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APPLIED COMPUTATIONAL ELECTROMAGNETICS SOCIETY (ACES)

NEWSLETTER

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NEWSLETTER ARTICLES AND VOLUNTEERS WELCOME

The ACES Newsletter is always looking for articles, letters and short communications of interest to ACES members. All individuals are encouraged to write, suggest or solicit articles either on a one-time or continuing basis. Please contact a Newsletter Editor.

AUTHORSHIP AND BERNE COPYRIGHT CONVENTION

President's Post

This year's ACES annual conference was held in Honolulu, Hawaii, April 3-7, 2005. The conference was held along with the IEEE conference on wireless communications. The combined event was named the 2005 IEEE/ACES International Conference on Wireless Communications and Applied Computational Electromagnetics.

The reason I am starting the President's Post with the above statement is to identify one of the new modifications that have been introduced to enhance member services and improve attendance, focus and format of the ACES annual conference. In the process, these modifications are also enhancing the financial viability of our society as can be verified by the great successes of the 2004 annual conference in Syracuse, NY and the 2005 annual conference. The 2006 annual conference will be held in Miami, Florida and promises to be another success. I would like to thank our colleagues, Professor Tapan Sarkar who led the efforts for the 2004 annual conference, and Professor Magdy Iskandar who led the efforts for the 2005 annual conference.

The new format and conference rotation are helping ACES increase its international focus, visibility and collaboration. We have been able to attract funding from industry as well as industrial exhibitors, financial support from the European Network ACE and grant funding from ARO and NSF. Such a support will help us provide broader benefits for addressing the research and development efforts for our members and increase our relevancy to industry and government agencies.

The Board of Directors met over two days during the annual conference in Hawaii and discussed many issues as well as received the budget with the financial status of the society improving. Our members will see in the near future many changes in the member services with many improvements in electronic access. More and more back issues of the Newsletter and Journal, credit card processing for membership fees and purchases and additional features on the website. I encourage all members to visit the website often to see all the new features and additions as they occur over the next few months.

An extended version of papers submitted and presented at the annual conference can be submitted for peer review and possible publication in the ACES Journal. As you know the ACES Journal is an ISI citation indexed Journal and can contribute to enhancing your career as well as research and development needs.

We are working hard and establishing liaisons with other organizations for the benefit of our members. So I encourage nonmembers to become members to get all the benefits. As you know we have lowered the membership fees to only $35.00 annually with electronic access to publications.

The next year's annual conference will be in Miami, Florida, March 12-16, 2006 at the Hyatt Regency Miami. I invite everyone to participate or organize a session. I also invite industrial colleagues to participate by organizing sessions and exhibiting their products at the conference.

I look forward to reporting on more progress in ACES operations and to your support in continuing to enhance this very important society for computational electromagnetics.

Osama Mohammed
ACES President

2004 FINANCIAL REPORT

ASSETS

BANK ACCOUNTS	**1 JAN 2004**	**31 DEC 2004**
Main Checking	16,377	13,772
Oxford Checking	0	24,021
Editor Checking	3,334	3,337
Secretary Checking	6,255	4,918
Savings	111	111
High Rate Savings	23,242	11,089
Credit Card	10,202	10,702
Transfer	0	4,767
CD 16406	14,523	14,812
CD 17227	14,204	14,437
CD 17228	14,401	14,700
CD 17673	14,413	14,625
TOTAL ASSETS:	$ 117,062	$ 131,291
LIABILITIES:	0	0
NET WORTH 31 December 2004:	$ 117,062	$ 131,291

INCOME AND EXPENSES

INCOME

Conference	53,880
Short Courses	4,065
Publications	1,885
Membership	19,598
Interest and Misc.	1,252
Miscellaneous	3,488
TOTAL INCOME	$ 84,168

EXPENSES

Conference	35,150
Short Courses	1,975
Publications	8,584
Services (Legal, Taxes, Secretarial)	5,384
Postage/Communications	7,365
Bank/Credit Card Fees	1,549
Website	7,344
Supplies and Misc.	2,177
TOTAL EXPENSES	$ 69,528

2004 NET PROFIT: $ 14,640

Notes:

1. The Oxford checking account was opened in January 2005 with the net proceeds of the 2004 conference held in Syracuse, NY in April 2004. The proceeds of the conference and the new Oxford account are included in the above report to more accurately reflect the 2004 financial status of the society.

2. The 2003 net operating profit was $2,718, while the 2004 net operating profit (including the 2004 conference profit; see Note 1 above) was $14,640.

3. The year-end net worth of ACES was $131,291, representing a increase of 12.2% over the 2003 year-end net worth.

Allen Glisson
Treasurer

Secretary's Report

ACES is actively serving a diverse group of CEM professionals via publications, the annual meeting, and formal as well as informal professional interactions. For 2004, there were sixty-nine (69) institutional, seventeen (17) expanded, fifty-five (55) intermediate, and one hundred eighty-two (182) basic members. The conference in Hawaii was well attended and the ACES Board of Directors is looking forward to seeing all of the society members, as well as new members, at the meeting in 2006 to be held in beautiful Miami, FL.

2005 ACES Board of Directors (BoD) Election

This year there are 6 candidates running for election to the ACES BoD, all of them highly experienced in the field of Computational Electromagnetics. In order to participate in the election you must log on to the ACES website and follow the links to the voting screens. The tallying is done electronically once you have made your selections. For your convenience, the candidate statements which appear in the following pages are also on the voting site as links to .pdf files.

Please participate in the election of our new BoD members and help to define the future of ACES.

Thank you,

Rene Allard
Nominations and Elections Chair

DR. LEO KEMPEL

GENERAL BACKGROUND

Leo Kempel was born in Akron, OH, in October 1965. He earned his B.S.E.E. at the University of Cincinnati in 1989 and participated in the cooperative education program at General Dynamics/Fort Worth Division. He earned the M.S.E.E. and Ph.D. degrees at the University of Michigan in 1990 and 1994, respectively.

After a brief Post-Doctoral appointment at the University of Michigan, Dr. Kempel joined Mission Research Corporation in 1994 as a Senior Research Engineer. He led several projects involving the design of conformal antennas, computational electromagnetics, scattering analysis, and high power/ultrawideband microwaves. He joined Michigan State University as an Assistant Professor in 1998 where he is conducting research in computational electromagnetics and electromagnetic materials characterization, teaching undergraduate and graduate courses in electromagnetics, and supervising the research of several M.S. and Ph.D. students. Prof. Kempel's current research interests include computational electromagnetics, conformal antennas, microwave/millimeter wave materials, mixed-signal electromagnetic interference techniques, and measurement techniques. Prof. Kempel has been awarded a CAREER award by the National Science Foundation and the Teacher-Scholar award by Michigan State University in 2002. He also received the MSU College of Engineering's Withrow Distinguished Scholar (Junior Faculty) Award in 2001.

Dr. Kempel served as the Chapter IV Vice-Chair for the Southeast Michigan chapter of the IEEE as well as the technical chairperson for the 2001 ACES Conference. He has organized several sessions at recent URSI and ACES meetings. He is an active reviewer for several IEEE publications as well as JEWA and Radio Science. He co-authored *The Finite Element Method for Electromagnetics* published by IEEE Press. Dr. Kempel is a member of Tau Beta Pi, Eta Kappa Nu, ACES, Commission B of URSI, and is a Senior Member of IEEE.

PAST SERVICE TO ACES

Leo has been a member of ACES for a number of years. He has presented papers, organized sessions, served as Vendor Chairperson for the 2000 meeting, Technical Chairperson for the 2001 meeting, and Co-Chairperson for the 2002 meeting. Dr. Kempel has been a member of the ACES Board of Directors since 2003.

CANDIDATE'S PLATFORM

ACES is making important contributions to the technical community, both over the lifetime of the society and especially in recent years. As a member of the Board of Directors for the past three years, I have seen significant changes in the society and in the profession. ACES will over the next three years make further changes to align the society with the needs of our technical constituency. We need to re-engage our members so that the society serves the needs of all CEM professionals. We need to include, increasingly, CEM professionals and organizations outside the United States and ensure that ACES meets the needs of an international community. This includes holding the conference overseas and encouraging greater involvement by our colleagues in Europe and Asia. We must also look for partnerships with complementary professional organizations such as AMTA, ACE, and IEEE. Finally, we must continue to engage colleagues across various components of our professional lives – academia, industry, and Government.

DR. OSAMA A. MOHAMMED

GENERAL BACKGROUND

Dr. Osama A. Mohammed is the current President of ACES and is an IEEE Fellow. Dr. Mohammed is a Professor of electrical and computer engineering at Florida International University, Miami, Florida. He received his Master of Science and Ph.D. degrees in Electrical Engineering from Virginia Tech., Blacksburg, Virginia in 1980 and 1983, respectively. He has many years of teaching, curriculum development, research and industrial consulting experience. He authored and co-authored more than 200 technical papers in the archival literature as well as in National and International Conference records. In addition, he has several book chapters, edited volumes, numerous technical and project reports and monographs to his credit. He specializes in Electromagnetic Field Computations in Nonlinear Devices, Design Optimization of Electromagnetic Devices, Artificial Intelligence Techniques and their Applications to Electromagnetic and Energy Systems. He is also interested in low power design considerations for Digital Mobile Telecommunication Applications He received many awards for excellence in research, teaching and service to the profession. He has extensive experience in organizing major national and international conferences. Professor Mohammed was the General Chairman of the 1993 COMPUMAG, October 31- November 4, 1993, and was Vice-Chairman of COMPUMAG-RIO, November 3-7, 1997, Brazil. He was also the General Chairman of the 1996 IEEE International Conference on Intelligent Systems Applications (ISAP'96), Orlando, Florida, January 28-February 2, 1996 as well as the General Chairman of the 1994 IEEE Southeastcon, Miami, Florida, April 10-13, 1994. He also was a member of the technical program committee for the 1996 IEEE/CEFC conference, Okayama, Japan, and the 1997 (IEEE-IEMDC), Milwaukee, Wisconsin. May 18-21, 1997. Dr. Mohammed was the editorial board chairman for the 2000 IEEE/CEFC conference, June 4-7, 2000 and the editor of the associated issue of the IEEE Transactions on Magnetics. He was recently a member of the editorial board for IEEE/CEFC Conference, Seoul, Korea, June 2004 and the 2005 IEEE-IEMDC, San Antonio, Texas, May 2005. Professor Mohammed was also one of the organizers of the 2004 IEEE Nanoscale Devices and Systems Integration Conference, Miami, Florida, February 2004. He is the General Chair of the 2006 ACES annual conference, March 2006 and the 2006 IEEE/CEFC conference, May 2006; both in Miami, Florida.

Professor Mohammed is an editor of the IEEE Transactions on Energy Conversion and an editor of the International Journal for Computation and Mathematics in Electrical and Electronic Engineering (COMPEL). Furthermore, he chaired sessions and programs in numerous National and International Conferences and has delivered numerous invited lectures and tutorials at scientific organizations worldwide. Professor Mohammed serves on several IEEE committees and boards. He is the Chairman of the Miami Section of IEEE as well as the past chairman of the Florida Council of IEEE. He was a member of the IEEE/PES Governing Board (1992-1996) and he is a currently on the board as chairman of the constitution and bylaws committee. He currently serves as chairman, officer or as an active member on several IEEE society committees, sub-committees and technical working groups.

PAST SERVICE to ACES

Dr. Mohammed is the current President of ACES since his election by the ACES BoD in March 2002 and April 2004. He has served as ACES vice president during 2001 and is serving as the ACES conference committee chairman. He has concentrated on enhancing the ACES annual conference, publications and on increasing industrial and government activities and participation in ACES. He has also concentrated on increasing the value of membership in ACES by reducing fees, enhancing services and communication with members as well as the conference format. The ACES journal has become listed on the ISI citation index. Dr. Mohammed has already introduced several changes to the ACES conference format, location and its publicity as well as member communications and services through web site. He has further organized sessions and presented papers at many of the ACES conferences. Dr. Mohammed has been a long time active member of ACES and its journal editorial board as well as a recipient of ACES service award. He is the general chair for the 2006 ACES annual conference in Miami, Florida.

CANDIDATE's PLATFORM

As the current president of ACES, I have developed a vision statement that simply say *.ACES will strive to be the main venue for the EM community that is best suited for theorists, experimentalists and practitioners.*. During the past two decades, computer modeling and numerical methods have matured as problem-solving tools in real-world electromagnetics applications. The interdisciplinary scope of ACES is pivotal and should be directed towards maintaining "cross-fertilization" between the high-frequency and low-frequency applications as well as to experimentation and practice. In addition to the services that ACES now offer its members, we must introduce modifications that would enhance member services and improve attendance and format of the ACES annual conference. The ACES Journal should become widely distributed, referenced and utilized by the CEM community worldwide.

In times of accelerating technological and social shifts, ACES services should be modified to deal with these changes. Electronic dissemination of information is changing our interface with the society members and is changing the way we will distribute our products and services. This represents an opportunity for us to increase the value of ACES membership in general as well as to organizational groups such as industry and government institutions in particular. We need to understand what member value is today and what it should be in the future to ensure that we continue to serve industry and the CEM community. With more of our information coming from the Internet, we need to offer a full range of electronic products and services. ACES needs to be the first source that its members to go for information. We must help the member identify the right information and then provide appropriate access to it. We also need to improve our focus on industry and government agencies and increase the relevancy of ACES to these organizations. This can take many forms such as publications and products that give practical information such as standards. We could also focus on professional networking by facilitating communications among practitioners and experimentalists in specialized technical areas in the applied electromagnetics field. ACES also need to increase its continuing education focus. For industrial members and their managers, continuing education is very important to keep up to date and may fill a need of acquiring professional development hours (PDH) required to maintaining licensure as well as it could be a source of income for the society.

The overall future of ACES is very bright and we can be valuable in our members. ACES can be extremely relevant if we work hard and execute the above items in our vision. The key for us is to utilize the opportunities inherent in the dramatic changes in our world and our technology. Relevance to our members with products and services will increase membership and enhance attendance at the annual symposium. As ACES president and BoD member I will continue to promote ACES on all fronts; work towards tasks outlined in our vision and increase communication and cooperation with other organizations.

DR. NATALIA K. NIKOLOVA

GENERAL BACKGROUND

Natalia K. Nikolova received the Dipl. Eng. degree (in Radio-electronics and Automation) from the Technical University of Varna, Bulgaria, in 1989, and the Ph.D. degree (in Computational Electromagnetics) from the University of Electro-Communications, Tokyo, Japan, in 1997. From 1998 to 1999, she held a Postdoctoral Fellowship of the Natural Sciences and Engineering Research Council of Canada (NSERC), during which time she was initially with the Microwave and Electromagnetics Laboratory, DalTech, Dalhousie University, Halifax, Canada, and, later, for a year, with the Simulation Optimization Systems Research Laboratory, McMaster University, Hamilton, ON, Canada. In July 1999, she joined the Department of Electrical and Computer Engineering, McMaster University, where she is currently an Associate Professor. Her research interests include theoretical and computational electromagnetism, high-frequency analysis, as well as CAD methods for high-frequency structures and antennas. She has published more than 45 journal papers in technical and physics journals, and has contributed to more than 40 refereed conferences in the field of microwave and antenna engineering, theoretical and numerical electromagnetism, etc. She has contributed chapters to two books, and was one of the principal developers of the multimedia CDROM accompanying the 7th edition of the textbook of Hayt, Jr., and Buck, *Engineering Electromagnetics*. Currently, she works on response sensitivity analyses with numerical electromagnetic solvers and their integration into automated design, on time-domain numerical analysis based on the electromagnetic potentials, on the problem of nonradiating electromagnetic sources and the nonuniqueness of the solution to inverse problems. She is a member of the IEEE (the Microwave Theory and Techniques Society, the Antennas and Propagation Society), the Applied Computational Electromagnetics Society (ACES), the International Union of Radio Science (URSI) – member of the Canadian National Committee of URSI representing Commission D, and the Professional Engineers of Ontario Association.

Dr. Nikolova currently holds a University Faculty Award of NSERC, awarded in 2000 and renewed in 2003.

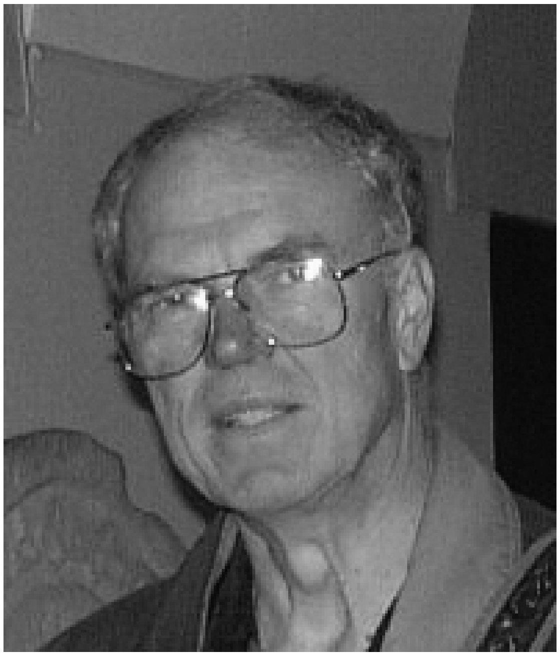
PAST SERVICES TO ACES

The candidate has participated in the ACES annual conferences every year since 2002, and has been a session chair in 2004.

