APPLIED COMPUTATIONAL ELECTROMAGNETICS SOCIETY (ACES)

NEWSLETTER

Vol. 10 No. 3 November 1995

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ACES NEWSLETTER STAFF

ACES NEWSLETTER AND JOURNAL COPY INFORMATION

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March	January 13
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For further information on the ***ACES JOURNAL,*** contact Prof. Duncan Baker at the above address.

For the ***ACES NEWSLETTER*** send copy to Ray Perez in the following formats:

1. A hardcopy.
2. Camera ready hardcopy of any figures.
3. If possible also send text on a floppy disk. We can read any version of MICROSOFT-WORD and ASCII files on both IBM and Macintosh disks. On IBM disks we can also read WORDPERFECT and WORDSTAR files. If any software other than MICROSOFT WORD has been used on Macintosh Disks, contact the Managing Editor, Richard W. Adler BEFORE submitting a diskette. If it is not possible to send a Macintosh disk then the hardcopy should be in Courier font **only** for scanning purposes.

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.

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The opinions, statements and facts contained in this Newsletter are solely the opinions of the authors and/or sources identified with each article. Articles with no author can be attributed to the editors or to the committee head in the case of committee reports. The United States recently became part of the Berne Copyright Convention. Under the Berne Convention, the copyright for an article in this newsletter is legally held by the author(s) of the article since no explicit copyright notice appears in the newsletter.

OFFICER'S REPORTS

PRESIDENT'S STATEMENT

I trust that you had a pleasant summer, and are now buckling down to the task at hand. In order to help you, I will be brief. Through diligent attention to cost cutting, and because of the success of our conferences, we are maintaining fiscal integrity. In order to ensure this strength, we want to promote membership in ACES as being beneficial to a career that involves computational electromagnetics. There are several ways that ACES can achieve its goal of being a leader in the CEM community: organize conferences that are valuable to this community, organize short-courses that are well attended, and provide user services, such as outstanding publications. We are doing well in all of these areas, but I do not want to see us slowing down in any of them.

You can help ACES by volunteering to promote a session in a conference, or presenting a short course, or, perhaps most easily, by writing a good paper for our publications, especially our Journal. Our members span a variety of backgrounds in CEM, and I believe that many of you have something worthwhile to tell the rest of us about your work. Please consider submitting a paper to our Journal editor, Duncan Baker. You will be helping ACES and the greater CEM community. If you are uncertain about your proposed subject matter, contact Duncan for suggestions and guidance. You'll be glad that you did, and so will we.

Best wishes,

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COMMITTEE REPORTS

ACES PUBLICATIONS

Two issues of the ACES Journal and ACES Newsletter were produced on schedule earlier in 1995, and this report appears in the third issue of the Newsletter. Vol. 10, No. 1 of the ACES Journal (March 1995) had 76 pages and was comprised of contributed papers. Vol. 10, No. 1 of the ACES Newsletter had 104 pages. Vol. 10, No. 2 of the ACES Journal (July 1995) had 120 pages, and was also comprised entirely of contributed papers. Vol. 10, No. 2 of the ACES Newsletter had 101 pages. In summary, the new "department" area approach to the Newsletter implemented by Editor-in-Chief Ray Perez has clearly boosted contributions to the Newsletter in 1995. For this year, page count in the Journal was reduced somewhat as more pages were utilized in the Newsletter. Overall, the financial resource requirements of ACES Publications in 1995 have closely tracked the projections made in 1994. We have produced slightly more total pages than originally projected, but our printing costs (per page) have decreased so that expenses continue to compare favorably with budgeted allocations.

We gratefully acknowledge the many hours of excellent work rendered by Guest Editors Allen Glisson and Ahmed Kishk on Vol. 10, No. 3 of the ACES Journal, which you have received bundled with this Newsletter. This Special Issue of the ACES Journal is particularly timely and important, and we believe it is highly responsive to the information needs of the ACES membership. If you find material in this Special Issue to be useful to you in your work, solving applied problems using Computational Electromagnetics, both the Guest Editors and ACES Publications would be pleased to hear your comments.

The new Style guidelines developed by Duncan Baker for the ACES Journal have now been in effect for five issues, with very positive results. In addition to conserving space, the new format guidelines give the Journal a solid, professional appearance which has drawn numerous compliments.

