (for imaging antenna arrays) can be ranked from our previous discussions. Low sensitivity to polarization (passive imaging requirement) was ranked according to the computed axial ratio. Manufacturability was also considered: for instance, it is easier to manufacture a bow tie than a fractal structure, as the latter requires high resolution on the whole antenna area. Coupling to intermediate frequency (IF) lines is related to the easiness of routing the output signals from an antenna array. In those criteria being added, log-periodic comes first, although close to the sinuous structure.

5 Conclusion and Further Developments

Four types of UWB antennas were investigated before finally selecting the log-periodic antenna. The substrate, having a high dielectric permittivity, stores the energy and prevents optimizing the coupling to free space. A first improvement was to put a back metal on the substrate and so avoid back radiation. Another improved solution will be to put a high permittivity (e.g., silicon) focusing lens over the substrate. Both improvements are compatible with THz HEB mixer block technology. Antenna arrays for THz imaging are considered in the next chapter.

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UWB Antennas for CW Terahertz Imaging: Cross Talk Issues

A.J. Kreisler, I. Türer, X. Gaztelu, and A.F. Dégardin

Abstract There is a strong need for wideband and sensitive receivers in the terahertz (THz) region. When considering imaging issues, cross talk between neighbor pixels is of key importance. The case of continuous wave imaging with superconducting hot-electron bolometers as heterodyne mixing sensors is considered, when the coupling to the incident THz radiation is accomplished by means of planar multi-octave self-complementary log-periodic antennas. The minimum distance between two antennas is considered in terms of resolution, diffraction, and electromagnetic coupling criteria. Back-end cross talk between the output lines at the intermediate frequency is briefly addressed.

Keywords THz imaging array • Log-periodic antenna • Cross talk

1 Introduction

The terahertz field (THz, from 500 GHz to 10,000 GHz, typically) is nowadays encompassing a large range of applications: spectroscopy, space and environment studies, security, health, transport, etc. [1]. The approach considered in the previous chapter dealt with the detection of continuous waves (CW) in the THz, using innovative bolometric sensors working as mixers for heterodyne reception (with down conversion to the GHz range). In this chapter, we are considering imaging issues and more specifically cross talk between neighbor pixels, which is of key importance for the system spatial resolution.

The main ingredients of our design are summarized below in Fig. 1. An elementary 3-pixel demonstrator is sketched in Fig. 1a; Fig. 1b shows the pixel detail, where a planar micro-antenna (of the self-complementary log-periodic type

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Fig. 1 (a) Sketch of a 3-pixel passive imaging demonstrator with heterodyne mixing. (b) Each pixel consists of a planar antenna structure connected to a HEB superconducting micro-bridge mixer [4]. (c) Superconducting transition of a high- $T_{\rm C}$ oxide HEB

in our design [2]) feeds the sensing device, a superconducting hot-electron bolometer (HEB); Fig. 1c shows the superconducting transition of a high critical temperature oxide (YBaCuO) HEB that requires moderate cooling [3]. The micro-antenna should be impedance matched to the HEB device, i.e., $50-150 \Omega$.

Log-periodic planar micro-antenna useful properties for cross talk evaluation are recalled in Sect. 2. Section 3 is devoted to the front-end THz and back-end GHz cross talk performances.

2 Planar Log-Periodic Antenna Properties

We consider planar antennas designed to exhibit UWB properties according to the self-complementarity principle (invariance as the dielectric part is replaced by the metal part, and vice versa). From Mushihake [5], the input impedance of a self-complementary antenna is frequency independent: $Z_{ant} = (\eta/2)\varepsilon_{eff}^{-1/2}$, where η is the free-space impedance and ε_{eff} the effective permittivity of the antenna medium.

A planar log-periodic antenna, with *n* arms on each side of the bow-tie trunk defined by the flare angle α , is sketched in Fig. 2a. The self-complementarity condition is $\alpha + \beta = \pi/2$, where β is the arm angle. The arm ratio $\tau = R_{p+1}/R_p$,



Fig. 2 The 8-arm log-periodic antenna designed for the 0.9–7 THz bandwidth on MgO substrate with metal back plane. (a) Sketch. (b) For the design values ($\tau = 1.1365$, $R_{in} = R_1 = 7.5 \mu m$, $R_{out} = R_{17} = 58.1 \mu m$, $\alpha = \pi/2 - \beta = \pi/3$ [2]), effective wavelength and diameter of the radiating disk as a function of frequency. D_{max} was evaluated at the 14 resonant frequencies given by the slot-line model [6] and extrapolated according to Fig. 3 results

where R_{p+1} and R_p are the outer and inner radii of a given arm, respectively, is obtained from the targeted bandwidth frequency limits f_{max} and f_{min} as