CANDIDATE'S PLATFORM

For more than ten years now, computational electromagnetics (EM) has been in the core of my job and my research interests. The subject is fascinating since it combines the depth of theoretical knowledge with the art of building computing algorithms. The advent of powerful computers and computational EM has brought tremendous changes to the way we design high-frequency hardware, even to the way we teach our students electromagnetics, microwaves and antennas. Contrary to the opinion of many, I believe that this field of study has not matured yet and there is still a long way to acceptable quality and efficiency. One example is my current involvement with design sensitivity analysis – a subject which is barely heard of in our society while structural engineers have already started incorporating it in commercial CAD. We must admit that the efficiency and speed of our analysis algorithms, especially for transient field computations, is still insufficient, making the design of realistic structures infeasible. I would like to see practical actions taken by ACES to promote: theoretical electromagnetic and mathematical research which may have impact on the numerical approaches in computational EM; increased collaboration (joint meetings, etc.) with other computational societies, e.g., fluid mechanics or structural engineering; fair comparative evaluations of the performance of techniques and simulators, which constitute the state-of-the-art in analysis; integration of numerical electromagnetics with design automation and optimization; studies of inverse electromagnetic problems such as target identification, microwave imaging and nondestructive testing; as well as graduate studies in computational EM through scholarships, awards and grants.

DR. JOHN D. NORGARD



GENERAL BACKGROUND

John D. Norgard is a Professor of Electrical & Computer Engineering at the University of Colorado at Colorado Springs, the President and CEO of ElectroMagnetic Technologies (EMT), Inc., and the Chief Scientist of zeeWAVES, Inc. He is also a Distinguished Visiting Professor (DVP) at the US Air Force Academy, where he is the Director of the Computational Electromagnetic (CEM) Division of the High Performance Computer (HPC) Center. He has taught graduate and undergraduate courses in Electromagnetic Field Theory for over 30 years and is the Director of the Electromagnetics Laboratory at the University of Colorado.

Before coming to the University of Colorado, he worked in the Electrical Engineering Department at Georgia Tech and was a Post-Doctoral Fellow at the Norwegian Defense Research Establishment (NDRE) in Kjeller, Norway. He worked at the Jet Propulsion Laboratory (JPL) while studying at Caltech and was a Co-Op student at Georgia Tech while working at the Charleston Naval Shipyard (CNS).

He has worked on numerous computational electromagnetic (CEM) problems, including RF antennas/apertures, strip lines/microstrips, waveguides and transmission lines, propagation of waves through various plasma media (polar ionosphere, rocket soundings), interaction and coupling of waves to wires (cross-talk, NEMP, and lightning), EMI, EMC, EMS/V, backscatter from clutter targets, ESD, and IR metrology. He has developed a 2D thermal mapping technique using infrared thermography and microwave holography to measure electromagnetic fields and to validate numerical codes. He has been a Visiting Professor at the Tel-Aviv University and was a member of the technical staff of the Bell Telephone Laboratories.

He is a Fellow of the IEEE for IR measurements of EM fields, on the Board of Directors for the IEEE/ EMC Society serving as the Vice President for Technical Services, on the Board of Physics and Astronomy for the National Academy of Sciences, Past Chairman for Commission A/Metrology of URSI, and an Associate Editor for the IEEE/EMC Transactions in the area of antenna metrology. He has authored over 175 technical papers, reports, and journal articles.

DR. C. J. REDDY

GENERAL BACKGROUND

C. J. Reddy received B.Tech. degree in Electronics and Communications Engineering from Regional Engineering College, Warangal, India in 1983. He received his M.Tech. degree in Microwave and Optical Communication Engineering and Ph.D. degree in Electrical Engineering, both from Indian Institute of Technology, Kharagpur, India, in 1986 and 1988 respectively. From 1987 to 1991, he worked as a Scientific Officer at SAMEER (India) and participated in radar system design and development. In 1991, he was awarded NSERC Visiting Fellowship to conduct research at Communications Research Center, Ottawa, Canada. Later in 1993, he was awarded a National Research Council (USA)'s Research Associateship to conduct research in computational electromagnetics at NASA Langley Research Center, Hampton, Virginia. Dr. Reddy worked as a Research Professor at Hampton University from 1995 to 2000, while conducting research at NASA Langley Research Center. During this time, he developed various FEM codes for electromagnetics. He also worked on design and simulation of antennas for automobiles and aircraft structures. Particularly development of his hybrid Finite Element Method/Method of Moments/Geometrical Theory of Diffraction code for cavity backed aperture antenna analysis received Certificate of Recognition from NASA.

Currently, Dr. Reddy is the President and Chief Technical Officer of Applied EM Inc, a small company specializing in computational electromagnetics, antenna design and development. At Applied EM, Dr. Reddy successfully led many Small Business Innovative Research (SBIR) projects from the US Department of Defense (DoD). Some of the technologies developed under these projects are being considered for transition to the DoD.

Dr. Reddy also serves as the President of EM Software & Systems (USA) Inc. At EMSS (USA), he is leading the marketing and support of commercial 3D electromagnetic software, FEKO in the US, Canada, Mexico and Central America.

Dr. Reddy is a reviewer for international publications such as *IEEE Trans. Antennas and Propagation (USA), Electronics Letters* (London), *IEE Procedings-H* (London), *IEEE Antennas and Propagation Magazine* (USA), *IEEE Transaction on Antennas and Propagation* (USA) and *IEEE Microwave and Guided Wave Letters* (USA). He is also a contributing author for the book *Comprehensive Dictionary of Electrical Engineering* published by CRC Press Inc. (USA). Dr. Reddy is a Senior Member of Institute of Electrical and Electronics Engineers (IEEE), USA. He is also a member of Applied

Computational Electromagetic Society. He has published more than 60 referred journal articles and conference papers so far.

PAST AND CURRENT SERVICES TO ACES

- Attended ACES conferences from 2003
- Organized and co-chaired special sessions at ACES Conferences in 2004 and 2005.
- Currently serving as Publicity Chair for ACES Conference in 2006.

CANDIDATES PLATFORM

In the recent past, the field of computational electromagnetics (CEM) matured enough to see proliferation of commercial electromagnetic codes in the market. As new techniques and methods are being worked out in the academia, there is an active use of these commercial codes for design and development of antennas, placement of antennas on large platforms, analysis of EMI/EMC and many other applications. I believe, it will be of benefit to ACES membership to learn from the experiences of the users of these commercial codes. I would like to see ACES intensify efforts to see validation of the commercial codes for various electromagnetic problems. It will also benefit the ACES membership to learn about various numerical techniques used in different commercial codes and their strengths and weaknesses. I believe, attracting papers that validate commercial codes and discuss their applicability to specific electromagnetic problems will strengthen ACES membership. I am looking forward to contributing to the growth of ACES in the coming years.

DR. TAPAN K. SARKAR

GENERAL BACKGROUND

Tapan K. Sarkar received the B.Tech. degree from the Indian Institute of Technology, Kharagpur, in 1969, the M.Sc.E. degree from the University of New Brunswick, Fredericton, NB, Canada, in 1971, and the M.S. and Ph.D. degrees from Syracuse University, Syracuse, NY, in 1975. From 1975 to 1976, he was with the TACO Division of the General Instruments Corporation. He was with the Rochester Institute of Technology, Rochester, NY, from 1976 to 1985. He was a Research Fellow at the Gordon McKay Laboratory, Harvard University, Cambridge, MA, from 1977 to 1978. He is now a Professor in the Department of Electrical and Computer Engineering, Syracuse University. His current research interests deal with numerical solutions of operator equations arising in electromagnetics and signal processing with application to system design. He obtained one of the "best solution" awards in May 1977 at the Rome Air Development Center (RADC) Spectral Estimation Workshop. He has authored or coauthored more than 280 journal articles and numerous conference papers and 32 chapters in books and fifteen books, including his most recent ones, Iterative and Self Adaptive Finite-Elements in Electromagnetic Modeling (Boston, MA: Artech House, 1998), Wavelet Applications in Electromagnetics and Signal Processing (Boston, MA: Artech House, 2002) , Smart Antennas (John Wiley & Sons, 2003) and History of Wirless (John Wiley & Sons, 2005 – to be published).

Dr. Sarkar is a Registered Professional Engineer in the State of New York. He received the Best Paper Award of the IEEE Transactions on Electromagnetic Compatibility in 1979 and in the 1997 National Radar Conference. He received the College of Engineering Research Award in 1996 and the Chancellor's Citation for Excellence in Research in 1998 at Syracuse University. He was an Associate Editor for feature articles of the IEEE Antennas and Propagation Society Newsletter (1986-1988). He was the Chairman of the Inter-commission Working Group of International URSI on Time Domain Metrology (1990–1996). He was a distinguished lecturer for the Antennas and Propagation Society from 2000-2003. He is currently a member of the IEEE Electromagnetics Award board and an associate editor for the IEEE Transactions on Antennas and Propagation. He is the vice president of the Applied Computational Electromagnetics Society (ACES) and was the ACES Liason for the combined IEEE 2005 Wireless Conference along with ACES held in Hawaii. He was the general chair for the ACES 2004 conference held at Syracuse. He is on the editorial board of Journal of Electromagnetic Waves and Applications and Microwave and Optical Technology Letters. He is a member of Sigma Xi and International Union of Radio Science Commissions A and B.

He received Docteur Honoris Causa both from Universite Blaise Pascal, Clermont Ferrand, France in 1998 and from Politechnic University of Madrid, Madrid, Spain in 2004. He received the medal of the friend of the city of Clermont Ferrand, France, in 2000.

PAST AND CURRENT SERVICE TO ACES

Tapan has contributed to both invited and survey papers to the ACES conference, journal, on the board of directors (2000-2003, 2003-2006) and has actively supported exhibit booth at the ACES conferences. In addition, sessions were also organized in the last two ACES conferences and short courses were presented by him. Tapan is currently the vice-president of the ACES society, was the general chair of the ACES conference at Syracuse in 2004 and was associated with the conference in Hawaii this year. The meeting at Syracuse was the first one outside the Monterrey area in twenty years and did result in a modest surplus for the society.

CANDIDATE'S PLATFORM

The strength of ACES lies in the following areas:

1. Providing a more pragmatic and user-oriented approach to computational electromagnetics thereby providing a strong coupling between the code users and the code developers.
2. Providing focused articles, which are more practical and more meaningful to the users.
3. Providing an open forum with an extended summary of the presentations, which makes it very convenient for a reader to understand what the speaker talked about in the conference long after the conference, is over.

Therefore to continue along the niche areas of strengths of ACES, my participation in the past had been to organize technical sessions along these themes, particularly on use of electromagnetic simulation codes, from the users perspective. An attempt was made to delineate pros and cons of some of the commonly used codes so that light can be generated without heat. If selected again, I shall continue my efforts in this direction and might increase the technical scope in generating tutorial articles on this topic from experienced scientists belonging to this community. In addition, I am also one of the members of ADCOM for the IEEE Antennas and Propagation (AP) Society. This may result in a stronger interaction between AP which is quite theoretical with ACES which is more user oriented, in the future. The interaction may result in having topical AP meetings perhaps in conjunction with ACES.

In summary, the strength of ACES lies in building a stronger connection between the code developers and the users and that should be further enhanced. This is what I propose to strengthen further.

Also as a vice chair of the society, my goal is to strengthen the linkage between the Antenna sand Propagation Society, which focuses primarily on theory and AMTA, which is more experimentally oriented and finally, ACES which is the perfect forum for the user community.

OTHER UNIQUE QUALIFICATIONS

I believe I can create a stronger link between the CEM theoreticians, experimenters and the users by making an attempt to reorient the theoreticians and experimentalists into looking at the solution of a CEM application from a practical pragmatic standpoint. In addition valuable information can be gained from such interactions between researchers of such diverse background as we have tried to follow the same philosophy during the last decades by not only developing computer codes but also making them user friendly for researchers who are not familiar with them. In future, we need to perhaps bring in the AP and the AMTA community to benefit the members of ACES in having an interactive perspective on applied computational electromagnetics.

CEM NEWS FROM EUROPE May 2005-05-31

Tony Brown (Anthony.brown@manchester.ac.uk)

1)ACES-UK

It is not without some sadness I have to report that the ACES UK Chapter held an Extra-ordinary General Meeting on 5th April 2005. By a majority of nearly 3:1 it was agreed to wind up the organisation as a independent Chapter of ACES. The principal reason was lack of support from the UK CEM community. Despite efforts by a hard working committee, the community of CEM users in the UK have not, for whatever reason, felt the organisation useful and given the necessary support. Attendance over the years at the annual meeting has been difficult to maintain. Last year's event had to be cancelled due to lack of support. Clearly the considerable organisational effort needed to keep the Chapter running was not serving a useful need.

ACES UK had its foundation in the NEC Users Group founded by this correspondent and Pat Foster in 1985.

It was the strong wish of those individuals present to continue to promote CEM in the UK on a less formal basis; this could include occasional workshops or meetings, possibly using the IEE or IEEE organisations. There will be a final meeting Tuesday 28th June at which this will be discussed. All current ACES UK members will administer their membership directly with ACES US from the end of 2005.

2) NEWS FROM EUROPE

I would like to invite contributions to this column. I am particularly keen to hear from organisations and individuals whom it would be appropriate to 'spotlight' in this column. I would also like to include notifications of technical meetings of interest, etc.

3) MEETINGS OF INTEREST

A selection of European forthcoming meetings:

IEEE International Conference on Ultra-Wideband (ICU 2005)
5-8 September 2005, Zurich, Switzerland (www.icu2005.ee.ethz.ch)

IEE Pulsed Power Symposium 2005
8th September 2005, Apollo Hotel, Basingstoke, UK (www.iee.org)

EMCUK 2005
11-12 October 2005, Newbury Racecourse, UK (www.emcuk.info)

European Microwave Week
3-7 October 2005, Paris (www.eumw2005.com)

10th AMPERE International Conference on Microwave and High Frequency Heating
12 – 15 September 2005, Modena and Reggio Emilia University, Italy (http://www.ampereeurope.org)

Perspectives on Computational Electromagnetics

It can be hugely productive to look at one subject as it interfaces with another. This is particularly important when one considers the benefit of knowledge from one sphere being translated to another. *Perspectives on CEM* aims to do this. Its objective is to try to strengthen the practice of CEM by looking at it from the angles other than mathematical or methodological. The technical notes in this newsletter, plus the papers in various journals and conferences are better able to do this. This section will be a success if it helps us to look at, and think about, CEM in different ways. Over the next few issues we will attempt to do this through a mix of essays and perspectives. The anticipation is that these perspectives will provoke some thought, opinion and debate. Some of the intended topics to be covered are:

- Does a knowledge/study of history help us to build better codes and better models now?
- What has philosophy got to do with CEM?
- Organisational culture: what shapes it and what are the benefits of managing it?
- Is knowledge the most valuable resource?
- Should the teaching of electromagnetics be biased towards understanding the mathematics or being able to drive the packages really well?
- What would it take for CEM to warrant the level of governmental research support as some of the more fashionable subjects, such as nanotechnology?
- How can we predict technological development?
- IPR: when is it a good idea not to bother?
- What makes a brand?
- How can project teams be made more effective?

Please feel free to make comments or observations about the intended and actual content.

I hope that this section will provide some food for thought over the next few issues

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Where Intangibles Collide – Stakeholders and Software projects

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There is a school of thought that says software systems do not work properly until they have failed repeatedly in action. There are various methodologies for software development, some very academic, some less so. In all cases, a limiting factor on whether the software will do what it needs to do is how well stakeholders have been included in the project. This article addresses some of the stakeholder factors pertinent to software-heavy (computational electromagnetics) projects. The importance of identifying and addressing the needs of stakeholders can be clearly seen in a wider context by looking at Concorde and the Brent Spa oil rig [1] although neither were computational electromagnetics projects, they do illustrate a point. The Concorde design is perhaps best described as iconic, but as a project it was not quite as successful because although it was designed for transatlantic flights, the USA had virtually no involvement an as a result it took quite some time for the routes to be opened up. Similarly, if the disposal of the Brent Spa oil rig had involved negotiations with environmental groups from the outset, substantial amounts of money could have been saved, not to mention avoiding a public relations disaster.

From a software angle, "In the real world, stakeholders are interested in solutions, not some abstract developer-centric set of requirements" [2]. So it is important to identify who the stakeholders are, what their interaction is with the project, what they want to get out of it and what the conflicts are between stakeholder requirements. This article will look at defining and identifying stakeholders and then how the development process can benefit from this.

The benefits of identifying stakeholders are both direct and indirect. Direct benefits include ensuring that all the groups who should be involved in the project have their contributions and expectations clearly built into the project plan, this includes involving the stakeholder groups when required and identifying conflicting stakeholder requirements that need managing. Indirect benefits include the resultant awareness of possible weak signals that, as a project team member, being sensitised to "stakeholders" can produce. Weak signals are deviations in the noise floor that can signify a growing signal, spot it early enough and you can deal with it, leave it too long and it can become a problem: the problem with the noise floor is that you need to be looking at it carefully and you need to adopt an appropriate filter. However, by cycling through stakeholder requirements and negotiating conflicts, many of these issues can be minimised.

Stakeholders

There are probably as many definitions of stakeholders as there are people discussing it. Even having decided on a definition, it can be difficult, if not impossible, to account for all stakeholder relationships, but as noted above, being sensitised can be a benefit in itself. Some approaches to stakeholder identification include lists of possible involvements. Such as in [3] where the author cites a World Bank view of stakeholders including those:

- Who might be affected by the development concern to be addressed
- Whose behaviour has to change for the effort to succeed?

We find these very helpful but prefer a simpler push-pull view of stakeholders: a stakeholder is someone (or a group) who contributes something to the project in order to get something out of it. This loose definition has the disadvantage that it requires more than a tick list to identify the stakeholders but has the advantages that virtually all stakeholders can be identified this way by thinking creatively about the push and pull. Internally, stakeholders will include the programmers (contributing their effort and giving up the opportunity to work on other projects in order to gain remuneration, job satisfaction and intellectual challenge). Externally, the stakeholders will include (but not be limited to) customers who will contribute money, possible IPR and give up the potential for working with another organisation in order to receive a product that allows them to do the job they want doing.

A helpful tool in identifying stakeholders is to group them [3] into primary, secondary, external and extended stakeholders.