We are tracking the evolution toward eventual maturity of electronic publication and distribution of technical material. The present position of ACES Publications is that traditional publication remains the most practical vehicle for us. However, we recognize that this is subject to change, perhaps within the next two to three years. Comments from ACES members with special interest and/or expertise in the area of electronic publishing are both welcome and invited. If you believe that we should be addressing this issue more aggressively now, please share your knowledge and views with a member of the ACES Publications Committee at your earliest convenience.

Finally, ACES Publications is considering how we can increase technical cooperation with other societies involved with aspects of Computational Electromagnetics that are kindred to the scope of ACES, such as the Compumag Society. If you have any ideas which would improve the dissemination of useful and important technical information to the ACES membership, please let us hear from you.

W. Perry Wheless, Jr.
ACES Editor-in-Chief / Publication Chair
e-mail wwheless@ua1vm.ua.edu (note: '1' = numeral one)

SOFTWARE EXCHANGE REPORT

ACES has recently received several diskettes that have been offered for distribution to our members. Some of them are restricted.

1. NECSHELL (Beta Version)

Three CEM experts from Lvov University in the Ukraine, Michael M. Mikhailov, Victor O. Lomtev and Andrew A. Efanov, have developed a WINDOWS replacement for the IGUANA 5.4 module of NEEDS 2.0. NECSHELL contains a duplicate version of the NEEDS NEC program NEC-81. It runs under Microsoft's POWER STATION FORTRAN. NECSHELL replaces the input deck preparation, NEC-81 running and output file viewing portions of IGUANA. In addition, viewing of wire structures is available, with rotation, zooming and coordinate point readout. The operation of the viewing portion of IGUANA makes it possible to print the structure views on any Windows-supported printer. A VGA monitor is required, and Windows 3.1 or newer. Disk storage used by NECSHELL is 3 MB. The NX card of NEC-81 is not supported in this version of NECSHELL. Surface patches are shown as hexagons. By operating a split screen showing input cards and the object view, side-by-side, one can identify any wire by clicking on it in the view window, and the corresponding card(s) will be highlighted in the input card window. All geometry cards have dialog boxes to assist the beginning NEC user, as do many of the control cards. Since this code is still in beta version, ACES will send it gratis to any member who will exercise it and report bugs. Upon completion, NECSHELL Version 1 will be distributed free to all registered NEEDS 2.0 users. Planned future releases of NECSHELL will provide output plots of radiation patterns, impedance and currents. Rectangular, polar, contour and 3-D color plots are expected.

2. NEC4.1 (PC Versions)

NEC4.1 is now available via Lawrence Livermore National Labs (Jerry Burke) to any U.S.A. user. Jerry distributes an IBM RISC 6000 UNIX version and a MAC version. ACES is offering several PC versions at no charge to anyone who is registered with LLNL. The versions available from ACES are:

a. DNEC4.1-LH for the Lahey Fortran Ver. 5.2 F77L-EM/32 compiler. Source and executable files, sample outputs and a make file are included. (Courtesy of Richard W. Adler.)

b. DNEC4.1-MS for the Microsoft POWER STATION 32-bit Fortran compilier. Same files as (a.) (Courtesy of David J. Pinion.)

c. NEC4.1S-WT for the WATCOM Ver. 9.5 F77 32-bit Fortran compilier. Executables and source code for 486 and 586 (Pentium) processors are supplied for Windows NT/95. (Courtesy of Richard Albus.)

Richard Adler for the Software Exchange Committee.

NOMINATIONS

In the coming months, ACES members will be asked to vote for three board members. For uniformity, each candidate will be asked to provide a short statement that addresses:

(1) GENERAL BACKGROUND (e.g., professional experience, degrees, employment, etc).

(2) PAST SERVICE TO ACES (e.g., service on ACES committees, or other contributions).

(3) CANDIDATES' STATEMENTS (e.g., short statement of the candidates views of major issues relevant to ACES). Candidates' statements will be no more than 500 words, unless otherwise directed by the Board.

(4) OTHER UNIQUE QUALIFICATIONS (An additional but optional statement).

It is hoped that these areas will provide data on each candidate that might otherwise be obscured in a general, unstructured statement. When the time comes, please take a few minutes to study the candidates' statements and vote.