$$\tau = \sqrt[2n]{f_{\text{max}}/f_{\text{min}}} = \sqrt[2n]{R_{\text{max}}/R_{\text{min}}},$$
(1)

where R_{max} and R_{min} are the outer and inner radii, respectively. The sequence of arm radii is simply generated as

$$R_{p+1} = R_{\min} \tau^p (p = 0 \dots 2n). \tag{2}$$

The absolute dimensions are deduced from the effective wavelength $\lambda_{\text{eff}} = \lambda_0/\varepsilon_{\text{eff}}^{4/2}$, where λ_0 is the free-space wavelength. Due to our electrically thick MgO substrate ($\varepsilon_r = 10$) with metal back plane [2], the empirical expression $\varepsilon_{\text{eff}} \cong \varepsilon_0(\varepsilon_r + 5)/2$ (ε_0 is the free-space permittivity, and ε_r is the dielectric constant of the substrate material) was preferred to the semi-infinite media expression $\varepsilon_{\text{eff}} = \varepsilon_0(\varepsilon_r + 1)/2$ [7].

Moreover, the antenna operation is such that one usually considers only one arm being mainly active at a given frequency, i.e., behaves as a quarter-wavelength resonator $L = \lambda_{eff}/4$, with $L = \beta R_p (1 + \tau)/2$ the mean length of the concerned arm. It has been recently shown, however, that in a preferred formulation, one should consider two successive arms behaving as a half-wavelength slot-line resonator $L' = \lambda_{eff}/2$, with $L' = \beta R_p (1 + \tau) + R_p (\tau - 1)$ the resonator length [6].

Those resonance effects are illustrated in Fig. 3. The active region of the antenna—identified by the higher surface currents—roughly appears as a disk, the diameter D_{max} of which can be considered as proportional to λ_{eff} . As computed from the slot-line model, D_{max} vs. frequency and λ_{eff} is shown in Fig. 2b.



Fig. 3 For the 8-arm log-periodic antenna, surface currents at (a) 0.8 THz, (b) 2.4 THz, (c) 4 THz, and (d) 6.8 THz. At low frequency, the antenna radiates like a bow tie. At higher frequencies, the active region is delimited by a disk that shrinks with decreasing wavelength

3 Cross Talk Studies

3.1 Imaging Array Spatial Resolution

Apart from the obvious limitations to resolution occurring from the maximum pixel density such as the antenna shape or the routing of output signal lines, electromagnetic cross talk between neighbor antennas or output lines will be evoked in the following sections. The fundamental limitation due to diffraction by the front-end aperture is considered first. As the incoming radiation is usually focused by a lens [8], we consider the Airy diffraction pattern radius at wavelength λ from a lens of diameter *d* and focal length *f*:

$$r_0 \approx 1.22 \frac{\lambda f}{d}.\tag{3}$$

We take the example of a hemispherical silicon lens (with a planar extension length)—widely used in mm or sub-mm waves [9]. From elementary geometrical optics, the *f*-number of such a lens is $f/d = \frac{1}{2}n/(n-1) \approx 0.7$ (refractive index of silicon $n \approx 3.42$). Taking the effective permittivity contrast between the lens and



log-periodic antenna media into account, one finally obtains $r_0 \cong 0.7D_{\text{max}}$. The diffraction disk is therefore satisfactorily matched to the active antenna region (cf. Fig. 2b). Moreover, the minimum distance between antennas will not be hampered by the diffraction limit.

3.2 Methodology for Cross Talk Evaluation

Considering two identical planar antennas A and B at the distance d_{AB} (Fig. 4), we shall evaluate the in-plane cross talk C_{AB} . Assuming antenna A is transmitting and antenna B is receiving, we wish to use—for the sake of simplicity—the Friis transmission equation [10]:

$$C_{\rm AB} \equiv [P_{\rm B}/P_{\rm A}]_{\rm dB} = G_{\rm A}(\theta = \pi/2) + G_{\rm B}(\theta = \pi/2) - 20\log\left[\frac{d_{\rm AB}}{\lambda_0}4\pi\sqrt{\varepsilon_{\rm eff}}\right], \quad (4)$$

where the gains are evaluated at $\theta = \pi/2$ elevation angle along the d_{AB} direction. It should be stressed, however, that Eq. 4 is only valid provided that far-field conditions are satisfied [11]. First, d_{AB} should be larger than the Rayleigh distance $d_{R} = 2D^{2}/\lambda$: in our case, $D = D_{max}$ (the active disk diameter) and $\lambda = \lambda_{eff}$. The other two conditions—with the same notations—are $d_{AB} \gg D_{max}$ and $d_{AB} \gg \lambda_{eff}$. In our actual situation, however, $D_{max} \approx \lambda_{eff}$ so that only one condition is needed, usually stated as $d_{AB} > 3\lambda_{eff}$ in this case [12]. We are finally left with the two conditions:



$$d_{\rm AB} > d_{\rm R} = 2D_{\rm max}^2 / \lambda_{\rm eff}$$
 "Rayleigh criterion" (5a)

and

$$d_{\rm AB} > 3\lambda_{\rm eff}$$
 " 3λ criterion." (5b)

As shown in Fig. 5, both conditions are satisfactorily satisfied for our design $(d_{AB} = 7R_{out})$ within the log-periodic antenna operating frequency range.