- Primary stakeholders – who have authority or resource responsibilities, or who have the power to influence collaboration outcomes. Their involvement is vital otherwise problems may ensue. Naturally, a customer for a solver will be intimately involved in the project planning process.
- Secondary stakeholders – have an indirect interest in the outcome such as with employees or consumers. They should be involved in elements that impact on them but not in other areas. Here, for example, an end-user may be involved in discussions about front-ends but not necessarily data structures.
- External stakeholders – are not part of the project but will have an impact on the project. An example of these may be regulators or grant awarding bodies. They will be expecting something out of the project and need to be involved carefully, possibly by lobbying, letters or discussions.
- Extended stakeholders – are more likely to influence the overall impact of the finished product (a conference audience, perhaps, or the readers of this article (?)). It is difficult to account for this group as easily as the other three groups. Their involvement may apparently be one way: they consume what has been output from the project. Certainly, their input to the project itself will be limited, the danger being that any input they have may not be predictable *a priori*.

The three main categories that software projects fall into are [4]:

- Standard package
- Customised
- Bespoke

In a rather more ordered world than the one we currently live in, these three categories would exist separately. However, for CEM, they can become somewhat intertwined. In the extreme case, a bespoke package will be developed for a customer, based on the customising of a current package that will also be included in a standard product. All the stakeholders need to be considered but possibly with different emphases.

The following table gives a starting point for identifying stakeholders for a project (some suggestions are in place). What it really suggests is that by creatively thinking about those involved, they can be categorised and treated in a way that is relevant to their impact on the project

	Primary	Secondary	External	Extended
Standard		Sales team	Government funding body	
Customised	Customer			Conference attendees
Bespoke	Customer (contract placer)	End-user		Buyer's customers

This approach can be further elaborated by considering the stakeholders as *personas* [5] although we are not going to discuss this further here.

Stakeholder involvement

We now have a potentially simple structure to identify the stakeholders and what they want out of their relationship with the project. In each case the following need to be clearly identified

- What are they contributing or giving up?
- What are they getting back from the project?

- How should they be involved? What is the best way of communicating with them
- Are their any conflicts with other stakeholders?

Naturally, major consideration needs to be given to those with the greatest ability to influence the outcome: the primary or secondary stakeholders. It must be remembered that all stakeholder groups can have an impact on the success or failure of a project, or on whether it comes in late and over budget. This section looks at how this involvement can be effectively utilized.

There is a certain amount of negotiation required to identify the most relevant stakeholders, their push-pull attributes and how they can and should be involved. A useful framework for undertaking this negotiation is the WinWin Spiral [6, 7]. The method is also claimed to enhance trust between stakeholders and to encourage flexibility within the team.

In essence, the WinWin Spiral Model [7] is based on the following activities:

1. Identify the key stakeholders.
2. Identify the stakeholders' win conditions (their goals from the project). If their win conditions are non-controversial, i.e. they do not conflict with other win conditions, they are accepted as they are. Otherwise, an 'issue artefact' is created and this becomes the topic of negotiations to resolve the issue.
3. Negotiate win-win reconciliations.
4. Establish objectives, constraints and alternatives.
5. Evaluate the alternatives and identify and mitigate risks.
6. Define the next level of product
7. Validate the product / process definitions
8. Review and commit
9. return to #1 for the next cycle to deal with the next level of detail.

Several cycles at increasing levels of complexity may be necessary. One example [7] demonstrates a four cycle approach for a multimedia project, although the general thrust can be a helpful starting point:

- Cycle 0 – determine feasibility of general approach.
- Cycle 1 – develop prototypes, plans, etc. and verify the existence of at least one feasible solution.
- Cycle 2 – develop more detailed plans and verify that there are no major risks in satisfying the specifications
- Cycle 3 – develop operational plans including user support plans.

Any other implementation of this may require more or fewer cycles, but consideration of the four cycle approach is a good starting point.

It should be noted that there are other software development methodologies which address similar issues. For example Agile Methods such as Extreme Programming [8], which develop, in short iterations, only a limited number of User Stories with customer involvement on the development team throughout.

However, embedding the identification of stakeholders as discussed earlier into the WinWin Spiral model should help in minimising 'nasty surprises' and maximising the quality of delivery, even if a formal software requirements specification (SRS) approach is being used [e.g. 9]. So, hopefully, repeated failure will not be as important in order to prove that the software works.

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HF Inverted Vee with Dipole Augmentation

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***Abstract*— In the amateur radio community, a common practice is to center-feed a dipole resonant in the 75-meter band (say at 3.9 MHz) with high-impedance ladder line, and use the same antenna on all eight amateur bands between 3.5 and 29.7 MHz with the aid of an antenna tuning unit. The dipole is typically 125-135 feet in length. Some users have employed longer "reflector" wires and/or shorter "director" wires in combination with the basic dipole element in an attempt to enhance the antenna's overall performance. This article reports a variation which has not been seen elsewhere in the published literature, namely, an inverted vee antenna is used as the driven element and a 75-meter half-wave dipole is then used to augment the gain of the inverted vee. The result is an antenna with higher gain across the HF spectrum at elevation angles of interest to practical communicators. The configuration reported here is not claimed to be optimum in any sense. Rather, the objective is to encourage others to apply CEM analysis to additional variations and report progress as it is achieved.**

I. Introduction

This is a specific case study, reported because the results are believed to be of interest and useful to practical radio communicators operating in the HF spectrum. The configuration is an inverted vee antenna as the driven element with a 75-meter half-wave dipole added to augment the gain of the inverted vee. EZNEC 4.0 antenna software by Roy Lewallen, W7EL, was used for the numerical analysis in this study. Real ground with $\sigma = 3$ mS and $\epsilon_r = 12$, typical of West Alabama soil, was assumed for all computer runs.

The (x,y,z) coordinates in feet of the wire ends are as follows: (a) for the dipole, end 1 is at (0,53,41) and end 2 is at (0,-83,57), and (b) for the inverted vee, which is off-center fed, end 1 is at (1,54,41) and end 2 is at (1,-84,57) with the feed point at (-36,0,10). The dipole and inverted vee ends are in close proximity, but not connected. The end elevations are different because they reflect the actual heights used in a prototype antenna, which has been constructed and used regularly for an extended period of time. The inverted vee feed point is off-center because it is the point closest to the transmitter location - only fifteen feet of transmission line are required to connect the transmitter to the antenna. There a 5-inch gap at the center of the dipole, and a section of 600Ω open-wire ladder line extends from the center feed point straight down to a height of 8 feet, where the two conductors are shorted. The transmission line and antenna are both constructed of #14 stranded copper wire. The active transmission line feeding the inverted vee is the same type 600Ω open-wire ladder line. The vee-dipole geometry is shown in Figure 1 below.

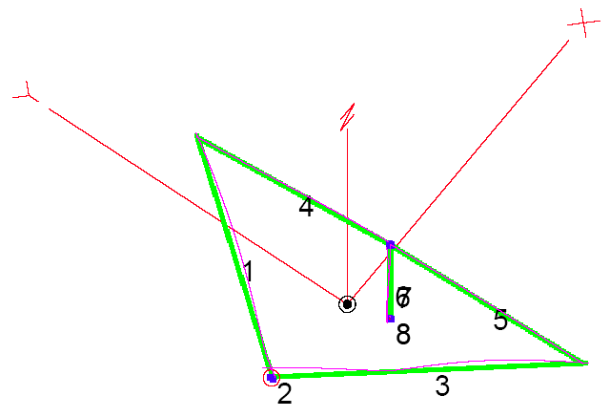

Fig. 1. Inverted vee - dipole configuration.

II. CEM Analysis Results

Note that the +x axis is toward the East, and the +y axis is toward the North. In the elevation plots that follow, for an East-West slice $(\varphi = 0°)$, East is to the right side of the plot, and West is to the left. Similarly, for North-South $(\varphi = 90°)$, North is to the right and South is to the left.

Performance at all eight HF amateur bands is too much data for presentation here. Four frequencies were selected for illustration purposes, two toward the ends of the HF spectrum, and two in between: 3.885, 7.290, 14.286, and 29.0 MHz. Elevation patterns for the dipole alone, which was the antenna previously in use alone, are superimposed on plots for the composite "inverted vee augmented with a dipole" antenna. The results are given successively as Figures 2 through 9 below.

Note that, in each of Figures 2-9, the "primary" trace is the inverted vee - dipole antenna, and the other trace is for the dipole alone at the frequency and azimuth angle specified in the figure caption.

Figures 2-9 occupy the next four pages. Discussion resumes after that.

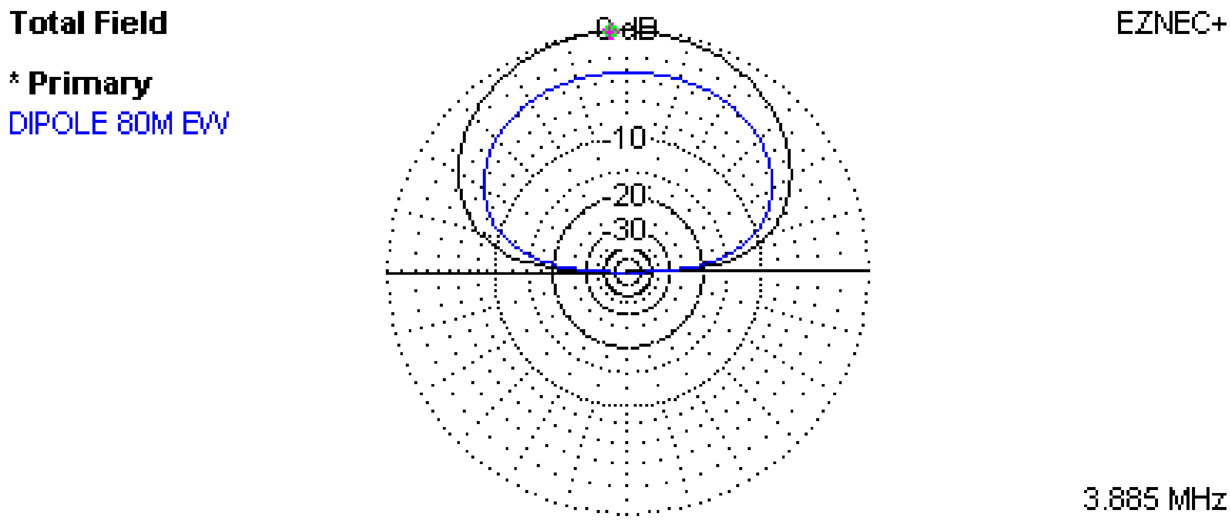

Fig 2. $\varphi = 0°$elevation plot, 3.885 MHz, outer ring 9.39 dBi.

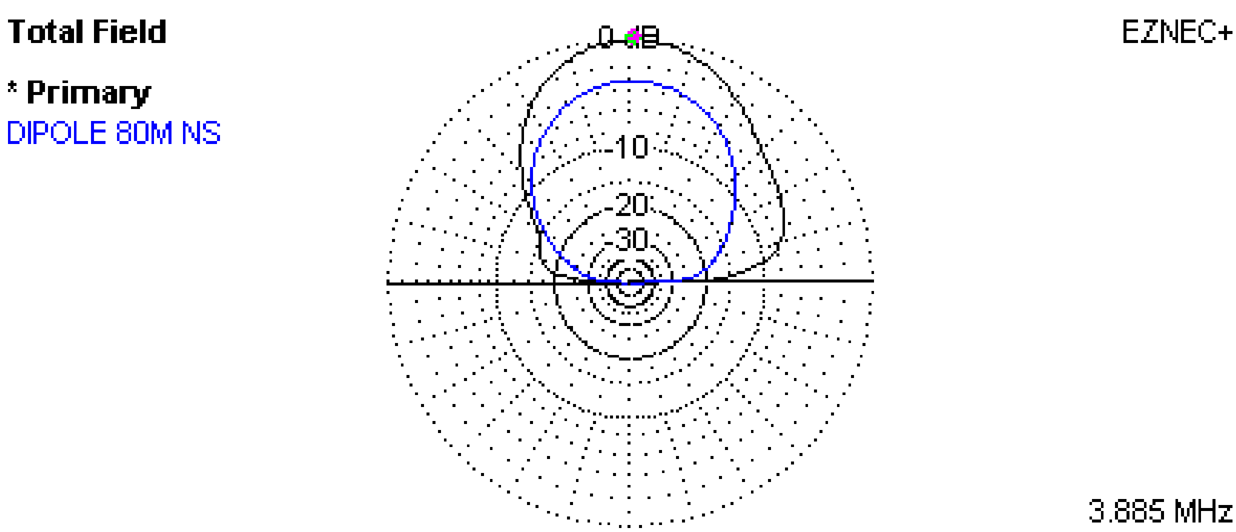

Fig 3. $\varphi = 90°$ elevation plot, 3.885 MHz, outer ring 9.38 dBi.

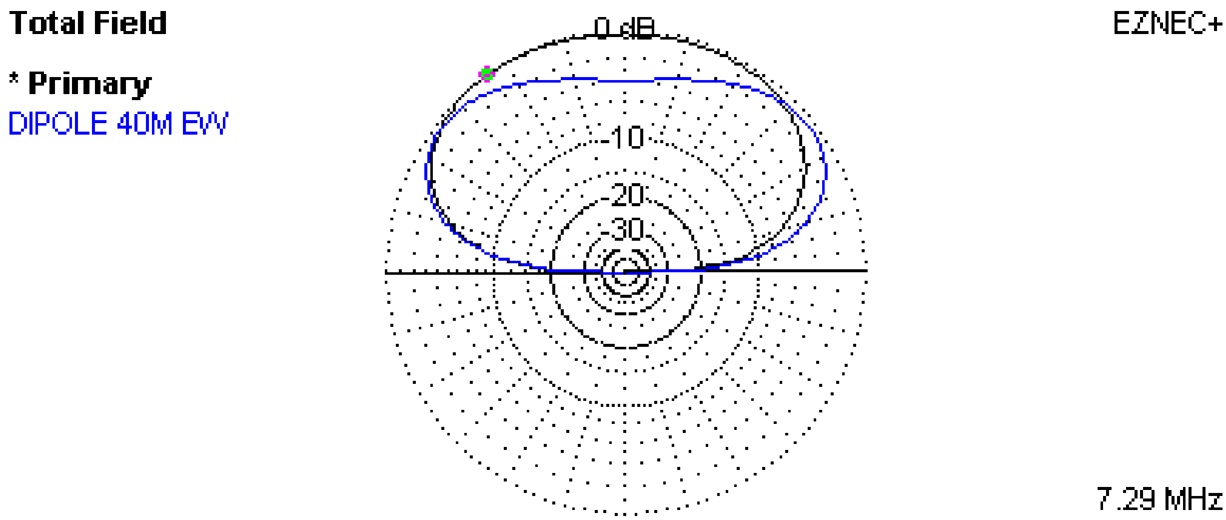

Fig 4. $\varphi = 0°$elevation plot, 7.290 MHz, outer ring 7.91 dBi.

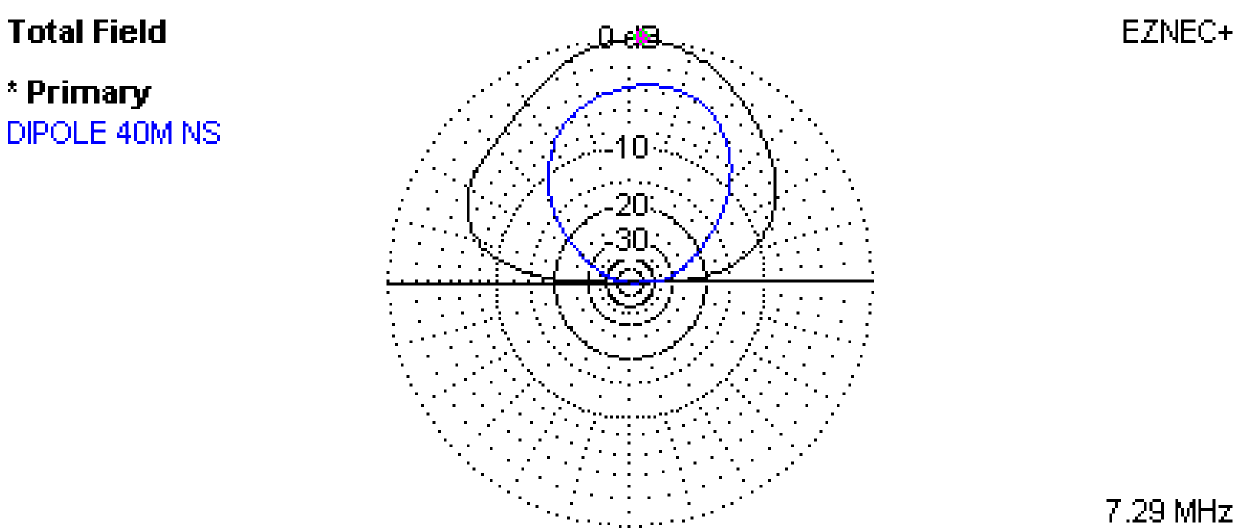

Fig 5. $\varphi = 90°$ elevation plot, 7.290 MHz, outer ring 7.7 dBi.