The names and terms of the present members of the Board are as follows:

Ray J. Luebbers	1996	Duncan C. Baker	1997	Pat Foster	1998
Harold A. Sabbagh	1996	Edmund K. Miller	1997	Todd Hubing	1998
W, Perry Wheless, Jr.	1996	Andrew F. Peterson	1997	Adalbert Konrad	1998

Adalbert Konrad
Chairman

ACES UK CHAPTER REPORT

ACES UK activities during this half year have been highly successful. Most recently we had our one day meeting, which included a half day code validation short course from Ed Miller. This provoked a lively discussion from the 25 attendees. The afternoon was given over to seven short papers from a variety of code users and developers. Scattering was the predominant theme.

Our AGM, held to coincide with this one day meeting, welcomed two new committee members, Greg Cook (University of Sheffield) and Tony Monk (Queen Mary and Westfield College, University of London). Brian Austin retired after many years of support and Jeff Cox moved from the committee to the newly formed Newsletter sub-committee. Jeff has agreed to remain Newsletter Editor.

In July ACES UK pioneered a NEC one day course. This combined hands-on and theoretical day was over subscribed (16 attended) and attracted users of a wide range of abilities with lectures from leading NEC experts in the UK. Useful feedback from attendees indicated a recommendation for a repeat in two years time. This ACES UK initiative was co-sponsored by the IEEE UK/RI Chapter, a first link between the two Chapters. Thanks are due in particular to Pat Foster for the organisational effort behind the day.

Finally, the UK Newsletter was the largest we have produced, particularly by including the short form papers from last years meeting and largely due to our editor, Jeff Cox, and we hope to repeat the success in 1996.

Tony Brown
Chairman
ACES UK Chapter

CEM NEWS FROM EUROPE

Coordinated by Pat Foster, MAAS, UK[1]

There are two inputs for this News from Europe column. One is from the National Radiological Protection Board of the UK where Simon Mann and Peter Dimbylow of the Non-Ionising Radiation Department have much work of interest in CEM underway. The other is from Christian Hafner of the Institute fur Feldtheorie, Zurich who has contributed so much to work on Multiple Multipole Theory.

National Radiological Protection Board (NRPB)

The National Radiological Protection Board (NRPB) is an independent statutory body in the UK, which was set up by the Radiological Protection Act: 1970. Its responsibilities are to carry out research in order to enhance knowledge about the protection from radiation hazards and to provide information and advice to government departments and other concerned organisations and individuals. The majority of board staff are concerned with protection from ionising radiation; however increasingly in recent years activities concerned with non-ionising radiation - and in particular electromagnetic fields - have grown substantially.

The major application of Computational Electromagnetics (CEM) is in the area of dosimetry, where the techniques employed are intended to provide the link between external measurement quantities and dose quantities in the body. In electromagnetic dosimetry the external quantities are electric and magnetic fields strengths and the dose quantities are induced electric field strength, current, current density, SAR (Specific Absorption Rate) and temperature rise.

Restrictions on current density and SAR established from biological consideration form the basis of NRPB exposure guidelines and derived field strengths, termed investigation levels, provide a framework for assessing compliance. At low frequencies, quasi-static approximations can be used to determine investigation levels for electric and magnetic fields separately, while at higher frequencies a plane wave is assumed. Where an exposed individual is sufficiently close to an RF source it may be necessary to carry out source-specific modelling work to show compliance directly with the internal dose quantities.

A program to develop anatomically realistic numerical phantoms from greyscale Magnetic Resonance Imaging (MRI) scan images has been on going at NRPB. This has produced models of the human head in which the tissues are identified at a resolution of 1mm, and more recently a model of the complete body. Published values for the conductivities and permittivities of the tissues are fed into a Finite Difference Time Domain (FDTD) code which is used to calculate the SARs and induced currents corresponding to a variety of exposure scenarios, including plane-waves and a short dipole close to the head. The plane-wave exposure condition is used to develop investigation levels for compliance with the basic restrictions.

More recently, models of cellular telephone handsets have been coupled to the head model in order to deduce the relationship between radiated powers and SARs produced. Consideration of a variety of exposure geometries has yielded the results that in order to restrict the SAR to $10\mathrm{Wkg}^{-1}$ in any 10g of tissue during normal operation, handsets should not radiate RMS powers greater than 3.2W at 900MHz at 1.8GHz. FDTD predictions of the near-field from handsets in the absence of a head have been compared with predictions from wire-grid modeling using NEC and have shown agreement to within a few percent.