The simulations were performed with the CST Microwave Studio[®] software. We also built large-scale microwave models with a size factor of ×400 by wavelength while keeping $\varepsilon_r = 10$ for the scaled model substrate materials. The 1–7 THz bandwidth so became 2.5–17.5 GHz. The antennas were built on Rogers RT/duroid[®] 6010LM copper-laminated substrates that were glued to a $10 \times 30 \times 30$ cm³ slab (ECCOSTOCK[®] HiK500F) for testing in an anechoic chamber (Fig. 6).

3.3 Front-End Cross Talk Results

We first consider the cross talk between log-periodic structures oriented according to Fig. 4, i.e., the x'-axis is parallel to the yz-planes (or H-planes) of the antennas, as in Fig. 2a. The cross talk decreases as frequency increases, as expected (Fig. 7), with a value close to -20 dB at the bandwidth lower frequency. This latter value is in line with previous simulation results at 1 THz that exhibited an xy-plane (elevation angle $\theta = \pi/2$) directivity value of \cong -9 dB for one antenna [2].

Two other orientations are then considered: the antennas are orthogonally oriented in Fig. 8a, whereas the x'-axis is parallel to the xz-planes (or E-planes) of



Fig. 6 Photograph of the ECCOSTOCK[®] slab with two large-scale antennas (*inset*: log-periodic model)





the antennas in Fig. 8b. In the latter case, the higher cross talk at low frequencies can be related to the bow-type-like stronger radiation along the *x*-axis (Fig. 3a).

The use of a focusing dielectric lens was mentioned in Sect. 3.1. It allows avoiding power loss due to substrate modes, provided the lens has the same dielectric constant as the antenna substrate [8]. Due to the lens/free-space interface mismatch, an anti-reflection treatment of the lens should be, however, required.



Fig. 8 For alternate antenna orientations, cross talk as a function of frequency. (a) Orthogonally oriented antennas. (b) x'-Axis parallel to antenna xz- (or E-) planes of Fig. 2a



A silicon lens ($\varepsilon_r = 11.7$) will offer a satisfactory match with our MgO substrate ($\varepsilon_r = 10$). The use of such a lens with a 3-pixel demonstrator (Fig. 1a) is illustrated in Fig. 9. The effect on cross talk of introducing the high permittivity lens medium is shown in Fig. 10, with a 15 dB improvement, at least below 4 THz.



3.4 Back-End Cross Talk

We are briefly considering here the cross talk between output lines at the intermediate frequency (IF). As sketched in Fig. 9, there are two line types. On the MgO substrate bearing the HEBs and antennas, 50 Ω microstrip lines (μ SL) are connected to those latter by means of matching tapers. The microstrip lines are bonded to 50 Ω coplanar waveguides (CPW) on alumina substrates. For both line types, the cross talk has been simulated through the whole IF bandwidth (1–32 GHz). Cross talk increases with increasing frequency. For μ SL, it remains below –24 dB (first neighbors) and –35 dB (second neighbor); for CPW, it remains below –18 dB (first neighbors) and –29 dB (second neighbor) [4].

4 Conclusion and Further Developments

We have simulated the front-end cross talk between THz imaging log-periodic micro-antennas designed for the 1–7 THz bandwidth. For neighbor pixels, the cross talk decreases as frequency increases and remains below -25 dB above 1.5 THz for the three considered antenna orientations: common *H*-plane, common *E*-plane, and orthogonal. At 1 THz, the value is $\cong -20 \text{ dB}$ for common *H*-plane and orthogonal but $\cong -15 \text{ dB}$ for common *E*-plane. These results are confirmed by measurements on large-scale microwave models. To design a low-cross talk imaging array, one should therefore consider arrangements where any antenna is surrounded by first neighbor orthogonal antennas.

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Evaluation of Imaging Algorithms for Prototype Microwave Tomography Systems

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Abstract We present imaging results using scattered field data obtained from microwave tomography prototypes and evaluate various state-of-the-art inversion algorithms and regularization schemes. The algorithms are based on either the Gauss-Newton inversion or the contrast-source inversion methods. The evaluation is based on image reconstruction accuracy of experimental datasets.