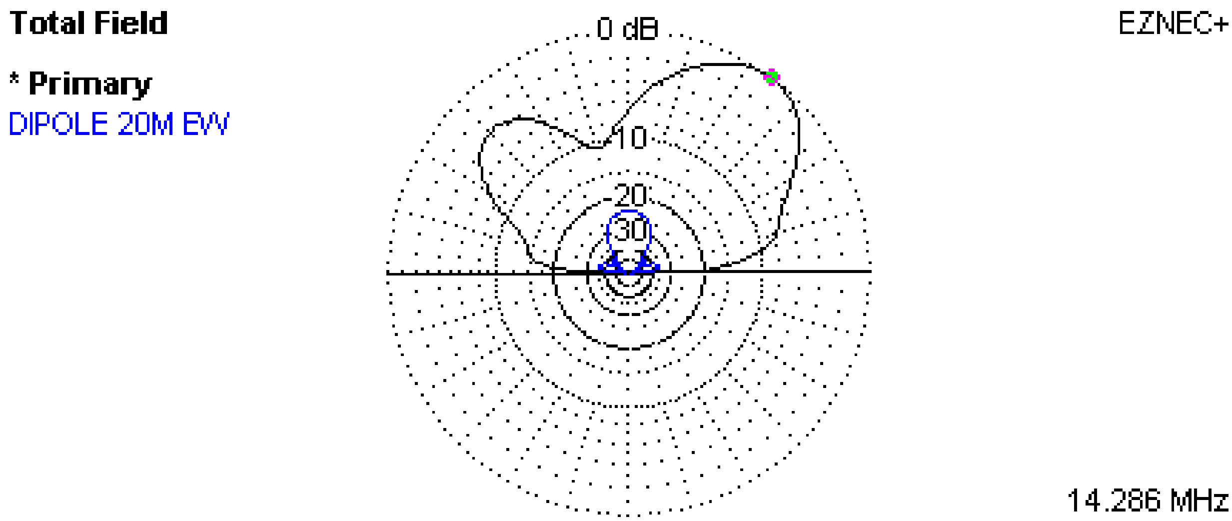

Fig 6. $\varphi = 0°$elevation plot, 14.286 MHz, outer ring 9.8 dBi.

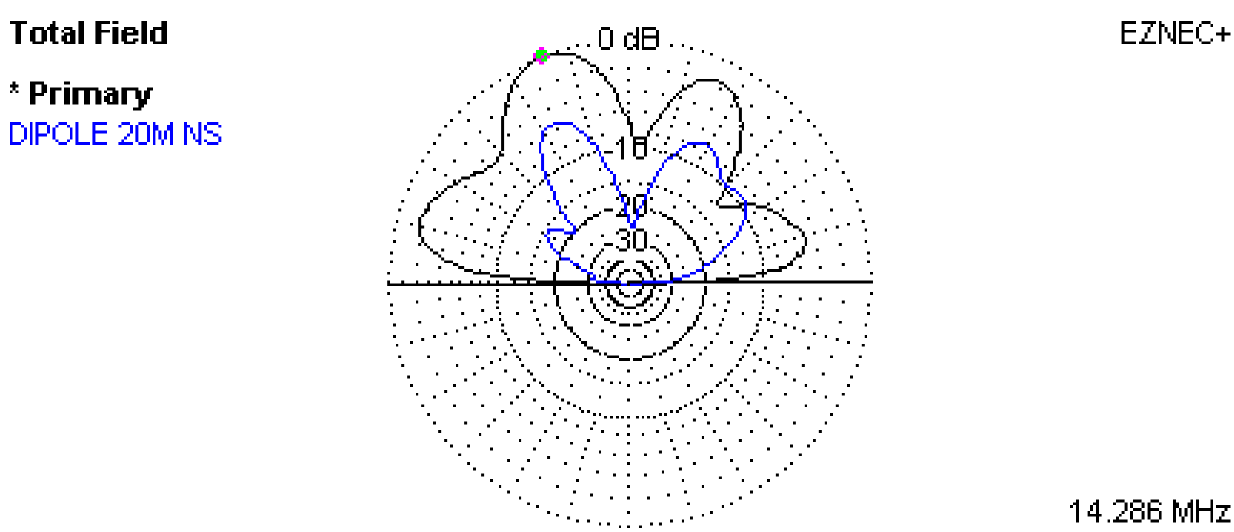

Fig 7. $\varphi = 90°$ elevation plot, 14.286 MHz, outer ring 10.56 dBi.

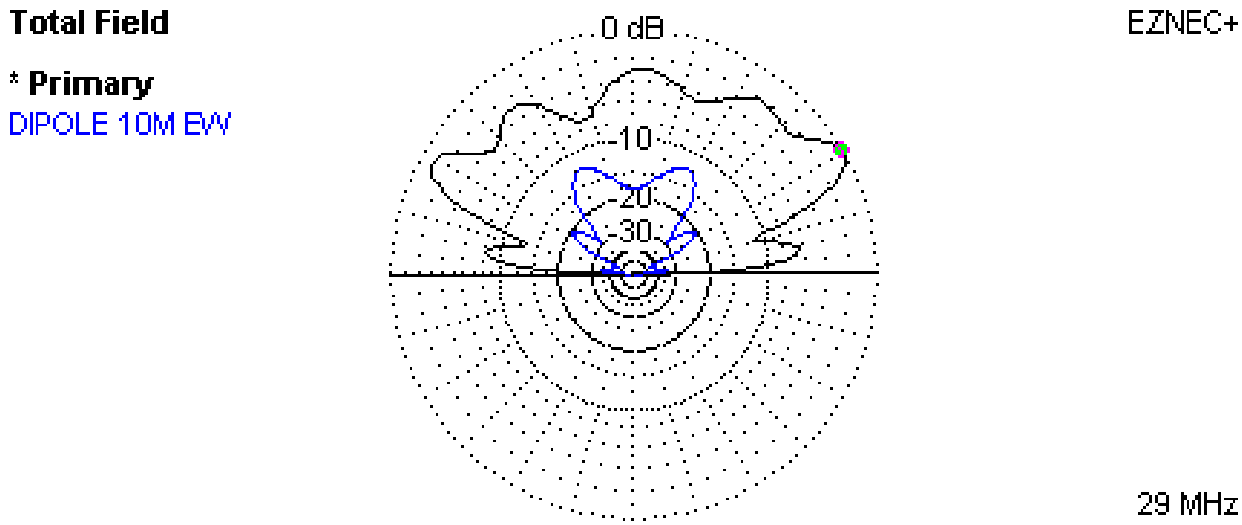

Fig 8. $\varphi = 0°$elevation plot, 29.0 MHz, outer ring 8.91 dBi.

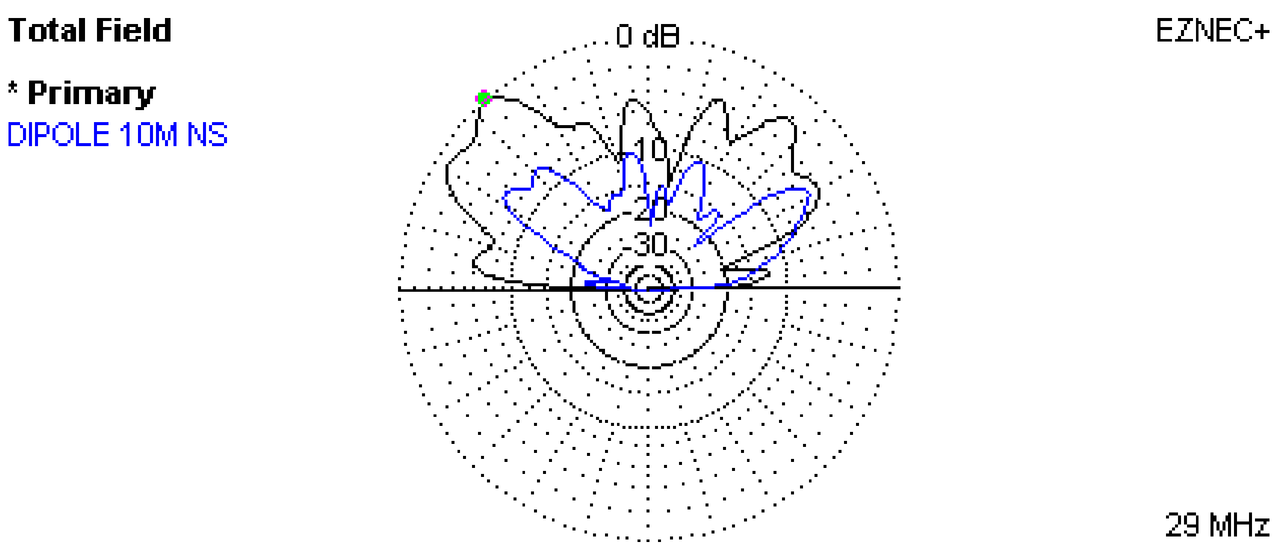

Fig 9. $\varphi = 90°$ elevation plot, 29.0 MHz, outer ring 9.38 dBi.

III. Radiation Pattern Remarks

The preceding figures (2-9) give four East-West and four North-South elevation plots comparing the inverted vee - dipole combination versus the dipole alone (before the vee was added and made the driven element) at the four frequencies 3.885, 7.290, 14.286, and 29.0 MHz.

Three general comments apply. First, because the southern end point is almost twenty feet higher than the northern end point, the antenna configuration has a slight "sloper" behavior favoring radiation toward the North. For example, from Fig. 5 the dipole is almost 6 dB stronger toward the North at elevation angle $45°$. Second, the inverted vee - dipole combination geometry is not symmetrical North-South or East-West, and thus one would expect asymmetric patterns with the effect more pronounced at the higher frequencies. Third, the feed point for the inverted vee is notably close to ground at a height of only ten feet, and the presence of real ground was taken into account throughout the CEM analysis of this antenna. Again, a low feed point was selected so that the transmission line feeding the antenna could be made very short.

At 3.885 MHz, the antenna elevation in fractional wavelengths is small, so its gain is highest straight up. However, for practical communications purposes, there is still quite adequate radiation at elevation angles down to $25° - 30°$. The vee-dipole combination is superior to the alternative of using the dipole alone at all angles, and shows a particularly useful increase for long-distance communication links at elevation angle $15°$ toward the North.

The $\varphi = 0°$ plot of Fig. 4 for 7.290 MHz shows one of the rare instances where the dipole alone would be superior, and this occurs at elevation angles less than $60°$. So, if the user's principal interest was high-angle radiation for links of shorter distance, the vee-dipole has the advantage.

Results at 14.286 MHz require some clarifying comments. The dipole is 2λ at this frequency, and so its azimuth pattern (at, say, $20°$ elevation) exhibits the well-known four-leave clover appearance with lobes toward the Northeast, Northwest, Southwest, and Southeast. The lobes all peak right at 9.0 dBi. For the given geometry and taking into account real ground, the nulls in the $\varphi = 90°$ (North-South) plane are moderately deep, and the $\varphi = 0°$ (East-West) plane are deeper. N-S and E-W slices were selected because they are of greatest interest in the actual deployment of this antenna, and limited space here precludes the presentation of additional patterns. Therefore, Figs. 6 and 7 perhaps imply an unfair advantage for the vee-dipole combination but, nonetheless, the vee-dipole variation is indeed generally superior.

At 29.0 MHz, both antennas are both electrically large and significantly elevated. In various azimuth plots, the dipole exhibits radiation characteristics that change considerably with elevation angle. At $30°$ elevation, the dipole pattern has two major lobes roughly North-South and a pair of sidelobes positioned about $20°$ North and South of the x-axis $(\varphi = 0°)$. Toward the North at elevation angle $30°$, the dipole alone is competitive with the composite antenna but, otherwise, the vee-dipole combination is still clearly superior in overall radiation performance.

IV. VSWR

A plot of VSWR at the inverted-vee feed point from 3.5 to 29.5 MHz is given as Figure 10 below.

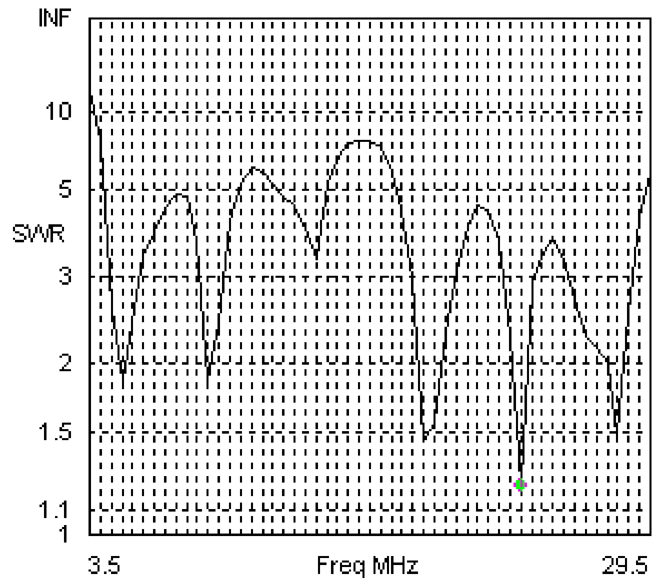

Fig. 10. Vee-dipole combination VSWR.

The notable feature is that SWR is below 10:1 over almost all the HF spectrum, only exceeding that value at the low-frequency end. Further, the SWR is below 5:1 from approximately 17.75 to 27.5 MHz. At HF, a 10:1 SWR is quite acceptable if open-wire feeders are used in conjunction with a high-quality antenna tuning unit. Open-wire transmission line lengths up to 200 feet have been found acceptable, and losses for the fifteen foot length of this study are insignificant.

V. Concluding Remarks

The inverted vee - dipole antenna described here has been found by CEM modeling to be superior to the classical dipole multiband antenna across the HF spectrum. When the rf feed is moved from the old dipole to the vee, the dipole open-wire transmission line need not be removed but simply truncated and shorted at about eight feet above ground, keeping it available for reactivation if the future, if so desired.

Adding the vee is convenient, as the original dipole end supports/anchors can be also be used for the vee ends, and the vee feed point can be located in close proximity to the transmitter with need for only a modest ten-foot high support. For this study, a fiberglas WonderPole^{TM} push-up mast was used to support the vee feed point.

It is assumed that the user will use 450Ω or 600Ω open-wire feeders, and use of coaxial cable with a balun is not recommended.

This is a specific case study, and is intended to stimulate CEM modeling of variations by practical radio communicators who continue to favor the HF spectrum for their operations.

Univariate Rational Macromodeling of High-Speed Passive Components: a Comparative Study

D. Deschrijver, T. Dhaene

Accurate broadband macromodels are of paramount importance for the study, design and optimization of RF, microwave, and millimeter-wave components and systems. These compact macromodels approximate the complex electro-magnetic (EM) behavior of high-speed multi-port systems at the input and output ports in the frequency domain by rational functions. It is well-known that the estimation of the system parameters is a numerically ill-conditioned problem. In literature, several techniques are proposed to relieve these numerical issues. This paper gives an overview and comparison of some rational fitting methods, which are most commonly used to model deterministic, simulation based data.

SECTION 1
INTRODUCTION : MACROMODELING

Compact rational macromodels, based on measurements or full-wave electromagnetic (EM) simulations, are very important for efficient time domain and frequency domain simulation of high-frequency/high-speed interconnections, components and systems. These macromodels characterize the electromagnetic behavior of electrical and electronical components at the input and output ports in the frequency domain $(s, H(s))$, using rational functions [24].

$$R(s) = \frac{N(s)}{D(s)} = \frac{\sum_{n=0}^{N} N_n s^n}{\sum_{d=0}^{D} D_d s^d} \qquad s = j2\pi f \tag{1.1}$$

The rational model should have a sufficient accuracy δ over a predefined frequency range of interest $[s_0, s_K]$

$$dB(|R(s) - H(s)|) < \delta, \ \forall s \in [s_0, s_K] \tag{1.2}$$

and it should satisfy several physical properties, which are inherent to linear time-invariant (LTI) systems, such as e.g. causality, system stability and passivity [1, 2].

(a) The coefficients of the numerator and denominator polynomial must be real, in order to avoid imaginary terms in the time-domain. This implies that the poles and zeros of the rational model are real or occur in complex conjugate pairs (i.e. $R(s) = R^*(s^*)$).

(b) The order of numerator must be smaller than (or equal to) the order of denominator in the case of scattering parameters. If $N > D$,

$$\lim_{s\to\infty} \frac{\sum_{n=0}^{N} N_n s^n}{\sum_{d=0}^{D} D_d s^d} = \infty \tag{1.3}$$

which violates the passivity constraint (e). Also, when considering the partial fraction expansion, the higher order polynomial terms s, s^2, ... would translate to unrealizable derivatives in the time domain.

(c) To enforce Bounded-Input-Bounded-Output (BIBO) system stability, all poles must be located in the left half of the complex plane. Unstable poles translate to unbounded exponentials in the time domain as s approaches infinity.

(d) All rational models (in the case of multiport systems) should share a common set of poles, in order to increase the efficiency of the time-domain convolutions.

(e) If the system is passive (unable to generate energy), the rational model must be passive as well, in order to avoid instabilities in time-domain simulations [12]. For scattering parameters, the rational function must be bounded real

$$I - H(s^*)H(s) \geq 0, \ \forall s \tag{1.4}$$

which is equivalent to

$$\max_{i,s}(\sigma_i(s)) \leq 1, \ \forall \sigma_i(s) \in \sigma(H(s)) \tag{1.5}$$

where σ represents the singular values, or positive real for hybrid parameters

$$\max_{i,s}(\lambda_i(s)) \geq 0, \ \forall \lambda_i(s) \in \lambda(\Re e(H(s))) \tag{1.6}$$

where λ represents the eigenvalues. Recall that eigenvalues and singular values of a matrix H are related by

$$\sigma_i \geq 0 \in \sigma(H) \Longleftrightarrow \sigma_i^2 \in \lambda(H^* H) \tag{1.7}$$

Unfortunately, in some cases, accurate simulation of complex (multi-port) LTI systems can be computationally very expensive and resource demanding, which is the case for full-wave EM simulations. One often wants to minimize the number of costly data samples, in order to find an accurate broadband model in an acceptable amount of time. Therefore, adaptive techniques [5, 8] are used which gradually build up a model by iteratively selecting new samples (and/or frequency derivatives) based on error estimates, while adjusting the model complexity as needed. In this report, the data that is used to characterize these models, is deterministic simulated data (i.e. repeatable and no measurement noise).

In literature, several rational interpolation and rational approximation techniques are proposed to calculate such a rational function. In this report, the most commonly used techniques are selected and compared from a practical point of view. Each approach will be discussed briefly, and some pro's and con's will be given. Afterwards, some numerical results will be given to compare the accuracy and usefullness of all techniques. Matlab code is available at : http://www.coms.ua.ac.be/urm_toolbox, which should encourage the reader to experiment and get some hands-on experience.