Apart from dosimetric calculations, computer codes are used to model sources of electromagnetic fields and assess their hazard potential by defining, where necessary, safe distances. For broadcast antennas, the NEC computer code has been found to be very useful and agrees well with measurements, where they are available. An aperture modelling code has been produced which is used to model reflector antennas such as those used with microwave links and radar installations. This code replaces the aperture with a current sheet and then performs an integration to find the field at points in front of the antenna.

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There is much interest in the radiation from cellular telephone base stations at present and a simple line-source modelling program is used to analyse omni-antennas consisting of vertical dipole arrays. It is intended to use NEC to model dipoles in corner reflectors in order to analyse sector antennas in the near future. The final use of CEM at NRPB is in the characterisation of our probe calibration facilities. Codes have been written to calculate electric fields inside parallel plate test cells and also to derive corrections for the calibrations of large meters in such cells. Finally a code has been written to calculate the magnetic fields produced by our Helmholtz coils.

Much of the RF dosimetry used in human exposure guidelines still employs models which predate the advent of practical FDTD computations and the developments of anatomical phantoms. Thus the main CEM task of NRPB today is the development of RF protection standards based entirely upon accurate modelling of anatomical human phantoms.

Dr. Simon Mann can be contacted on FAX (44) 1235 833891.

Activities and Trends of the Electromagnetics Group at the ETH, Zurich

The Multiple Multipole Program (MMP) is very closely related to analytical procedures. Therefore, MMP allows us to obtain extremely accurate and reliable results. It has initially been developed and used to clarify some theoretical questions that cannot be answered by analytical reasoning. MMP is based on a pure boundary method, the Generalized Multipole Technique (GMT) and therefore is especially advantageous and efficient for modelling lossy media and for near-field computations. This has allowed us to obtain the first useful simulations of a handheld radio near a human body. In addition to problems in bioelectromagnetics, MMP has been applied to EMC, antenna design, microwave applications, high voltage problems, etc. MMP has been tested over an extremely large frequency range from statics up to optics. Its ability to model small (quasi-static) details in large structures are of special interest in near-field optics and other topics of nano science. This explains the success of MMP simulations of Scanning Nearfield Optical Microscopes (SNOM).

For a long time, the electromagnetics group suffered from relatively poor equipment. The original MMP codes ran on machines comparable with XT compatible PCs. Therefore, implementations of PC versions were very easy. In 1990 and 1993, 2D and 3D MMP for PCs, were published as book-software packages for computational electromagnetics. Although the version for 3D MMP includes a GUI with excellent graphic field representations (including animation), it suffers from a lack of user friendliness in the modelling due to several facts; generality of the implementation of the kernel, the wide range of applications, the large number of special features for increasing the efficiency and also the small number of members of the electromagnetics group at the ETH working on the implementation of MMP, etc. The structure of the kernel and of the GUI is now being entirely redesigned. The 3D MMP release planned for 1996 should require much less knowledge from the user. It should include automatic routines for generating MMP expansions and features for adaptive modelling. Moreover, the new 3D MMP version includes an eigenvalue solver for computing resonators and guided waves and special features for easily computing periodic structures.

Since fast PCs are quite adequate for obtaining MMP solutions of many interesting EM problems, MMP can be used as a forward problem solver for inverse scattering and in EM optimization problems. This requires the robust, accurate and efficient computation of many similar problems. Therefore, new techniques are being developed for tuning MMP. A strategy for drastically reducing the computation time is the Parameter Estimation Technique (PET). The PET takes advantage of the knowledge obtained in previous runs in order to achieve an excellent guess of the parameters to be computed in the actual run. Moreover, the PET and several parts of the MMP kernel (for example, the automatic procedures mentioned above) contain various parameters that heavily affect the performance of MMP. These system parameters are provided by the user. Although useful default values are known, this is the main reason why experienced MMP users are much more successful than others. The automatic optimization of the system parameters in such a way that typically applications of a specific user will be computed most efficiently, will be a key to user friendliness and to the efficiency of the future MMP release. Therefore, the focus of our current research is on the development of efficient optimization strategies for complicated tasks based on evolutionary ideas. The future MMP code should be able to compute inverse scattering and optimization tasks and it should also be self-optimizing, i.e. it should adapt its system parameters to the needs of a specific user.

Dr. Hafner is available on FAX (41) 1632 1198. Email:hafner@ifh.ee.ethz.ch. More information is available on the World Wide Web at www.ifh.ee.ethz.ch.