Keywords Microwave tomography • Imaging algorithms • Microwave imaging

1 Introduction

Microwave tomography (MWT) is an emerging modality which has gained interest in various applications such as medical imaging [1] and geophysical surveying [2]. In MWT, the goal is to determine the shape, location, and electrical properties (permittivity and/or conductivity) of an object of interest (OI) confined within a chamber. With the OI immersed in a known background medium, it is successively illuminated by various sources of electromagnetic radiation. The variation of the electrical properties between the OI and the background medium results in a field that is measured at several observation points outside the OI. This field is known as the *total field*. In the absence of the OI, the field produced by the same sources is referred to as the *incident field*. The difference between the total and incident fields is defined as the *scattered field*.

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Mathematically, MWT is associated with an inverse scattering problem that is nonlinear and ill-posed. The nonlinearity arises due to the existence of two unknowns: the field within the OI and the electrical properties of the OI. The nonlinearity can be dealt with using iterative-based optimization algorithms that attempt to minimize a cost functional by estimating variables representing the unknown electrical properties of the OI. Such algorithms utilize the measured scattered field data. Furthermore, due to ill-posedness of the inverse problem, its solution is not guaranteed to exist, be unique or be stable (i.e., small change in the measured fields can yield large changes in the reconstructed unknowns). The ill-posedness of the problem is treated via different regularization methods.

The purpose of the research work presented herein is to review and evaluate two state-of-the-art iterative-based optimization algorithms: the Gauss-Newton inversion (GNI) method and the contrast-source inversion (CSI) method. While a detailed description of both algorithms is beyond the chapter's scope, the presented work will emphasize on the algorithms' performance in inverting experimentally collected datasets and how this performance is affected using different regularization schemes when applicable.

2 Mathematical Formulation

We consider a two-dimensional (2D) chamber that contains an imaging domain D, within which an isotropic, nonmagnetic object of interest (OI) is located. The imaging domain is immersed in a background medium with a relative complex permittivity $\varepsilon_{\rm b}(q)$, where $q \in D$ is a 2D position vector. The electrical contrast of the OI is given as $\chi(q) = (\varepsilon_{\rm r}(q) - \varepsilon_{\rm b}(q))/\varepsilon_{\rm b}(q)$, where $\varepsilon_{\rm r}(q)$ is the complex relative permittivity of the OI. Outside $D, \chi = 0$.

The chamber is illuminated successively by one of the Tx transmitters, producing an incident field E_t^{inc} defined as the field in the absence of the OI. With the presence of the OI, the resultant field in the chamber is the total field E_t that corresponds to the scattered field $E_t^{\text{sct}} = E_t - E_t^{\text{inc}}$ (see Fig. 1). Both fields are measured at receivers' locations $p \in S$, where S is the measurement surface. The scattered field on S is governed by $E^{\text{scat}}(p) = G_S[\chi(I - G_D\chi)^{-1}(E^{\text{inc}})]$ where $G_S[.]$ and $G_D[.]$ are the data and domain operators, respectively [3]. The data operator returns the scattered field on the measurement surface S, whereas the domain operator returns the scattered field inside the imaging domain D. These operators can be presented in integral form using the appropriate Green's function [3] or can be solved by direct discretization of the Helmholtz equation governing the problem using techniques like the finite-element method (FEM) [4]. In this work, the GNI algorithm utilizes the integral form of these operators, whereas the CSI algorithm is formulated using the FEM discretization of the operators.



Fig. 1 Two-dimensional geometrical models for the microwave tomography problem (a) with OI (total field measurement) and (b) without the OI (incident field measurement). Tx and Rx represent the transmitting and receiving antennas, respectively. The domain D, which contains the OI, is the imaging domain. The domain S, which contains Tx and Rx antennas, is the measurement surface and is outside the OI

3 Inversion Algorithms

Broadly speaking, iterative-based inversion algorithms can be divided into two types: conventional and modified-gradient methods.

The conventional type attempts to minimize a cost functional solely in terms of the scattered field measured on surface *S*. The GNI method belongs to this class of algorithms, and its functional is written as

$$C^{\text{GNI}}(\chi) = \frac{\sum_{t=1}^{T_{\chi}} \left\| G_S \left[\chi (I - G_D \chi)^{-1} \left(\boldsymbol{E}_t^{\text{inc}} \right) \right] - \boldsymbol{E}_{\text{meas}, t}^{\text{scat}} \right\|_S^2}{\sum_{t=1}^{T_{\chi}} \left\| \boldsymbol{E}_{\text{meas}, t}^{\text{scat}} \right\|_S^2}, \qquad (1)$$

where $E_{meas,t}^{\text{scat}}$ is the experimentally measured scattered field for a transmitter *t*. At each iteration, the contrast χ is updated, and the resultant scattered field is evaluated. This requires a forward solver call that can be a computational burden [3].