SECTION 2
RATIONAL MODEL

A rational analytic model $R(s)$ is defined as a quotient of two polynomials $N(s)$ and $D(s)$.

$$R(s) = \frac{N(s)}{D(s)} = \frac{\sum_{n=0}^{N} N_n s^n}{\sum_{d=0}^{D} D_d s^d} \qquad s = j2\pi f \tag{2.1}$$

where N and D represent the order of numerator and denominator respectively, and N_n and D_d the polynomial coefficients. The rational function provides an approximation of the spectral response of the system over the interval $[f_{min}, f_{max}]$. Since there are $N + D + 1$ unknown coefficients (D_0 can be chosen arbitrarily, e.g. $D_0 = 1$), a set of $K + 1 = N + D + 1$ samples $(s_k, H(s_k))$ is required to identify $R(s)$ completely. $R(s)$ is then an interpolating curve passing through the values $H(s_k)$ at the complex frequencies s_k, for $k = 0, 1, ..., K$. In some cases, it's possible to obtain tth order frequency derivatives $(s_k, H^{(t)}(s_k))$ from the simulator. Frequency derivatives are scaled moments (coefficients of the Taylor series at a given expansion point), which can often be simulated at a significantly lower computational cost than data samples. Taking them into account can significantly reduce the overall simulation cost, since they provide additional information to the modeling process.

2.1 Unattainable points

When calculating an interpolant, the problem can easily be linearized by multiplying the left hand side and right hand side of (2.1) with the denominator expression.

$$\sum_{n=0}^{N} N_n s^n = \left(\sum_{d=0}^{D} D_d s^d \right) H(s) \tag{2.2}$$

While the solution of these homogeneous equations is in general straightforward, the solution of the linearized problem will not always satisfy the rational interpolating conditions (2.1). Moreover, there may even not be a solution to the rational interpolation problem. An interpolating function satisfying (2.1) satisfies (2.2), however the inverse relation doesn't hold in general [33, 16].

Definition 1 *Two rational functions*

$$H_1(s) = \frac{N^1(s)}{D^1(s)}, H_2(s) = \frac{N^2(s)}{D^2(s)} \tag{2.3}$$

are said to be equivalent if

$$N^1(s)D^2(s) = N^2(s)D^1(s) \tag{2.4}$$

Theorem 2 *If*

$$H_1(s) = \frac{N^1(s)}{D^1(s)}, H_2(s) = \frac{N^2(s)}{D^2(s)} \tag{2.5}$$

are both non-trivial solutions of the homogeneous linear system (2.2), then they are equivalent.

Proof. Let $X(s) = N^1(s)D^2(s) = N^2(s)D^1(s)$. Then for k=0,1,...,K

$$\begin{aligned} X(s_k) &= N^1(s_k)D^2(s_k) - N^2(s_k)D^1(s_k) && (2.6) \\ &= H(s_k)D^1(s_k)D^2(s_k) - H(s_k)D^2(s_k)D^1(s_k) && (2.7) \\ &= 0 && (2.8) \end{aligned}$$

Since the order of X doesn't exceed K-1, X must be identically zero. ■

This theorem shows, that if a solution of (2.2) exists, it is essentially unique. If there is no solution, there must be some point, which is not interpolated by the rational function. Such a point leads to the indefinite form 0/0, and is called unattainable. This situation occurs when the numerator and denominator polynomial of (2.1) are not relative prime (i.e. they consist of common factors). From now on, it will be assumed that condition (2.2) can be satisfied, and that the rational models don't have unattainable points.

2.2 Least squares linearization

To parameterize the rational model in a least-squares sense, the following non-linear cost-function needs to be minimized

$$\arg\min_{N,D} \sum_{k=0}^{K} \left| R(s_k) - \frac{\sum_{n=0}^{N} N_n s_k^n}{\sum_{d=0}^{D} D_d s_k^d} \right|^2 \tag{2.9}$$

which equalizes

$$\arg\min_{N,D} \sum_{k=0}^{K} \frac{1}{\left|\sum_{d=0}^{D} D_d s_k^d\right|^2} \left| \left(\sum_{d=0}^{D} D_d s_k^d \right) R(s_k) - \sum_{n=0}^{N} N_n s_k^n \right|^2 \tag{2.10}$$

Unfortunately, the D_d terms in the denominator which act as a weighting factor, make the least-squares problem non-linear in terms of the parameters, such that it can't be solved analytically. A possible approach is the use of optimization techniques, however they are computationally not very efficient and suffer convergence to local minima. A better alternative, is to solve a (sub-optimal) linearized variation of (2.10), which converges "close enough" to the optimal solution. Possible options, which are commonly used in engineering [31, 32] are

1. Kalman's method

$$\arg\min_{N,D} \sum_{k=0}^{K} \left| \left(\sum_{d=0}^{D} D_d s_k^d \right) R(s_k) - \sum_{n=0}^{N} N_n s_k^n \right|^2 \tag{2.11}$$

2. Shanks' method

$$\arg\min_{D} \sum_{k=0}^{K} \left| \left(\sum_{d=0}^{D} D_d s_k^d \right) R(s_k) \right|^2 \tag{2.12}$$

$$\arg\min_{N} \sum_{k=0}^{K} \left| R(s_k) - \frac{\sum_{n=0}^{N} N_n s_k^n}{\sum_{d=0}^{D} D_d s_k^d} \right|^2 \tag{2.13}$$

3. Steiglitz-McBride or Sanathanan-Koerner iteration

   Solve (2.11) or (2.12) to obtain an initial guess of model parameters D (poles) in iteration $t = 0$. Set $D^{(0)} = D$, and solve (2.10) iteratively [1]

$$\arg \min_{D^{(t)},N^{(t)}} \sum_{k=0}^{K} \frac{1}{\left|\sum_{d=0}^{D} D_d^{(t-1)} s_k^d\right|^2} \left| \left(\sum_{d=0}^{D} D_d^{(t)} s_k^d \right) R(s_k) - \sum_{n=0}^{N} N_n^{(t)} s_k^n \right|^2 \tag{2.14}$$

Due to the linearization, all techniques suffer unbalanced weighting if $K + 1 > N + D + 1$, however it was shown in [14, 29] that iterative least-squares techniques such as the Steiglitz-McBride iteration have favorable convergence properties, compared to Kalman's method or a sequential estimation of the poles and zeros such as Shanks' method. A detailed analysis of these linearizations can be found in literature [32], and will not be repeated here.

Figure 2.1 illustrates the effects of unbalanced weighting, introduced by suboptimal linearization. The reflection coefficient S_{11} of a multipole filter is sampled very densely (1500 equidistant samples), and is approximated by a rational function with 80 poles. This amount of poles is more than sufficient to obtain an overall accuracy of -65 dB, since this accuracy can be obtained using rational interpolation techniques. This fit was calculated in a least-squares sense using Kalman's method by solving (2.11) in a multiprecision floating point environment with a precision of 1000 bits [6]. This approach ensures that all numerical issues are circumvented, and reveals the effects of linearization. Clearly, the accuracy of the fit isn't spread equally over the frequency range of interest, and only an accuracy of -50dB is achieved.

[1] In [13], it was proposed to flip unstable poles of the denominator into the left half of the complex plane in each iteration, to achieve Bounded-Input-Bounded-Output models. Afterwards, the numerator expression is calculated to minimize the fitting error. Since the data coming from an LTI system has a physical behaviour, it can be fitted with stable poles. However, this approach will only be successfull when the sampling of the data is dense enough, and the model order is high enough. If an (undersampled) subset of the data is fitted with a lower order model, unstable poles may be needed to approximate the system with sufficient accuracy.

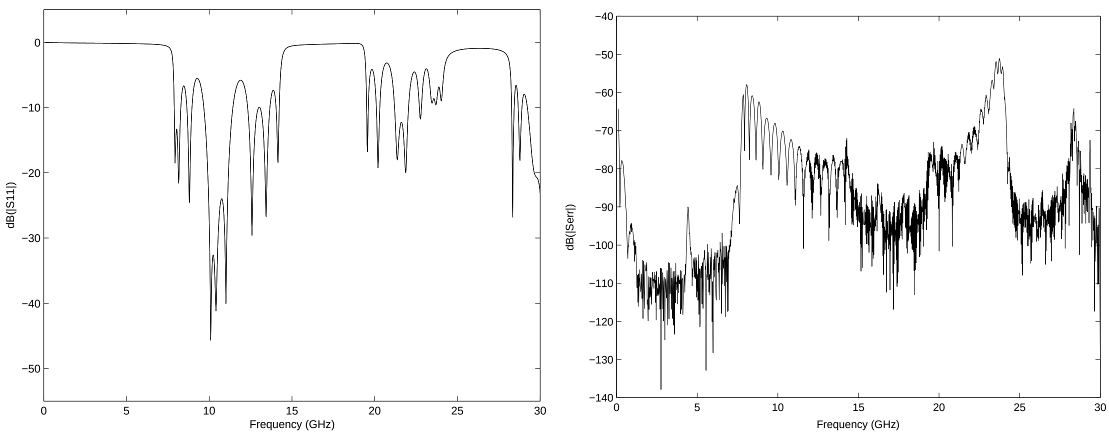

Figure 2.1: Magnitude of data and fit (left), and error of fit (right), based on Kalman linearization & Multiprecision floating-point arithmetic

SECTION 3
SELECTION OF OPTIMAL BASISFUNCTIONS.

3.1 Power Series

To solve the identification problem, it is desired to satisfy $R^{(t)}(s_k) = H^{(t)}(s_k)$ for all data samples and frequency derivatives $(s_k, H^{(t)}(s_k))$, $\forall k = 0, 1, ..., K$ and $\forall t = 0, 1, ..., T$. Traditionally, the Rational Linear Least Squares technique (RLLS) is used to calculate the coefficients of the rational model by solving a Vandermonde-like system of equations [24]. The identification problem is linearized, e.g. using Kalman's method, which leads to the following equations (2.11). At a given complex frequency point s_k, we get

$$A_k x = b_k \tag{3.1}$$

where

$$A_k = \left[\begin{array}{ccccccc} 1 & ... & s_k^N & -s_k H(s_k) & ... & -s_k^D H(s_k) \end{array} \right] \tag{3.2}$$

$$x = \left[\begin{array}{cccccc} N_0 & ... & N_N & D_1 & ... & D_d \end{array} \right]^T \tag{3.3}$$

$$b_k = [H(s_k)] \tag{3.4}$$

D_0 can be chosen as 1, since both the numerator and denominator can be divided by the same constant value without loss of generality. A direct solution with real coefficients N_n and D_d can be found, by writing out (3.1) for all frequencies s_k, and solving the following set of linear equations in a least squares sense

$$\left[\begin{array}{c} \Re(A) \\ \Im(A) \end{array} \right] x = \left[\begin{array}{c} \Re(b) \\ \Im(b) \end{array} \right] \tag{3.5}$$

When frequency derivatives of the data are available at the discrete frequencies s_k, (2.2) can be generalized. The coefficients N_n and D_d of the rational fitting model now satisfy :

$$\begin{aligned} H^{(t)}(s_k) \sum_{d=0}^{D} D_d {s_k}^d &= \sum_{n=t}^{N} N_n {s_k}^{n-t} \frac{n!}{(n-t)!} \\ &- \sum_{m=1}^{t} \sum_{d=m}^{D} \left(\begin{array}{c} t \\ m \end{array} \right) H^{(t-m)}(s_k) D_d {s_k}^{d-m} \frac{d!}{(d-m)!} \end{aligned} \tag{3.6}$$

where $H^{(t)}$ is the tth order derivative of the frequency domain data. All derivatives are relative to s. The set of equations at all frequencies s_k and for all derivatives t, can be written in a similar matrix form as equation (3.5).

Although some of the numerical problems can be relieved by scaling the frequencies [25], the conditioning of this system deteriorates fast when the order of the model increases, or when the frequency range of interest gets more broad.

3.2 Orthogonal Fitting Techniques

Chebyshev polynomials of the first kind

Definition 3 *An inner product $\langle .,. \rangle$ is a bilinear function of elements p_u,p_v,p_w of a vector space that satisfies the axioms :*

1. $\langle p_u, p_u \rangle \geq 0$ *with equality if and only if* $p_u \equiv 0$
2. $\langle p_u, p_v \rangle = \langle p_v, p_u \rangle$
3. $\langle p_u + p_v, p_w \rangle = \langle p_u, p_w \rangle + \langle p_v, p_w \rangle$
4. $\langle \alpha p_u, p_w \rangle = \alpha \langle p_u, p_w \rangle$ *for any scalar* α

Definition 4 *Two functions $p_u(x)$ and $p_v(x)$ in $L_2[a,b]$, are said to be orthogonal on the interval [a,b] with respect to a given continuous and non-negative weight function w(x) if*

$$< p_u(x), p_v(x) >= \int_a^b w(x) p_u(x) p_v(x) dx = 0 \tag{3.7}$$

Definition 5 *A Chebyshev polynomial $T_i(x)$ of the first kind is a polynomial in x of degree i, defined by the relation $T_i(x) = \cos(i\phi)$ when x=cos ϕ.*

From Simpson's rule, it follows that

$$\begin{aligned} \cos(i\phi) + \cos(i-2)\phi &= 2\cos\left(\frac{i\phi + (i-2)\phi}{2}\right) . \cos\left(\frac{i\phi - (i-2)\phi}{2}\right) \\ &= 2\cos(i-1)\phi\cos\phi \end{aligned} \tag{3.8}$$

This trigonometric identity leads to the fundamental recurrence relation $T_i(x) = 2xT_{i-1}(x) - T_{i-2}(x)$ together with the initial conditions $T_0(x) = 1, T_1(x) = x$.

If the continuous inner product is defined as

$$\langle T_u(x), T_v(x) \rangle = \int_a^b w(x) T_u(x) T_v(x) dx = 0 \tag{3.9}$$

using the interval $[a,b] = [-1,1]$ and weight function $w(x) = (1-x^2)^{-\frac{1}{2}}$, then the Chebyshev polynomials of the first kind satisfy $< T_u, T_v >= 0$ if u$\neq$v [21].

The numerator and denominator polynomial of the rational function $R(s)$ can be represented as a linear combination of Chebyshev polynomials of the first kind $T_i(\omega)$ where ω represents the angular frequency $2\pi f$. The even Chebyshev polynomials consist of even powers of ω, while the odd Chebyshev polynomials only consist of odd powers of ω. Since the Chebyshev polynomials have real coefficients and a real input variable ω, the odd terms are multiplied with complex j.

$$R(s) = \frac{N(s)}{D(s)} = \frac{\sum_{n,even} N_n T_n(\omega) + j. \sum_{n,odd} N_n T_n(\omega)}{\sum_{d,even} D_d T_d(\omega) + j. \sum_{d,odd} D_d T_d(\omega)} \tag{3.10}$$

The identification problem can be linearized, e.g. using Kalman's method, which leads to the following homogenous equations

$$\left(\sum_{d,even} D_d T_d(\omega) + j.\sum_{d,odd} D_d T_d(\omega)\right) H(j\omega) = \sum_{n,even} N_n T_n(\omega) + j.\sum_{n,odd} N_n T_n(\omega) \quad (3.11)$$

The linear equations are solved for all angular frequencies ω, in order to determine the unknown system parameters N_n and D_d. To ensure complex conjugacy of the poles and zeros, the real and imaginary part of each equation is separated, as in (3.5). The large variations of the Chebyshev polynomials make it possible to downsize the effects of the ill-conditioned matrix, by summing the orthogonal Chebyshev polynomials, instead of summing the power series, which show little variation with increase in order [23]. Therefore, the system of equations is usually better conditioned.

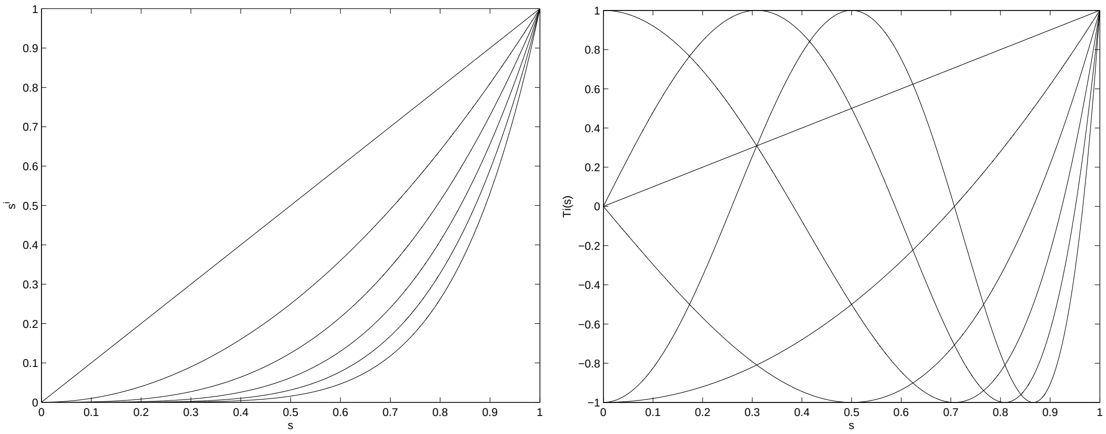

Figure 3.1: Power series and Chebyshev polynomials on the interval [0,1]

From (3.9), it follows that the angular frequencies must be scaled and shifted to the unit interval [-1,1] in order to take fully advantage of the orthogonality. Under normal circumstances however, this property can't be satisfied since the transfer function (and consequently also the frequency response) is required to be hermitian symmetric over the real axis. To solve this problem, the frequencies must be scaled into]0,1], e.g. by dividing all frequencies by $\omega_{\max}$. As the poles and zeros of the model are usually desired in terms of the monomial basis, an inverse basis transformation from the Chebyshev polynomials to the power series is required. However, this requires the calculation of the polynomial coefficients, which can be ill-conditioned. Also, the polynomials should be evaluated recursively to obtain best accuracy.