COMPUTATIONAL ELECTROMAGNETICS ON THE INTERNET

What are the dates for the next ACES symposium? How do I register? When are the final papers due? How do I submit a paper to the ACES Journal? If you have questions about ACES or upcoming ACES events, check out the ACES world wide web page at http://www.emclab.umr.edu/aces. The web page contains up-to-date information on the ACES organization, activities, dates, deadlines, etc....

If you have a question about ACES that is not answered by the web page, send an email message to Todd Hubing (thubing@ee.umr.edu). He will try to answer your question and post the new information to the web page. Also, if you have information of your own that you feel should be posted to the web page, send it to Todd. Committee reports, conference or short course announcements, new software announcements, and other news items of interest to the CEM community can be posted to the web.

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MODELER'S NOTES

Gerald J. Burke

There were no contributed articles for Modeler's Notes for this issue, but readers have provided new information on the NEC running times on PCs to add to the discussion in previous issues. Also, a couple of bugs have been uncovered by NEC–4 users, and fixes are given below. If anyone has any ideas for articles on modeling that they would like to write they would be very welcome here, although I guess it is up to the editor to come up with some more specific guidelines and maybe start twisting some arms. If anyone has comments on material covered here, suggestions or observations on using NEC or other modeling codes or can submit an article on EM modeling topics they are encouraged to submit them to

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NEC–4 was first released about three years ago and has now been sent out to over 100 sites, so it should be of interest to a reasonable number of ACES readers, although not approaching the number of NEC-2 users. A couple of bugs in NEC-4 have come to light recently through user's experiences. Brent Campbell found that the maximum coupling calculation (CP command) may give the wrong result for a structure that includes transmission lines. It works correctly with transmission lines as long as another command (XQ, NE, RP, etc.) precedes it to initiate the solution for currents, which is the reason that the problem slipped by pre-release testing. If CP is the first command to initiate the solution the transmission line parameters do not get set up. Note that in NEC–2 the CP command does not initiate the solution but must be followed by an EX-XQ sequence, and the NEC–2 CP command does not give the correct maximum coupling for structures with networks or transmission lines. NEC–4 should give the correct result with transmission lines. The problem that Brent encountered can be fixed by adding a call to the subroutine SETTLX just before the call to subroutine COUPLE at about line 347 of the main program. Note that there are two calls to subroutine COUPLE in the NEC–4 main, and this is the first one. The inserted line and preceding and following lines are shown below:

```
C
C     COMPUTE MAXIMUM COUPLING AND RETURN TO READING INPUT
C
      IF(NONET.GT.0)CALL SETTLX              ! insert
      CALL COUPLE(ICPT1,ICPS1,ICPT2,ICPS2)
```

Another problem was found by Roy Lewallen when he tried running a "MININEC ground", using a perfectly conducting ground under the antenna and real ground for the radiated field. This can be done in NEC–4 by specifying a perfectly conducing ground

(GN1) and using the GD command to define the parameters of a second medium beyond a specified line (the "cliff"). If the radius of the cliff is set to zero the primary ground (perfectly conducting) will still be used in the solution for currents, and the second-medium parameters will be used in computing radiated field. However, Roy found that doing this in NEC–4 resulted in a division by zero, because the second-medium parameters did not get set when the primary ground was perfectly conducting. This problem resulted from code in subroutine SETGND. There is another problem there, that when upper-medium parameters are set (UM command) with a perfectly conducting round there is no indication of the upper-medium parameters in the output, although they do take effect. Both of these problems can be fixed with the changes shown below to the code in subroutine SETGND.

```
C
C   SET FREQUENCY DEPENDENT GROUND PARAMETERS
C
      FRATI=(1.,0.)
      IF (IPERF.EQ.1)THEN
        WRITE(3,906)
                                          ! delete RETURN
      ELSE                                ! change
        IF(GSIG.LT.0.)GSIG=-GSIG*TP*FMHZ*1.E6*EPSRZ
        EPSC1=CMPLX(GEPS,-GSIG/(TP*FMHZ*1.E6*EPSRZ))
        CEPSL=EPSC1
        XK=CSQRT(EPSC1)
        XKL=TP*FMHZ*1.E6*XK/CVEL
        GCK1=TP*XK
        GCK1SQ=GCK1*GCK1
        CK1=GCK1
        ETAL=ETAZ/XK
      END IF                              ! insert
      IF(ICLIFT.NE.0)THEN
         IF(SIG2.LT.0.)SIG2=-SIG2*TP*FMHZ*1.E6*EPSRZ
         EPSC2=CMPLX(EPSR2,-SIG2/(TP*FMHZ*1.E6*EPSRZ))
         XK=CSQRT(EPSC2)
         ETAL2=ETAZ/XK
      END IF
      IF(IPERF.NE.1)THEN                  ! insert
        IF (NRADL.GT.0)THEN
          WRITE(3,907) NRADL,SCNRAD,SCNWRD
          WRITE(3,908)
        END IF
        IF(IPERF.EQ.2)THEN
          FRATI=(XKL*XKL-XKU*XKU)/(XKL*XKL+XKU*XKU)
          CALL GNDINO(EPSC1,21)
          WRITE(3,909)
        ELSE
          WRITE(3,910)
        END IF
        WRITE(3,912) GEPS,GSIG,EPSC1
      END IF                              ! insert
      IF(CABS(CEPSU-1.).GT.1.E-6)THEN
      ...
```