The modified-gradient-type algorithm minimizes an objective functional in terms of both the scattered field on S and the total field in D. This leads to a larger number of unknowns to solve for; nevertheless, forward solver calls are avoided. The CSI method is of the modified-gradient type and its functional given as

$$C^{\text{CSI}}(\chi, \boldsymbol{W}_{t}) = \frac{\sum_{t=1}^{T_{x}} \left\| \boldsymbol{E}_{\text{meas}, t}^{\text{scat}} - \boldsymbol{G}_{S}[\boldsymbol{W}_{t}] \right\|_{S}^{2}}{\sum_{t=1}^{T_{x}} \left\| \boldsymbol{E}_{\text{meas}, t}^{\text{scat}} \right\|_{S}^{2}} + \frac{\sum_{t=1}^{T_{x}} \left\| \chi \boldsymbol{E}_{t}^{\text{inc}} - \boldsymbol{W}_{t} + \chi \boldsymbol{G}_{D}[\boldsymbol{W}_{t}] \right\|_{D}^{2}}{\sum_{t=1}^{T_{x}} \left\| \chi \boldsymbol{E}_{t}^{\text{inc}} \right\|_{D}^{2}}, \quad (2)$$

where $W_t = \chi E_t$ is the contrast-source variable. At each step, the contrast χ and the contrast-source W_t variables are updated in an interlaced fashion to minimize the CSI functional [4].

4 **Regularization Methods**

Due to the ill-posedness of the inverse problem, the optimization algorithms are used in conjugation with an appropriate regularization technique to produce reasonable results.

The regularization techniques used with GNI can be classified into two categories distinguished by the type of the cost functional to be minimized [3]. The first strategy minimizes the data misfit functional, $C^{\text{GNI}}(\chi)$, by updating the unknown variables and then regularizing the solution using either penalty methods or projection methods. Tikhonov regularization is an example of a penalty method, whereas Krylov subspace method is an example of a projection method. The second strategy regularizes the cost functional $C^{\text{GNI}}(\chi)$ first and then applies the GNI algorithm on the regularized nonlinear cost functional. Additive, multiplicative, and additive-multiplicative regularizations are classified under this category.

In contrast to GNI, the CSI functional is already regularized by the Maxwell regularizer which is the second additive term in the cost functional, $C^{\text{CSI}}(\chi, W_t)$. Nevertheless, the CSI reconstruction results can be enhanced by augmenting the functional by multiplicative or additive regularizers [4]. Regularization in CSI only affects the update of the contrast variables χ .

4.1 Multiplicative Regularization

A successful regularizer used in conjugation with both GNI and CSI is the weighted L^2 -norm total variation multiplicative regularizer (MR). This regularizer provides an adaptive, edge-preserving regularization with the ability to suppress noise from measurement data. In conjugation with MR, the functional of GNI is given as

$$C_n(\chi) = C^{\text{GNI}}(\chi) \times \frac{1}{A} \int_D \frac{|\nabla \chi(\boldsymbol{q})|^2 + \alpha_n^2}{|\nabla \chi_n(\boldsymbol{q})|^2 + \alpha_n^2} d\boldsymbol{q}$$
(3)

while for CSI it is

$$C_n(\boldsymbol{\chi}, \boldsymbol{W}_t) = C^{\text{CSI}}(\boldsymbol{\chi}, \boldsymbol{W}_t) \times \frac{1}{A} \int_D \frac{|\nabla \boldsymbol{\chi}(\boldsymbol{q})|^2 + \delta_n^2}{|\nabla \boldsymbol{\chi}_n(\boldsymbol{q})|^2 + \delta_n^2} d\boldsymbol{q}.$$
 (4)

In Eqs. (3) and (4), α_n and δ_n are adaptive factors automatically calculated at each step of the algorithm, while A is the total area of the imaging domain.

To account for the imbalance between the real and imaginary components of the OI's contrast and to enhance the reconstruction quality, the GNI and CSI functionals with MR can be modified as follows:

$$C_n(\chi) = C^{\text{GNI}}(\chi) \times \frac{1}{A} \int_D \frac{|\nabla \chi_R(\boldsymbol{q})|^2 + Q^2 |\nabla \chi_I(\boldsymbol{q})|^2 + \alpha_n^2}{|\nabla \chi_{R,n}(\boldsymbol{q})|^2 + Q^2 |\nabla \chi_{I,n}(\boldsymbol{q})|^2 + \alpha_n^2} d\boldsymbol{q}$$
(5)