Orthonormal Forsythe Polynomials It is always possible to convert a continuous orthogonality relationship into a discrete orthogonality relationship simply by replacing the integral with a summation. Also the inner product can be defined on a discrete data set, just as well as on a continuum. [21]

Forsythe polynomials are formal orthogonal polynomials, which are defined by the following three-term recurrence [10].

$$\begin{aligned} p_{-1}(f) &= 0 \\ P_0(f) &= 1 \\ &\cdots \\ P_i(f) &= j(2\pi f - \alpha_{i-1})p_{i-1}(f) + \beta_{i-1}p_{i-2}(f) \end{aligned} \tag{3.12}$$

The α and β coefficients are selected to make the orthogonality conditions hold over the discrete sample set $\{f_k\}$

$$\alpha_i = \frac{\sum_{k=0}^{K} 2\pi f_k (w_k P_i(f_k))(w_k P_i(f_k))^*}{\sum_{k=0}^{K} (w_k P_i(f_k))(w_k P_i(f_k))^*} \tag{3.13}$$

$$\beta_i = \sqrt{\sum_{k=0}^{K} (w_k P_i(f_k))(w_k P_i(f_k))^*} \tag{3.14}$$

and the polynomials are orthonormalized by

$$p_i(f) = \frac{P_i(f)}{\beta_i} \tag{3.15}$$

Summating over all negative and positive sample frequencies, and taking advantage of the fact that $w_k = w_{-k}^*$, the following simplified recursive relation (α_i=0) is obtained

$$\begin{aligned} p_{-1}(f) &= 0 \\ P_0(f) &= 1 \\ &\cdots \\ P_i(f) &= j2\pi f p_{i-1}(f) + \beta_{i-1}p_{i-2}(f) \end{aligned} \tag{3.16}$$

where the β coefficients are defined as in (3.14). This way, a simplified formulation of the polynomials is given as

$$p_i(f) = \frac{P_i(f)}{\beta_i} \tag{3.17}$$

Due to the orthonormality, the magnitude of the Forsythe polynomials is normalized in the sample frequencies with respect to the discrete inner product

$$\sum_{\forall k} w_k p_i(f_k)(w_k p_j(f_k))^* = \delta_{ij} \quad \forall 0 \le i, j < K \tag{3.18}$$

Similar to (3.10), the numerator and denominator polynomial of the rational function $R(s)$ can be represented as a linear combination of generalized Forsythe polynomials [26].

$$R(s) = \frac{N(f)}{D(f)} = \frac{\sum_{n=0}^{N} N_n p_n(f)}{\sum_{d=0}^{D} D_d q_d(f)} \tag{3.19}$$

$p_i(f)$ and $q_i(f)$ are orthonormal with respect to the inner product (3.18), provided that the weighting factor w_k for the numerator polynomial is set to 1, and the weighting factor for the denominator polynomial is set to $H(s_k)$. The identification problem can be linearized, e.g. using Kalman's method, which leads to the following homogenous equations

$$\left(\sum_{d=0}^{D} D_d q_d(f)\right) H(s) = \sum_{n=0}^{N} N_n p_n(f) \tag{3.20}$$

Let's define the vectors U_r,V_r,W and the matrices U and V as

$$\begin{aligned} U_r &= [p_r(s_0) \ldots p_r(s_K)\ p_r(s_0^*) \ldots p_r(s_K^*)]^T && (3.21)\\ V_r &= [H(s_0)q_r(s_0) \ldots H(s_K)q_r(s_K)\ H^*(s_0)q_r(s_0^*) \ldots\ H^*(s_K)q_r(s_K^*)]^T && (3.22)\\ U &= [U_0 \ldots U_N]\ and\ V = [V_1 \ldots V_D] && (3.23) \end{aligned}$$

Then the normal equations are obtained as

$$\begin{pmatrix} Y & X \\ X^T & Z \end{pmatrix} \begin{pmatrix} N \\ D \end{pmatrix} = \begin{pmatrix} G \\ F \end{pmatrix} \tag{3.24}$$

with $Y = (U^*)^T(U)$, $Z = (V^*)^T(V)$, $X = -\Re((U^*)^T(V)), G = \Re((U^*)^T(V_0))$, and $F = \Re((V^*)^T(V_0))$. Due to the orthonormality of the Forsythe polynomials, $Y = Z = I$, where I represents the identity matrix, and F=0. This way, real coefficients are obtained when following two systems of equations are solved consecutively.

$$\begin{aligned} (I - (X^T X))D &= -X^T G && (3.25)\\ N &= G - XD && (3.26) \end{aligned}$$

In [27, 11], it is proven that this approach makes the normal equations best conditioned, in a sense that no other polynomial basis can be found resulting in a better conditioned form of the normal equations. Determination of the polynomial coefficients can be ill-conditioned, so although it is time-consuming, a recursive evaluation of the polynomials is required in order to get accurate results. As new samples become available, all orthogonal polynomials need to be recalculated, which is computationally not efficient. Moreover, the technique doesn't extend well to MIMO systems, because the orthogonality of the denominator polynomials depends on the S-parameters.

When frequency derivatives of the data are available, the Forsythe polynomials are generalized. If $p_i^{(t)}(f_k)$ represents the t^{th} order derivative of the i^{th} Forsythe polynomial, evaluated in frequency f_k, it can be expressed as

$$p_i^{(t)}(f_k) = \frac{t p_{i-1}^{(t-1)}(f_k) + j2\pi f p_{i-1}^{(t)}(f_k) - \beta_{i-1} p_{i-2}^{(t)}(f_k)}{\beta_i} \tag{3.27}$$

Hence, the coefficients N_n and D_n of the rational fitting model now satisfy :

$$\begin{aligned} H^{(t)}(s_k) \sum_{d=0}^{D} D_d q_d^{(0)}(f_k) &= \sum_{n=t}^{N} N_n p_n^{(t)}(f_k) && (3.28)\\ &\quad - \sum_{m=1}^{t} \sum_{d=m}^{D} \begin{pmatrix} t \\ m \end{pmatrix} H^{(t-m)}(s_k) D_d q_d^{(m)}(f_k) \end{aligned}$$

where $p_n^{(t)}$ is the t^{th} order derivative of the n^{th} order numerator Forsythe polynomial, $q_d^{(t)}$ is the t^{th} order derivative of the d^{th} order denominator Forsythe polynomial, and $H^{(t)}$ is the t^{th} order derivative of the frequency domain data. All derivatives are relative to $j2\pi f$. The set of equations at all frequencies f_k and for all derivatives t, can be solved in terms of the unknowns N_n and D_d. A breakdown of the orthogonality occurs first for $i = K{-}1$, when (3.13) fails. This problem can remedied by introducing virtual samples.

A modification of Forsythe's algorithm, due to Clenshaw, is to represent the polynomials by their Chebyshev expansion [3]. Although this method is computationally less efficient than the original in terms of arithmetic operations, this form is more concise and convenient in terms of storage requirements.

Lanczos-based Methods A similar approach was made in [4, 18]. Let's define matrix V and vector H as

$$V_r \;=\; [s_0^r\ s_1^r\ ...\ s_K^r]^T \quad and \tag{3.29}$$

$$H \;=\; diag(H(s_0), .., H(s_K)) \tag{3.30}$$

Then the identification problem can be formulated in terms of the unknowns, e.g. using Kalman's method, as

$$\begin{pmatrix} V_{0:N} & -HV_{1:D} \end{pmatrix} \begin{pmatrix} N \\ D \end{pmatrix} = (HV_0) \tag{3.31}$$

The columns of the Vandermonde matrix V are $[\mathit{1}, S\mathit{1}, S^2\mathit{1}, ..., S^r\mathit{1}]$ where S=diag($[s_1,s_2,...,s_K]$) and $\mathit{1}$ is a K column vector with all entries set to 1. Hence, the columns of V form a Krylov subspace $K_{r+1}(S, \mathit{1})$. Since S is symmetric, an orthonormal basis is generated using a symmetric Lanczos method, which produces the factorization $SQ = QT$ [28]. It follows that Q has orthonormal columns which span $K_{r+1}(S, \mathit{1})$, and T is tridiagonal. Based on the elements of T, the polynomial basis is defined by a three-term recurrence relation similar to (3.12),

$$p_i(s) = \left(\frac{s - t_{i,i}}{t_{i+1,i}}\right) p_{i-1}(s) - \frac{t_{i-1,i}}{t_{i+1,i}} p_{i-2}(s) \tag{3.32}$$

To obtain polynomials which satisfy complex conjugate symmetry, it is required that the basis polynomials are orthogonal on both sides of the frequency axis. This corresponds to generating $K_{r+1}(\tilde{S}, \tilde{\mathit{1}})$ where $\tilde{S}$=diag($[s_1,...,s_K,s_1^*,...,s_K^*]$), $\tilde{H} = diag(H, H^*)$ and $\tilde{\mathit{1}}$ is a $2K$ column vector with all entries set to 1. In practice a similar result is obtained by using a modified Arnoldi process where only the real projections are used. This way, the orthogonal polynomial basis is defined by a simplified three-term recurrence relation ($\tilde{t}_{i,i} = 0$), similar to (3.16).

$$p_i(s) = \frac{s}{\tilde{t}_{i+1,i}} p_{i-1}(s) - \frac{\tilde{t}_{i-1,i}}{\tilde{t}_{i+1,i}} p_{i-2}(s) \tag{3.33}$$

Hence, the coefficients of the rational model

$$R(s) = \frac{N(s)}{D(s)} = \frac{\sum_{n=0}^{N} N_n p_n(s)}{\sum_{d=0}^{D} D_d p_d(s)} \tag{3.34}$$

are calculated in the orthogonal basis by solving

$$\begin{pmatrix} \tilde{Q}_{0:N} & -\tilde{H}\tilde{Q}_{1:D} \end{pmatrix} \begin{pmatrix} N \\ D \end{pmatrix} = \begin{pmatrix} \tilde{H}\tilde{Q}_0 \end{pmatrix} \tag{3.35}$$

Since the dual-basis approach with Forsythe polynomials was optimal in the sense that it makes the normal equations best conditioned, no real breakthrough was achieved with this technique. Furthermore it shares the same advantages and disadvantages, however, using the same basis in numerator and denominator simplifies the extension to MIMO systems with common poles.

When frequency derivatives of the data are available, the orthonormal polynomials are generalized. If $p_i^{(t)}(s)$ represents the t^{th} order derivative of the i^{th} polynomial, evaluated in frequency s, it can be expressed as

$$p_i^{(t)}(s_k) = \frac{t p_{i-1}^{(t-1)}(s_k) + s p_{i-1}^{(t)}(s_k) - \tilde{t}_{i-1,i} p_{i-2}^{(t)}(s_k)}{\tilde{t}_{i+1,i}} \tag{3.36}$$

Hence, the coefficients N_n and D_n of the rational fitting model now satisfy :

$$\begin{aligned} H^{(t)}(s_k) \sum_{d=0}^{D} D_d p_d^{(0)}(s_k) &= \sum_{n=t}^{N} N_n p_n^{(t)}(s_k) \\ &\quad - \sum_{m=1}^{t} \sum_{d=m}^{D} \begin{pmatrix} t \\ m \end{pmatrix} H^{(t-m)}(s_k) D_d p_d^{(m)}(s_k) \end{aligned} \tag{3.37}$$

where $p_n^{(t)}$ is the t^{th} order derivative of the n^{th} order numerator polynomial, $p_d^{(t)}$ is the t^{th} order derivative of the d^{th} order denominator polynomial, and $H^{(t)}$ is the t^{th} order derivative of the frequency domain data. All derivatives are relative to $j2\pi f$. The set of equations at all frequencies s_k and for all derivatives t, can be solved in terms of the unknowns N_n and D_d.

3.3 Vector Fitting

Recently, a new approximation technique was introduced, which fits the frequency response with a causal pole-residue model [13] [2].

$$R(s) = \sum_{p=1}^{P} \frac{c_p}{s - a_p} + d \tag{3.38}$$

$H(s)$ with $s = j2\pi f$ can be approximated by a rational function, based upon an initial set of P starting poles $\bar{a}_p$, multiplied with an unknown function $\sigma(s)$. Since $\sigma(s)$

[2] To make the order of numerator significantly larger than the order of the denominator, higher-order polynomials terms can be added to the pole-residue form. By interpolating the inverse of the data samples $H(s_k)$, the order of numerator and denominator can be interchanged.

is also rational, it can be represented in pole-residue form, which leads to the following augmented problem :

$$\begin{bmatrix} \sigma(s)R(s) \\ \sigma(s) \end{bmatrix} = \begin{bmatrix} \sum_{p=1}^{P} \frac{c_p}{s-\bar{a}_p} + d \\ \sum_{p=1}^{P} \frac{\tilde{c}_p}{s-\bar{a}_p} + 1 \end{bmatrix} \tag{3.39}$$

The problem can be linearized in terms of the unknowns c_p, d and $\tilde{c}_p$ by multiplying the second line of the vector equation (3.39) with $R(s)$.

$$\left(\sum_{p=1}^{P} \frac{c_p}{s-\bar{a}_p} + d\right) = \left(\sum_{p=1}^{P} \frac{\tilde{c}_p}{s-\bar{a}_p} + 1\right) R(s) \tag{3.40}$$

or

$$(\sigma R)_{fit}(s) = \sigma_{fit}(s) R(s) \tag{3.41}$$

Since R(s) should equal H(s) at each frequency sample, we get

$$A_k x = b_k \tag{3.42}$$

where

$$A_k = \begin{bmatrix} \frac{1}{s_k-\bar{a}_1} & ... & \frac{1}{s_k-\bar{a}_P} & 1 & \frac{-H(s_k)}{s_k-\bar{a}_1} & ... & \frac{-H(s_k)}{s_k-\bar{a}_P} \end{bmatrix} \tag{3.43}$$

$$x = \begin{bmatrix} c_1 & ... & c_P & d & \tilde{c}_1 & ... & \tilde{c}_P \end{bmatrix}^T \tag{3.44}$$

$$b_k = \begin{bmatrix} H(s_k) \end{bmatrix} \tag{3.45}$$

Writing out (3.42) for all frequencies s_k gives an overdetermined system of equations :

$$Ax = b \tag{3.46}$$

To ensure that complex conjugacy of the residues, corresponding to complex poles $\bar{a}_p$ and $\bar{a}_{p+1}$ is guaranteed, the following modification is performed. Suppose that the poles $\bar{a}_p$ and $\bar{a}_{p+1}$ of two partial fractions constitute a complex conjugate pair $\bar{a}_p = a' + ja'', \bar{a}_{p+1} = a' - ja''$. Then the corresponding vector columns $A_{k,p}$ and $A_{k,p+1}$ of (3.43) can be replaced by

$$A'_{k,p} = \frac{1}{s_k - \bar{a}_p} + \frac{1}{s_k - \bar{a}_p^*} \quad and \quad A'_{k,p+1} = \frac{j}{s_k - \bar{a}_p} - \frac{j}{s_k - \bar{a}_p^*} \tag{3.47}$$

such that the residues $c_p = c' + jc'', c_{p+1} = c' - jc''$ satisfy complex conjugacy as well. This has the effect that the corresponding residues in the solution vector x, become equal to c' and c'' respectively.

To preserve that the coefficients of the rational functions are real, (3.46) is formulated as :

$$\begin{bmatrix} \Re(A) \\ \Im(A) \end{bmatrix} x = \begin{bmatrix} \Re(b) \\ \Im(b) \end{bmatrix} \tag{3.48}$$

After the parameterization of the rational model, the following function approximations are obtained

$$(\sigma R)_{fit}(s) = \frac{\prod_{p=1}^{P}(s - z_p)}{\prod_{p=1}^{P}(s - \bar{a}_p)} \quad and \quad \sigma_{fit}(s) = \frac{\prod_{p=1}^{P}(s - \tilde{z}_p)}{\prod_{p=1}^{P}(s - \bar{a}_p)} \tag{3.49}$$

From (3.49), $R(s)$ can be calculated

$$R(s) = \frac{(\sigma R)_{fit}(s)}{\sigma_{fit}(s)} = \frac{\prod_{p=1}^{P}(s - z_p)}{\prod_{p=1}^{P}(s - \tilde{z}_p)} \tag{3.50}$$

The residues of $R(s)$ can be obtained by solving (3.38) with the zeros of $\sigma(s)$ as new poles for $R(s)$.

When frequency derivatives are available, a modification is performed by deriving the left hand side and right hand side of equation (3.40) [9]. This leads to the following system of equations for a certain frequency point s_k :

$$A_k = \begin{bmatrix} \psi_{1,0}(s_k) & \ldots & \psi_{P,0}(s_k) & 1 & \phi_{1,0}(s_k) & \ldots & \phi_{P,0}(s_k) \\ \psi_{1,1}(s_k) & \ldots & \psi_{P,1}(s_k) & 0 & \phi_{1,1}(s_k) & \ldots & \phi_{P,1}(s_k) \\ \ldots & \ldots & \ldots & \ldots & \ldots & \ldots & \ldots \\ \psi_{1,T}(s_k) & \ldots & \psi_{P,T}(s_k) & 0 & \phi_{1,T}(s_k) & \ldots & \phi_{P,T}(s_k) \end{bmatrix} \tag{3.51}$$

$$x = \begin{bmatrix} c_1 & \ldots & c_P & d & \tilde{c}_1 & \ldots & \tilde{c}_P \end{bmatrix}^T \tag{3.52}$$

$$b_k = \begin{bmatrix} H^{(0)}(s_k) & H^{(1)}(s_k) & \ldots & H^{(T)}(s_k) \end{bmatrix}^T \tag{3.53}$$

where $\psi_{p,t}(s_k)$ is defined as

$$\psi_{p,t}(s_k) = \frac{d^t}{ds^t}\left[(s_k - \bar{a}_p)^{-1}\right] = (-1)^t t!(s_k - \bar{a}_p)^{-(t+1)} \tag{3.54}$$

and based on Leibniz' identity, $\phi_{p,t}(s_k)$ represent

$$\phi_{p,t}(s_k) = \frac{d^t}{ds^t}[-H(s_k)(s_k - \bar{a}_p)^{-1}] = -\sum_{\tau=0}^{t} \left(\begin{array}{c} t \\ \tau \end{array} \right) H^{(t-\tau)}(s_k)\psi_{p,\tau}(s_k) \tag{3.55}$$

To enforce complex conjugacy of poles and residues, a modification similar to (3.47) is performed, by replacing the corresponding column vectors $A_{k,p}$ and $A_{k,p+1}$ by $A'_{k,p} = \psi_{p,t}(s_k) + \psi_{p+1,t}(s_k)$ and $A'_{k,p+1} = j\psi_{p,t}(s_k) - j\psi_{p+1,t}(s_k)$.