Roy also has been testing the Sommerfeld ground model for high ground conductivity to compare with perfectly-conducting ground, and has found differences between NEC–4 and NEC–2. Actually, the Sommerfeld ground model in NEC–4 is essentially the same as that in NEC–3, but it differs from NEC–2. Both NEC–2 and NEC–3/4 generate tables of the Sommerfeld integral values for the field over ground and interpolate for values needed to fill the impedance matrix. The NEC–3/4 code also uses functions from the asymptotic form of the field over ground that are fit to the computed values and used to interpolate in a scheme known as Model-Based Parameter Estimation. All of these evaluations have been adjusted to work for the typical range of ground parameters encountered in antenna installations, so you can get into trouble with values far out of this range.

NEC–2 uses three interpolation grids with fixed limits, as shown in Figure 12 of the NEC–2 Theory Manual (Part I). In the standard configuration, with the limit on the distance parameter for grid 1 of $R_1/\lambda_0 = 0.2$, the NEC–2 solution fails for a horizontal wire approaching the surface of seawater ($\sigma = 4$ S/m) at 1 MHz. To get accurate results for this case the parameters of the interpolation grid must be adjusted. To handle seawater at 1 MHz the maximum R_1/λ_0 for grid 1 could be decreased from the standard 0.2 to 0.004 which is about five times the skin depth/λ_0. This is done by reducing DXA(1), the increment for R_1/λ_0 in grid 1, from 0.02 to 0.0004. Then with the number of R_1 samples in the grid, NXA(1), equal to 11, the final value in grid 1 is (NXA(1)-1)*DXA(1) = 0.004. The starting values for grids 2 and 3, XSA(2) and XSA(3), must be set to 0.004 to avoid leaving a gap between the grids. These changes in increment and starting values are made in program SOMNEC in the BLOCK DATA SOMSET, or in older versions in the DATA statements in the SOMNEC main program. If you want to increase the number of values in the grids to increase the overall interpolation accuracy, you must increase the array dimensions in SOMNEC and also in the corresponding COMMON block in NEC–2. Some information for making these changes can be found on page 392 of the NEC–2 Code Manual (Part II). Obviously, changing the grid parameters is not as easy as it could be, but it can be done if you are careful to check that things are still working afterward.

In NEC–3 and 4 the interpolation is more complicated, since it involves a three dimensional space to allow for wires below and above the ground surface. However, the adjustment of the grid boundaries is done automatically for the frequency and ground parameters, so NEC–3 and 4 should give accurate results for a wire approaching seawater at 1 MHz. The problem that Roy Lewallen ran into was with a radial-wire ground screen floating just above the ground which is a known potential source of trouble. One requirement for a screen above ground is that the lengths of segments at the junction be equal to or less than the height of the screen, which is usually accomplished by tapering the segment lengths. However, despite following this rule, Roy found that NEC–4 (and NEC–3) gave incorrect input impedances for ground conductivities of about 10 S/m or greater, while NEC–2 gave a result close to that for the perfectly conducting ground. At this point I have not had a chance to look into the cause of this problem, but can just recommend that, unless you are doing validation tests, the perfect ground (GN1) option should be used in such cases of high ground conductivity, both for accuracy and speed.

Back on the subject of PC running times, we have some new data provided by John Grebenkemper for PowerMacs and Roy Lewallen for a Pentium. For a direct comparison with previous results, John ran NEC–2 on a PowerMac 7100/66, with 66 MHz PowerPC processor. The NEC–2 codes had been compiled without the -N9 option in the Absoft compiler to obtain maximum speed without the option for interrupting. Running times for the 300, 600 and 1200 segment test cases used previously are shown in Table 1.