$$C_n(\boldsymbol{\chi}, \boldsymbol{W}_t) = C^{\text{CSI}}(\boldsymbol{\chi}, \boldsymbol{W}) \times \frac{1}{A} \int_D \frac{|\nabla \boldsymbol{\chi}_R(\boldsymbol{q})|^2 + Q^2 |\nabla \boldsymbol{\chi}_I(\boldsymbol{q})|^2 + \delta_n^2}{|\nabla \boldsymbol{\chi}_{R,n}(\boldsymbol{q})|^2 + Q^2 |\nabla \boldsymbol{\chi}_{I,n}(\boldsymbol{q})|^2 + \delta_n^2} d\boldsymbol{q} \qquad (6)$$

where $\chi_R = \text{Re}(\chi), \chi_I = \text{Im}(\chi)$, and Q is a scaling factor dependent on the ratio of the real to the imaginary components' magnitudes of the OI's relative complex permittivity.

5 Inversion Results

5.1 UPC Barcelona Dataset: BRAGREG

The UPC Barcelona dataset is collected using a near-field 2.33 GHz microwave scanner system [5]. This system consists of 64 antennas distributed on a circular array of radius 0.125 m. When using one of the 64 antennas as a transmitter, the 33 antennas in front of it are active as receivers. The data collection tank is filled with a background medium of water with relative permittivity $\varepsilon_b = 77.3 - j8.66$ at 2.33 GHz. For the BRAGREG dataset, the scattering object is a human forearm, which is immersed in the tank and is surrounded by the antennas.

The GNI reconstruction results using the first and second strategies of regularization are shown in Figs. 2 and 3, respectively. The CSI result and its variant with MR are given in Fig 4. In GNI, the second regularization strategy gives better results than the first strategy especially for the imaginary component of the OI. For both GNI and CSI, the use of multiplicative regularization produces the best-quality reconstructions. With the use of MR, the contours of the bones and muscle tissues as well as that of the forearm are clear and distinguishable.



Fig. 2 BRAGREG dataset: the reconstructed relative complex permittivity of a real human forearm using GNI with (a, c) an identity Tikhonov and (b, d) a Krylov subspace regularization



Fig. 3 BRAGREG dataset: the reconstructed data using GNI with (a, d) a multiplicative regularization, (b, e) an additive regularizer, and (c, f) an additive-multiplicative regularizer



Fig. 4 BRAGREG experimental dataset: the reconstructed data using (a, c) CSI and (b, c) MR-CSI

5.2 University of Manitoba MWT System: Wood-Nylon Target

The University of Manitoba MWT system utilized here is air filled with 24 Vivaldi antennas used as transmitters and receivers [6]. The Vivaldi antennas are evenly distributed on a circle of radius 0.13 m. For each transmitter antenna, 23 measurements are collected. The measured data are calibrated using the procedure described in [6]. The dataset is acquired at a frequency of 3 GHz.

The OI consists of a wooden block and a nylon cylinder as depicted in Fig. 5a. The relative complex permittivities at 3 GHz are $\varepsilon_r^{\text{wood}} = 2.0 - j0.2$ and $\varepsilon_r^{\text{nylon}} = 3.0 - j0.03$. The reconstruction results using MR-GNI and MR-CSI are shown in Fig. 5b, d and Fig. 5c, e, respectively. The enhanced reconstruction results using a balanced form of the multiplicative regularizer are given in Fig. 6.



Fig. 5 (a) The OI consisting of a wooden block and a nylon cylinder and the reconstructions at 3 GHz using (b, d) MR-GNI and (c, e) MR-CSI



Fig. 6 Wood-nylon cylinder experimental dataset reconstruction using balanced multiplicative regularization with (a, c) GNI and (b, d) CSI

The reconstruction results for the real component of the OI are similar using either MR or its balanced form or using GNI or CSI; the algorithms predicted the real component of the OI accurately. Using MR, the imaginary part of the reconstruction is not satisfactory; on the other hand, with the use of a balanced multiplicative regularizer, the algorithms reconstructed the imaginary component of the wooden block properly with an average value of 0.11 for $- \text{Im}(\varepsilon_r)$. As for the nylon cylinder, considering it is almost lossless and the MWT system has a limited signal-to-noise ratio as well as dynamic range, the reconstruction of the cylinder's imaginary part is difficult.

6 Conclusion

Two state-of-the-art MWT inversion algorithms, GNI and CSI, have been overviewed, along with a discussion of the various regularization strategies used in association of these algorithms. The algorithms, as well as the different regularizers, were evaluated using experimental datasets. The balanced form of multiplicative regularization had the best performance among the regularizers when used either in conjugation with GNI or CSI.