When the system is solved in a least squares sense, the new poles can be reused as initial starting poles and the fitting process is repeated several times.

Essentially, this method is an "intuitive" reformulation of a Sanathanan-Koerner iteration with rational basis functions [14,7,15].

Suppose the numerator and denominator expression of the rational function is expanded as a linear combination of partial fractions based on common poles, which spans the same space as (3.38).

$$R(s) = \frac{\sum_{p=1}^{P} \frac{c_p}{s - \bar{a}_p} + d}{\sum_{p=1}^{P} \frac{\tilde{c}_p}{s - \bar{a}_p} + 1} \tag{3.56}$$

Then the identification problem can be linearized using Kalman's method which reduces to the following set of homogenous equations

$$\left(\sum_{p=1}^{P}\frac{c_p}{s-\bar{a}_p}+d\right)-R(s)\left(\sum_{p=1}^{P}\frac{\tilde{c}_p}{s-\bar{a}_p}+1\right) \tag{3.57}$$

Note that equation (3.57) reduces exactly to the pole-identification of Vector Fitting (3.40).

Since usually, the poles and residues of the model require less significant digits than the model parameters, more accurate results are achieved than provided by the former techniques. Also, the inversion of a Vandermonde-like matrix, and the use of orthogonal polynomials is avoided by solving a Cauchy-like system of equations. The main disadvantage of the technique, is that the conditioning of (3.46) is highly dependent on the initial choice of starting poles. However, the conditioning often improves when the poles converge. For a detailed analysis of the optimal initial pole-location, the reader is referred to [13,30].

3.4 Thiele-type Continued Fractions

The use of continued fractions as rational interpolants in the design of microwave circuits was proposed in [19], and later extended to the multivariate case [20]. The rational model can be represented as a convergent of a corresponding Thiele-type continued fraction

$$R_k(s) = H(s_0)+\cfrac{s-s_0}{\phi_1(s_1,s_0)+\cfrac{s-s_1}{\phi_2(s_2,s_1,s_0)+\ldots\cfrac{}{\ldots+\cfrac{s-s_{k-1}}{\phi_k(s_k,s_{k-1},\ldots,s_0)}}}} \tag{3.58}$$

$$= H(s_0)+\sum_{m=1}^{k}\frac{s-s_{m-1}\quad|}{|\ \phi_m(s_m,s_{m-1}..,s_0)}\quad for\ k=0,...,K \tag{3.59}$$

Each rational expression $R_k(s)$ is a k^{th} order partial fraction expansion of (2.1), together constituting a set of interpolants which exhibit increasing accuracy as k increases, reaching a convergent value at k=K. The inverse differences ϕ_k, are the partial denominators of (3.58), and are essentially the coefficients that define $R_k(s)$. They are determined recursively from the samples and are defined as follows [33] :

$$\phi_1(s_m,s_0)=\frac{s_m-s_o}{H(s_m)-H(s_0)} \tag{3.60}$$

for $m=1,2,...,K$ and

$$\phi_k(s_m,s_{k-1},...s_0)=\frac{s_m-s_{k-1}}{\phi_{k-1}(s_m,s_{k-2},...,s_0)-\phi_{k-1}(s_{k-1},s_{k-2},...,s_0))} \tag{3.61}$$

for $m=k,k+1,...,K$ and for $k=2,3,...,K$

The interpolating function $R_k(s)$ can be evaluated numerically with a three-term recurrence relation

$$N_k(s) = \phi_k(s_k, s_{k-1}, ..., s_0)N_{k-1}(s) + (s - s_{k-1})N_{k-2}(s) \quad for\ k = 2, 3, ..K \tag{3.62}$$
$$D_k(s) = \phi_k(s_k, s_{k-1}, ..., s_0)D_{k-1}(s) + (s - s_{k-1})D_{k-2}(s) \quad for\ k = 0, 1, ..K \tag{3.63}$$

which is initialized by $N_0(s) = H(s_0)$, $N_1(s) = \phi_1(s_1, s_0)N_0 + (s - s_0)$, $D_0(s) = 1$ and $D_1(s) = \phi_1(s_1, s_0)$. As a consequence of the continued fraction formulation, $N = D = \frac{k}{2}$ for k even and $N = \frac{k+1}{2}$ and $D = \frac{k-1}{2}$ for k odd. [1]

The main advantage of the technique is that it provides numerically accurate results, and interpolates the given dataset. Also, the continued fraction can easily be updated with minimal computational expense, when a new data sample is selected. In [17], a generalization is proposed, which can be applied to fit frequency derivatives. To make the coefficients real-valued, the samples on the negative part of the imaginary axis $(s_k^*, H^*(s_k))$ should be interpolated as well, such that $R(s_k^*) = R^*(s_k)$.

Unfortunately, there are some disadvantages too. This approach based on continued fractions can't be used to approximate the data in a least-squares sense, so the technique fails when the data is contaminated with noise. Also it doesn't extend well to MIMO systems. The conversion from a continued-fraction representation to a ratio of polynomials can imply a loss of accuracy, and successive values of the desired transfer function shouldn't have the same values in order to avoid singularities in the inverse differences. Although the technique provides the ability to precisely specify the frequency response at particular frequencies, it usually causes the response at other frequencies to vary widely from what would be expected. This implies that more samples are needed for smooth functions, compared to least-squares approaches. Also, the model complexity may become excessively large when the sampling of the data is not optimal (too dense).

[1] The technique can be extended to more generalized staircases, by combining rational with polynomial interpolation, and interpolating the inverse of the data samples $H(s_k)$.

SECTION 4
NUMERICAL RESULTS

In order to give a fair comparison of the techniques, one should consider which form of representation is most appropriate to represent the final transfer function. The answer to this problem is not straightforward, since some representations are well suited from a mathematical point of view, while engineers may argue that other representations are preferable from a practical perspective. For the sake of argument, it was decided by the authors to represent the transfer function in the form which is numerically most preferable, considering each technique separately.

Figure 4.1 shows the magnitude of the transmission-line coefficient of a lowpass filter over the frequency range of interest [2 GHz - 6 GHz]. To analyze the fitting techniques on this data, an iterative technique is applied. First, an initial set of 4 equidistantly spaced data samples is selected, and rational fitting models are calculated using the different techniques. The transfer function is chosen to be strictly causal ($N - 1 = D$), and the number of poles is chosen high enough, in order to obtain an interpolating function. This ensures that undesired linearization effects are excluded from the numerical results. In each iteration, the number of equidistant samples is incremented, and the number of iterations was set to 100.

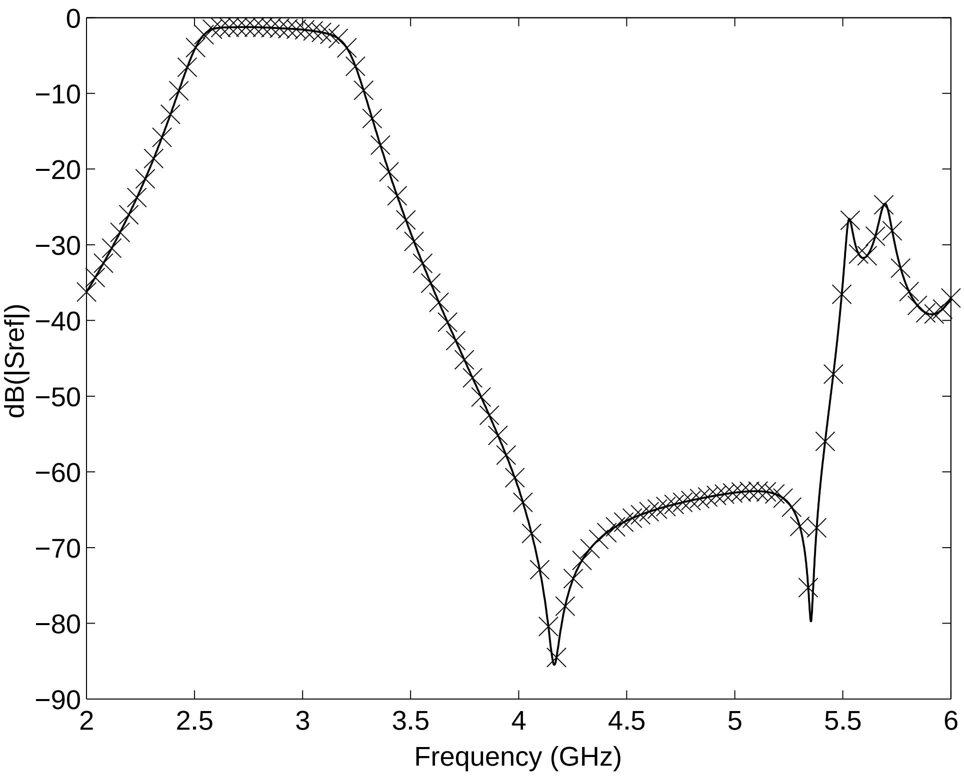

Figure 4.1: Magnitude of data vs. frequency

The equations of the least-squares techniques were solved using Matlab's backslash operator (QR decomposition with column pivoting).

Figure 4.2 shows the condition number of the equations which are solved in function of the number of selected samples. Clearly, the numerical conditioning deteriorates fast when the numerator and denominator polynomial of the transfer function are expanded in a power series basis. The other, more advanced techniques, perform significantly better.

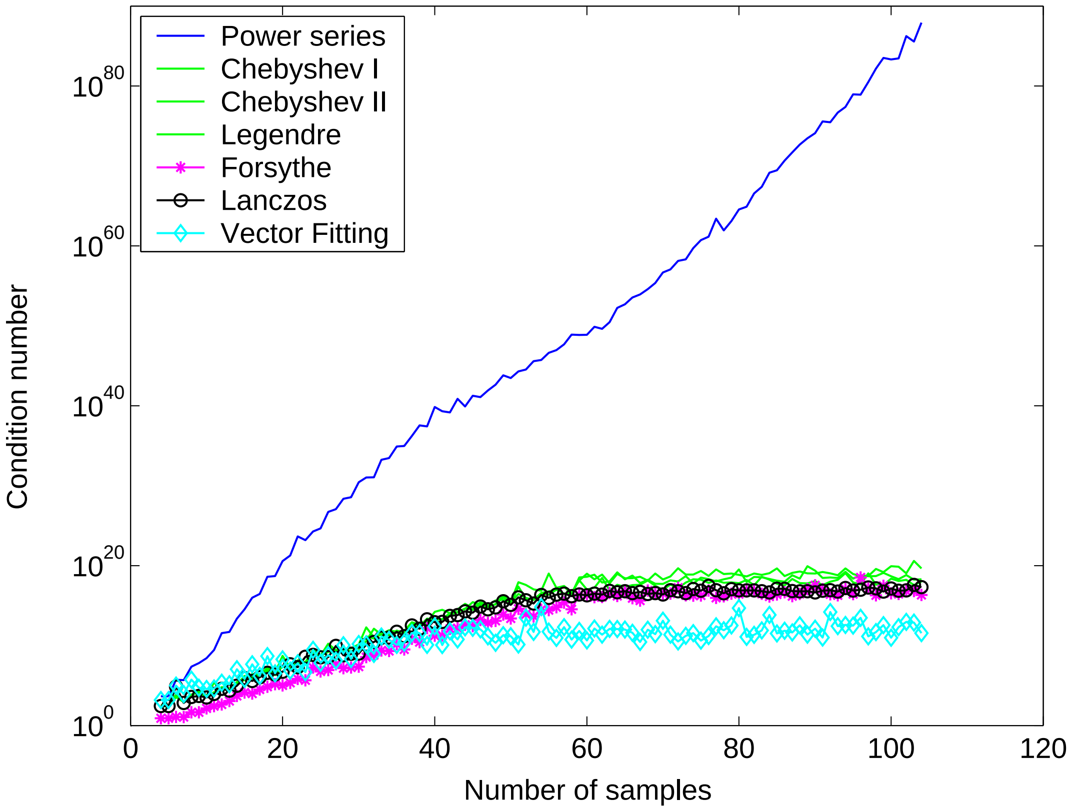

Figure 4.2: Condition number of fit vs. number of selected data samples

A close-up of Figure 4.2 is shown in Figure 4.3. When the polynomial basis functions are orthogonalized w.r.t. a continuous inner product (e.g. Chebyshev I, Chebyshev II, or Legendre polynomials), the condition number is very similar, and no real difference can be distinguished. The orthonormalization w.r.t. a discrete inner product (Lanczos) improves the structure of the equations, and leads to a small improvement in numerical conditioning. This can be further optimized, when the numerator and denominator basisfunctions are orthonormalized separately, w.r.t. an appropriate weighting function (Forsythe).

Vector Fitting, on the other hand, uses rational basis functions to estimate the poles and zeros of the fitting model. The algorithm starts from a well-chosen initial set of poles, and was run for 3 iterations. Figure 4.2 and 4.3 shows the condition number of the pole-identification in the final iteration. Before comparing this result to the polynomial fitting techniques, one should keep in mind that this technique estimates residues of partial fractions, which require less significant digits than coefficients of a polynomial. Therefore a pole-residue model is often a better representation for broadband solutions, since the coefficients of numerator and denominator polynomial may simply not be representable in 16-bit machine precision.

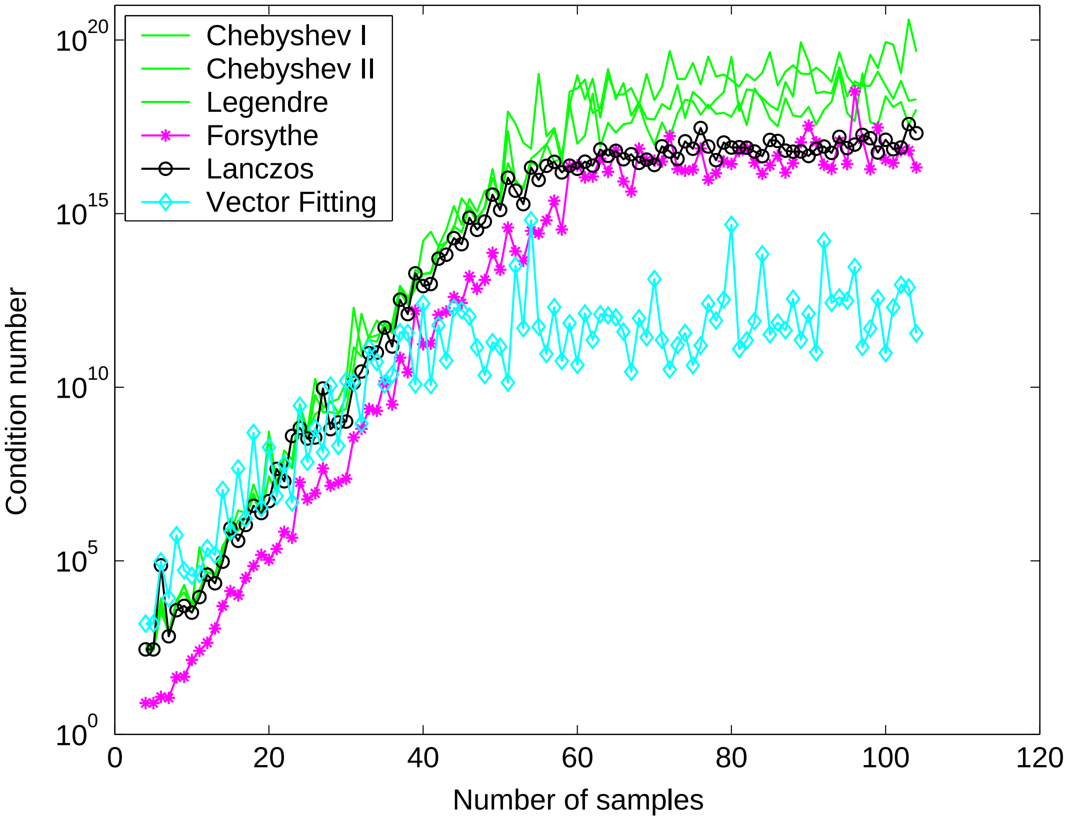

Figure 4.3: Condition number of fit vs. number of selected data samples

The comparison becomes more clear when the fitting error (i.e. the maximal error in the selected data samples) is considered. As can be seen in Figure 4.4, the same conclusions hold for the power series, and orthogonal fitting techniques. Especially interesting is the fitting error of the Vector Fitting technique which is highly accurate (near machine precision), and quite comparable to interpolation techniques based on continued fractions. The main advantage compared to the Tiele CF - approach, is that Vector Fitting can be applied to least-squares solutions, where overdetermined equations need to be solved and the data may be contaminated with noise. Also the pole-residue model is practically more useful and easy to handle, compared to a continued fraction.