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Index

A

Aircraft losses, 169, 172, 176 Analytical solutions, 32–35 Anisotropy, 159–166 Antenna factor, 72, 75 Antenna impulse response, 76, 78 Antenna theory, 13–28 Axially symmetrical structures, 137–144

B

Back projection, 433–440 Breakdown, 129, 190, 191, 196, 232, 233, 240, 251, 283–285, 299–307, 370, 415, 459

С

Cables, 13, 31, 34, 35, 37, 47, 201, 210, 213, 221, 289–296, 299–301, 304, 305, 307, 326–329, 331–340, 358, 364, 394, 399–401, 404, 405, 407, 451 Cavity resonator, 43, 45

CEM. See Computational electromagnetics (CEM)

Characteristic modes, 200, 203-207

Circular patch, 219-227

Cold plasma, 101-112

Collision frequency, 102-104

Complex images, 63, 64

Component, 13, 20, 32, 34, 50, 54, 55, 57, 64, 65, 85, 91–97, 102, 115–118, 124, 125, 128, 130, 147, 150, 151, 154, 160, 162, 163, 171, 240, 241, 255–259, 266, 268, 273–, 275, 280–282, 284, 285, 287, 291, 299–301, 336, 338, 345, 356, 371, 372, 375, 412, 413, 416, 425, 428–430, 434, 444, 447, 449, 450, 487, 491 Computational electromagnetics (CEM), 169, 176 Conformal electrodes, 238 Coupling, 13, 31–47, 57, 114, 135, 169, 170, 190, 191, 246, 267–269, 273, 280, 283–284, 287, 290, 303, 322, 331–340, 448, 458–460, 464, 469, 471 Critical infrastructure, 189, 266 Crosstalk, 458–460, 479–481

D

De-embedding technique, 394–397, 401 Detection, 89–91, 94, 97, 101, 269, 275–276, 291, 296, 331, 345–353, 356, 362, 367, 379–391, 426, 430, 434–440, 463, 466, 473 Dielectric lens, 245, 252, 479 Digital device, 309–317 Discrete random media, 89–98 Dispersion, 85, 90, 102, 110, 159, 253, 371, 372 Distortion analysis, 80–82, 85–86 Double ridged waveguide, 369–376 Dual polarized, 189–196, 467

E

EAC. See Exact absorbing conditions (EAC)
Electric field, 4, 28, 60, 73, 80–82, 84–86, 105–109, 114, 116, 118–119, 124, 125, 150, 161, 162, 170, 172, 181, 205, 224–226, 233, 236, 240, 253–258, 260, 299, 301, 304, 305, 314, 332–340, 356, 358, 362, 369, 371, 372, 376, 403
Electric-field sensor, 412–414
Electromagnetic attack warning system, 362

- Electromagnetic compatibility (EMC), 8, 13, 31, 32, 49, 50, 270, 280, 319–329, 332, 345 Electromagnetic environment, 113–125, 270,
- 272, 274, 345, 388
- Electromagnetic field sensor, 79-86
- Electromagnetic induction (EMI), 426, 427, 429–432
- Electromagnetic interference (EMI), 32, 265–277, 280, 289, 328, 356, 380–391, 401
- Electro-optics, 321, 356, 415
- EMC. See Electromagnetic compatibility (EMC)
- EMI. See Electromagnetic interference (EMI)
- EMP detector, 368
- ENG-metamaterial, 220-222, 226, 227
- Enhancement of image bearing component, 92 Equivalent circuit, 114, 128–130
- Exact absorbing conditions (EAC), 137, 138, 141

F

- FDTD. See Finite difference time-domain (FDTD)
- Finite difference time-domain (FDTD), 6, 9, 120, 121, 125, 129, 135, 159–166, 170, 211, 212, 215, 372
- Frequency-domain (FD), 80, 83, 107, 109, 163, 186, 212

G

Gain, 71, 72, 75, 77, 78, 80, 85, 133, 192–195, 220, 224–227, 245, 247, 248, 457, 471, 477
GPR. See Ground-penetrating radar (GPR)
Grounding systems, 113–125
Ground-penetrating radar (GPR), 199

H

Helical antenna, 246–247 Hermite–Gauss functions, 82–83, 86 HF antenna, 59 Higher order modes, 50, 53 High frequency, 13, 53, 84, 85, 110, 112, 159, 210, 252, 260, 291, 293, 295, 296, 356, 393–401, 444, 448, 449, 454, 455, 457, 459, 460 High-intensity radio frequency (HIRF), 169–176

- High power electromagnetic field detector, 355–367 High-power electromagnetics (HPEM), 189–196, 265, 266, 270, 273, 274, 379–391, 411–418 measurements, 411–418 High power microwaves (HPM), 246, 345–353, 369–376, 383 HIRF. *See* High-intensity radio frequency (HIRF)
- HPEM. See High-power electromagnetics (HPEM)
- HPM. See High power microwaves (HPM)
- Hybrid method, 125
- Hydrogen switch, 252, 256, 259