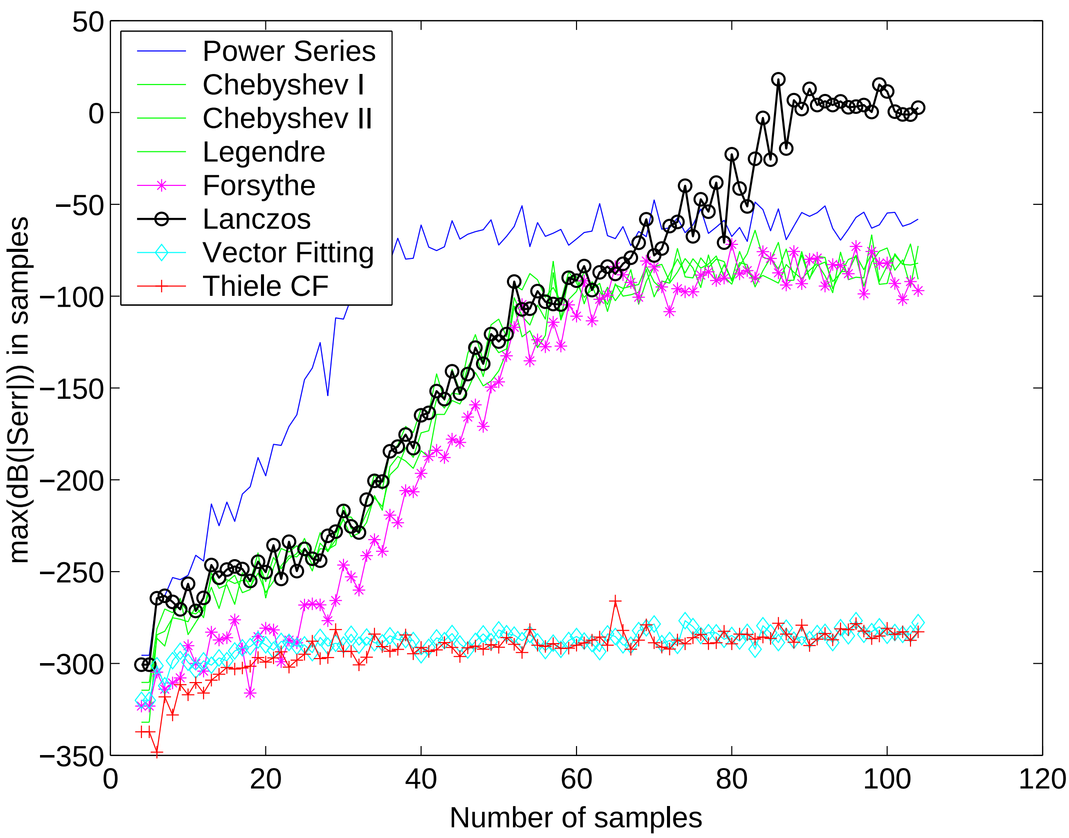

Figure 4.4: Accuracy in data samples vs. number of selected data samples

SECTION 5
CONCLUSIONS

Compact rational macromodels of high-speed passive components, based on full-wave electromagnetic simulation data, are paramount to the success and efficiency of circuit simulation. Solving the rational least-squares approximation problem in a power series basis gives rise to an overdetermined set of highly ill-conditioned equations. A polynomial basis, which is orthogonal w.r.t. a continuous inner product improves the conditioning and the accuracy of the fitting model. The use of a dual basis of orthogonal polynomials, which are orthogonal w.r.t. a discrete inner product with proper weighting improves the structure of the equations, and often gives a small additional advantage at the expense of a higher computational cost. Nevertheless, the advantage of using orthogonal polynomials is often lost, when the rational model needs to be converted into a suitable representation for SPICE-like circuit simulators (e.g. ratio of polynomials, state space representation, or partial fraction expansion). The use of rational bases, combined with iterative least-squares methods, gives rise to a better conditioning and an accuracy which is close to machine precision, and comparable to rational interpolation techniques (Thiele continued fractions). It allows to realize the transfer function as a pole-residue model, which is very fast to evaluate. It doesn't need to be transformed into any other representation, as it can easily be converted into a SPICE netlist.

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Loss Effects on the RCS of a Conducting Circular or Elliptic Cylinder with a Metamaterial Coating

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Abstract

The problem of electromagnetic scattering by a lossy metamaterial coated circular or elliptic cylinder is analyzed using elliptic waves expressed in terms of Mathieu functions with complex arguments. Numerical results are obtained for the far scattered field for different axial ratio, lossy metamaterial coating and angles of incidence. The numerical results show a significant reduction in the backscattering echo pattern width due to the presence of a lossy metamaterial coating when it is compared with losseless conventional dielectric and metamaterial coatings.

1. INTRODUCTION

The problem of electromagnetic scattering by dielectric elliptic cylinders provides a useful model to study the electromagnetic scattering by the fuselage of aircrafts. On the other hand, analytical solutions can be used to check the accuracy of approximate and numerical solutions of similar geometries.

Analytical solution to the problem of a plane electromagnetic wave scattering by a lossless homogeneous dielectric coated elliptic cylinder was investigated by many authors [1-2], and the solution is later extended to the nonconfocal dielectric case [3]. Axial slot antenna on a dielectric-coated elliptic cylinder was solved by Richmond [4]. Sebak obtained a solution to the problem of scattering from dielectric-coated impedance elliptic cylinder [5]. The scattering by mutilayered lossless dielectric elliptic cylinders was studied by many authors [6-8]. The scattering by a weakly lossy multilayer elliptic cylinder was obtained by Caorasi et al using a first order truncation of the Taylor expansion of each Mathieu function of real argument [9-10]. Recently, lossy and lossless metamaterials have gained considerable attention by many researches [11-14].

In this paper, numerical results of the electromagnetic wave scattering by a lossy metamaterial coated conducting elliptic cylinder is presented. These results are obtained for different configurations and compared with published data for the special case of a lossless conventional dielectric coated elliptic cylinder.

2. Numerical Results

Consider the case of a linearly polarized electromagnetic plane wave incident on a lossy metamaterial coated elliptic cylinder at an angle ϕ_i with respect to the x axis, as shown in Figure 1, with e^{jwt} time dependence. The complete formulation of this problem may be found in [16].

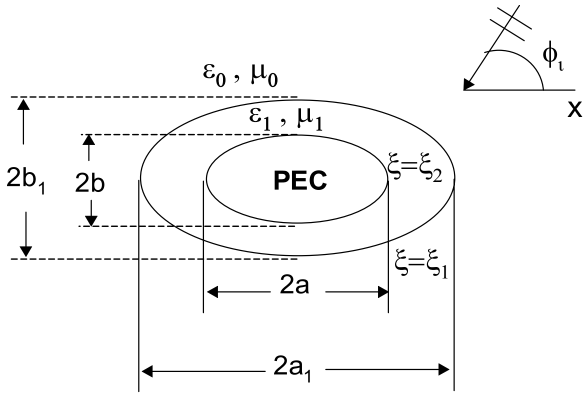

Fig. 1. Geometry of a lossy metamaterial coated elliptic cylinder.

Fig. 2 shows the echo width pattern for a dielectric coated elliptic cylinder against the scattering angle ϕ with an incident angle $\phi_i = 0^o$. The numerical results are plotted for the case of conventional dielectric, lossless metmaterial and lossy metamaterial coatings. The electrical dimensions of the elliptic cylinder are $k_o a = 2.50$, $k_o a_1 = 3.51$, $k_o b = 1.25$, and $k_o b_1 = 2.76$. The numerical results show a significant reduction in the echo width pattern especially at the scattering angles between 170 and 200 degrees due to the presence of the metamaterial, and the loss has little effect on the echo pattern in this range. The effect of the loss can be observed at the other scattering angles where the echo pattern decreases and the resonances vanish by increasing the loss of the metamaterial coating. Fig. 3 has the same electrical dimensions as in Fig. 2 except with $\phi_i = 90^o$.

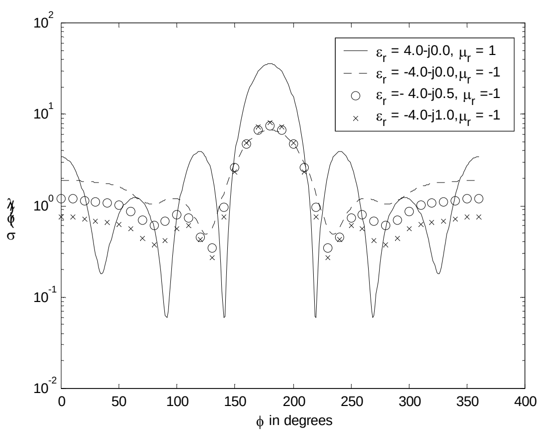

Fig. 2. Echo width pattern against the scattering angle ϕ of a lossy metamaterial coated elliptic cylinder with $k_o a = 2.5$, $k_o a_1 = 3.51$, $k_o b = 1.25$, and $k_o b_1 = 2.76$, and $\phi_i = 0^o$

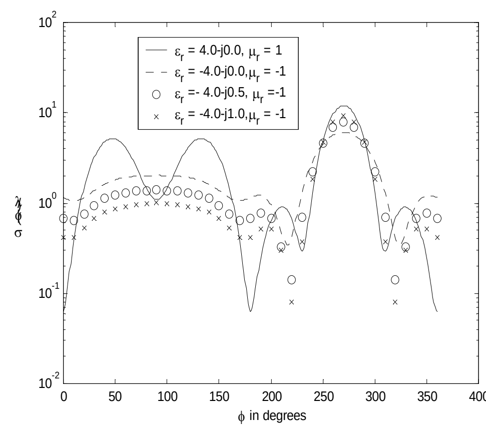

Fig. 3 Echo width pattern against the scattering angle ϕ of a lossy metamaterial coated elliptic cylinder with $k_o a = 2.5$, $k_o a_1 = 3.51$, $k_o b = 1.25$, and $k_o b_1 = 2.76$, and $\phi_i = 90^o$.

The effect of the metamaterial coating on the echo pattern at $\phi_i = 90^o$ is similar to that in Fig. 2. The echo pattern of a lossy metamaterial coated circular cylinder is also shown in Fig. 4 with $k_o a = 1.0$, $k_o a_1 = 2.0$, $k_o b = 1.0$, $k_o b_1 = 2.0$, and $\phi_i = 90^o$.

Fig. 5 shows the backscattering echo width pattern of a lossy metmaterial elliptic cylinder verses the incident angle ϕ_i. The electrical dimensions of the elliptic cylinder are $k_o a = 2.50$, $k_o a_1 = 3.08$, $k_o b = 0.63$, and $k_o b_1 = 1.88$. It can be seen that a significant reduction in the echo pattern occurs between 60 and 120 degrees due to the presence of the metamaterial coating and the reduction increases with an increase of the coating loss.

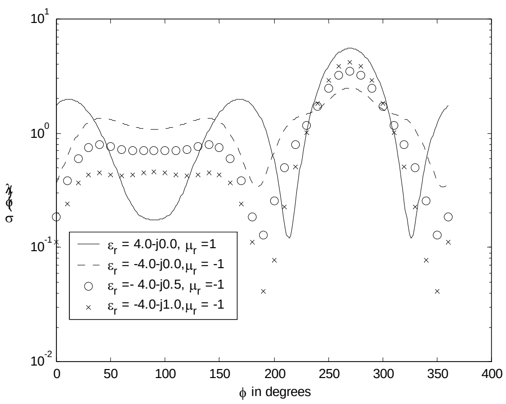

Fig. 4. Echo width pattern against the scattering angle ϕ of a lossy metamaterial coated circular cylinder with $k_o a = 1.0$, $k_o a_1 = 2.0$, $k_o b = 1.0$, and $k_o b_1 = 2.0$, and $\phi_i = 90^o$.

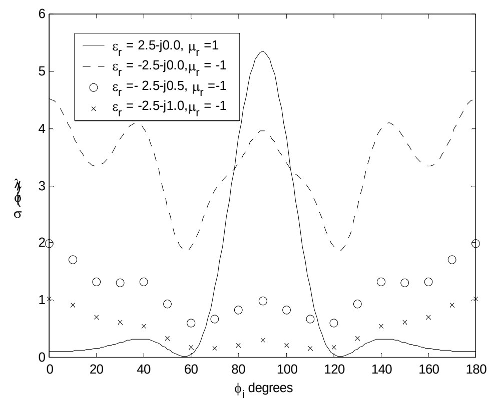

Fig. 5. Backscattering pattern against the incident angle ϕ_i of a lossy metamaterial coated elliptic cylinder with $k_o a = 2.5$, $k_o a_1 = 3.08$, $k_o b = 0.63$, and $k_o b_1 = 1.88$.

Fig. 6 shows the backscattering echo pattern of a lossy metamaterial elliptic cylinder versus the major axis of the dielectric coating (ka_1) with $k_o a = 0.63$, $k_o b = 0.50$ and $\phi_i = 0^o$. It can be seen that presence of the metamaterial coating reduces the number of resonances and a significant reduction in the echo pattern width also occurs at values of ka_1 greater than 2.0 with $\varepsilon_r = -4 - j1.0$. Fig. 7 is similar to Fig. 6 except for a lossy metamaterial coated circular cylinder with $k_o a = 0.63$. It can be seen that the echo pattern of a circular cylinder behaves similar to the elliptic cylinder with metamaterial coatings.

Fig. 8 shows the backscattering echo width pattern against $\varepsilon_r^{'}$ of a lossy and lossless metamaterial coated elliptic cylinder. The electrical dimensions of the elliptic cylinder are $k_o a = 1.25$, $k_o a_1 = 2.18$, $k_o b = 0.62$, $k_o b_1 = 1.88$ with $\phi_i = 0^o$. There is only one strong resonance occurs at small values of $\varepsilon_r^{'}$ for the case of lossels metamaterial coating and it vanishes due to the presence of the lossy coatings. Further, the loss has little effect on the echo pattern at larger values of $\varepsilon_r^{'}$, namely at $\varepsilon_r^{'}$ greater than -8.0. Fig. 9 is similar to Fig. 8 except for $\phi_i = 90^o$. It can be seen that changing the incident from 0 degree to 90 degrees has little effect on the echo pattern width.

3. CONCLUSIONS

Numerical results of the electromagnetic waves scattering by a lossy metamaterial coated circular or elliptic cylinder was obtained for the first time in the case of TM polarization. It was shown that a conducting cylinder coated with a lossy or lossless metamaterial has different backscattered cross section when compared to that coated with conventional dielectric material. The lossy and lossless metamaterials may be used to reduce the backscattered cross section over a certain lossy or lossless metamaterial coatings range.

ACKNOWLEDGEMENT

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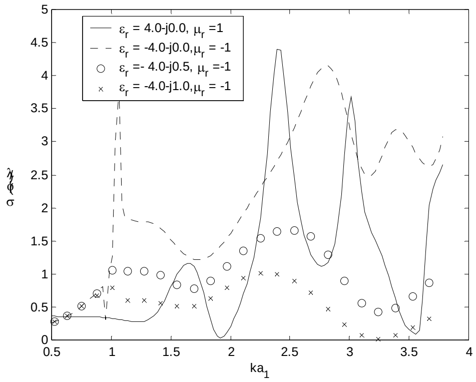

Fig. 6. Backscattering pattern against ka_1 of a lossy metamaterial coated elliptic cylinder with $k_o a = 0.63$, $k_o b = 0.50$, and $\phi_i = 0^o$.

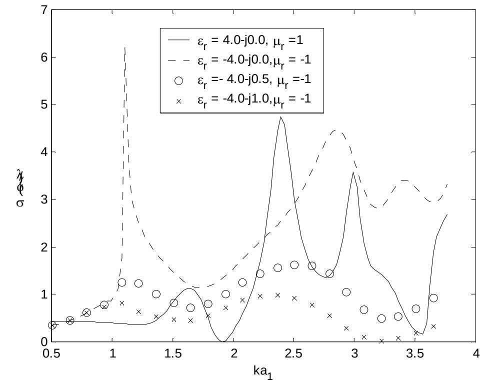

Fig. 7. Backscattering pattern against ka_1 of a lossy metamaterial coated circular cylinder with $k_o a = 0.63$ and $\phi_i = 0^o$.

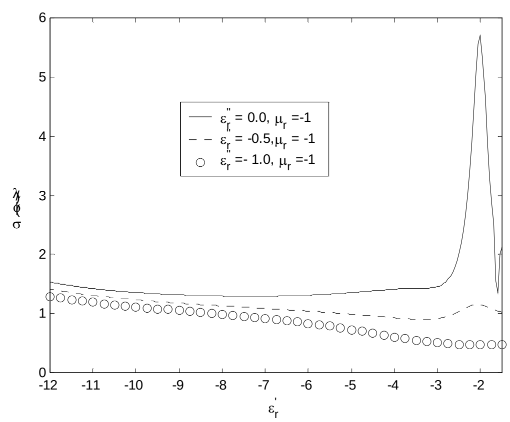

Fig. 8. Backscattering pattern against $\varepsilon_r^{'}$ for a lossy metamaterial coated elliptic cylinder with $k_o a = 1.25$, $k_o a_1 = 2.18$, $k_o b = 0.62$, $k_o b_1 = 1.88$ and $\phi_i = 0^o$.

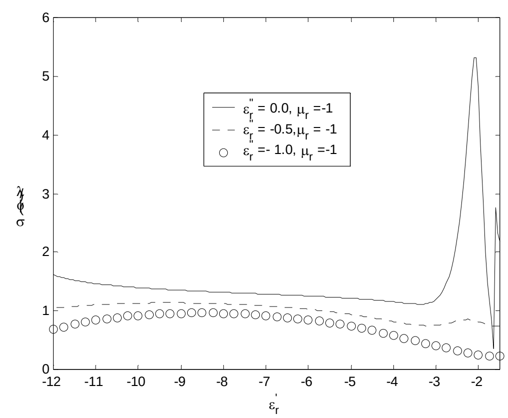

Fig. 9. Backscattering pattern against $\varepsilon_r^{'}$ for a lossy metamaterial coated elliptic cylinder with $k_o a = 1.25$, $k_o a_1 = 2.18$, $k_o b = 0.62$, $k_o b_1 = 1.88$ and $\phi_i = 90^o$.

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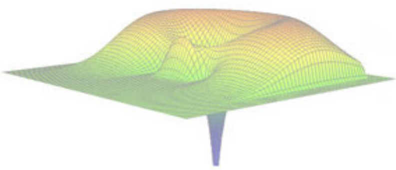

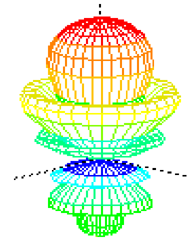

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