I

IEMI. See Intentional electromagnetic interference (IEMI) Imaging, 89-98, 433-440, 463, 464, 470, 471, 473, 474, 476–477, 481, 483–485, 487 Imaging algorithms, 435, 483-491 Immunity, 296, 300-302, 304-306, 309-317, 345.415 Impulse radiating antenna (IRA), 72, 102, 106, 108-110, 189, 251, 252, 331-340 Impulse radiation, 192 Integrated directional coupler, 449, 450 Integrated UWB sensor, 445 Intentional electromagnetic interference (IEMI), 266, 279-287, 289, 356, 362, 364, 367, 379-391 Inverse Fourier transform, 173 Ionosphere, 101-112 IRA. See Impulse radiating antenna (IRA)

K

Kraus monofilar axial mode, 247

L

Land mines, 425–432 Layout variation, 296 Lightning modeling, 127–135 Log-periodic antenna, 468–469, 471, 474–478

M

Magnetic polarizability, 425–432 Marx generator, 240–242 Index

Media converter (MCs), 404, 405, 407, 408.410 Method of moments (MoM), 51, 59, 60, 80, 84, 114, 203, 290 Microstrip resonator, 159–166 Microwave imaging, 483-491 Microwave pulse measurement, 369-376 Microwave tomography, 483-491 Minimization algorithm, 408, 410 Minimum phase, 179-186 Mode modification, 220, 225–227 MoM. See Method of moments (MoM) Morris, 8-11 M-sequence, 444-449, 453-471 Multi-target, 437, 439 Mutual coherence function, 89-98

N

Network, 128, 133, 179–181, 189, 191, 201, 213, 265, 273, 274, 280, 290, 299–307, 313, 314, 316, 320, 380, 394, 396–399, 401, 404, 405, 410, 443–451, 458 Nonlinear inverse problems, 484

0

Optical fiber, 249, 404, 406

P

Plasma frequency, 104, 105, 109, 110 Power line communication, 313 Propagation, 4, 50, 54, 60, 61, 64–69, 89–97, 101–112, 114, 121, 129, 138, 141, 147–156, 180, 184, 232, 249, 256, 268, 269, 272, 273, 301, 302, 332, 371, 398–400, 458 Pulse forming line, 240–245

R

Radiation efficiency, 142, 213, 215, 216 Radiation properties, 210, 212–215 Realized gain, 72, 75, 77, 78 Reciprocity, 74–75, 292, 296, 319–329, 427 Reciprocity theorem, 292, 296, 319–329 Resistive sensor, 369–376 RF testing, 403–410 RF weapon detector, 362 Risk analysis, 266, 269, 279–287 RWG, 203

S

Self-complementary antenna, 465, 467, 474 Sensor network, 380 Shielding, 166, 170, 172, 173, 329 Shielding effectiveness, 170, 172, 173 Shielding effectiveness assessment, 170, 319 Short pulse, 416-417, 433-440 SoC. See System-on-chip (SoC) Spiral antennas, 209-216 Stochastic collocation, 5-7 Superconductor, 162-163 Suppression of diffusion, 97 Surface wave, 64 Susceptibility, 272, 289-296 Switched oscillator, 231-238 System-on-chip (SoC), 449-451, 454-461

Т

TEM Horn, 179-186, 245-246 TEM-mode, 50, 53, 399 TEM-waveguide, 49-57, 301-303, 351, 353 Thin wires, 32-35 Threat Scenario, Effect, and Criticality Analysis (TSECA), 265-277 Through wall imaging, 433-440 Through-wall imaging radar (TWIR), 433-440 THz imaging, 471, 481 THz imaging array, 481 Time-domain (TD), 71-86, 91, 93, 107, 109, 110, 117-119, 121, 125, 128, 147, 148, 152, 154-156, 161, 172, 173, 185, 186, 193, 199, 212, 247, 302, 333, 334, 380, 384-388, 413, 426, 437 Time-domain analysis, 80 Time-domain circuit, 128 TLST. See Transmission-line super theory (TLST) Transfer function, 149-151 Transmission line, 113, 121, 337, 359 Transmission-line super theory (TLST), 13 - 28Transmission-line theory, 399 Trilateration, 434, 435, 437-440 Triple band, 219–227 TSECA. See Threat Scenario, Effect, and Criticality Analysis (TSECA) TWIR. See Through-wall imaging radar (TWIR)

U UTD, 149–151 UWB sensor, 454 UXO, 101

v

Vector fitting (VF), 147–156 Vectorial characterization, 412, 413, 415, 417 Vivaldi, 190–195, 199, 489 Vulnerability, 274

W

Wideband and ultra-wideband sources, 240 Wide-band pulse, 90, 97