Table 1. Times in seconds for NEC2S/D to fill and factor the impedance matrix. Total is the time printed at the end of the output and included computation of currents but no radiated fields.

Computer	Prec.	No. Seg.	Fill	Factor	Total
PowerMac 7100/66	S	300	7.4	3.0	11.9
PowerMac 7100/66	D	300	8.0	3.7	13.4
PowerMac 7100/66	S	600	27.5	24.8	55.0
PowerMac 7100/66	D	600	29.0	30.0	61.8
PowerMac 7100/66	S	1200	85.7	196.5	289.0
PowerMac 7100/66	D	1200	90.8	257.0	354.6

The 7100 had a 256 KB secondary cache and 32 MB of main memory. Compared to these times, the Quadra 650 with PPC card running at the same clock speed (see the July Newsletter) is about 19 percent slower in filling the matrix and 40 to 50 percent slower in factoring. This difference may be due to the bottleneck of transferring data from the card to the memory on the motherboard. John has now traded up to a PowerMac 7500/100, and has provided the times in Table 2 for the test cases run with NEC–4. The 7500 has a PowerPC 601 running at 100 MHz and a 128 bit wide memory bus, which should improve its speed, but it did not have a secondary cache installed. The 1200 segment double precision case is omitted since the 7500 had only 16 MB of RAM. The total times for the same code on the 7100/66 and the percent slower than the 7500/100 are also shown. The NEC–4 codes had been compiled with the -N9 option for interrupt checking at 1/6 second intervals which slows them down by 10 to 20 percent.

Table 2. Times in seconds for NEC4S/D to fill and factor the impedance matrix.

Computer	Prec.	No. Seg.	Fill	Factor	Total	7100/66	% slower
PowerMac 7500/100	S	300	6.6	2.5	10.2	15.2	49%
PowerMac 7500/100	D	300	6.4	3.1	10.5	15.9	51%
PowerMac 7500/100	S	600	25.4	20.2	48.7	70.4	45%
PowerMac 7500/100	D	600	30.0	24.2	59.1	76.3	29%
PowerMac 7500/100	S	1200	95.1	156.6	263.0	384.8	46%
Pentium, 90 MHz	S	300	8.1	4.7	13.7		
Pentium, 90 MHz	D	300	10.3	8.5	19.7		
Pentium, 90 MHz	S	600	31.0	39.5	72.8		
Pentium, 90 MHz	D	600	39.6	69.8	112.1		
Pentium, 90 MHz	S	1200	132.3	546.4	705.9		
Pentium, 90 MHz	D	1200	179.2	928.1	1157.2		

Roy Lewallen has provided running times for the same test cases on a 90 MHz Pentium processor, and the results are included in Table 2. He had 16 MB of RAM configured for 1 MB disk cache, and compiled NEC–4 with the Microsoft FORTRAN PowerStation, v. 1.00a with Release, 486 Code, and Optimize Time options. He said that the 1200 segment single and double precision cases used the NEC out-of-core solution and were further slowed down by fragmented disk space. He plans to set it up for dynamic allocation of matrix memory as in EZNEC, but had not had time to do that yet or to try shoehorning it in. The 1200 segment single precision case fits into a Mac with 16 MB of RAM, but it is tight and would depend on system size and other RAM users, as well as the size of the code generated by the compiler. The Pentium seems to be a little slower than the Mac 7500, adjusted for clock speed. It is considerably slower in double precision, but that may depend on the floating point unit.

At the deadline for the last Newsletter I was just starting to use the Absoft FORTRAN compiler version 4.1 that replaced version 4.0 for the PowerMac. Version 4.1 was claimed to be up to five times faster in compiling code with many small routines, which would apply to parts of NEC–4, and to include additional optimizations. Initial tests were somewhat disappointing. While it was faster in compiling small routines, the time to compile subroutine DATAGN in NEC–4 went from 43 seconds with v. 4.0 to 2970 seconds with v. 4.1. Also, the code was about 25 percent slower in factoring the matrix, but about the same in filling. I sent some sample code to Absoft for evaluation, but they have declined to comment. However, it was then found that the double precision code compiled with v. 4.0 for 1200 segments produced all zeros due to what looks like a linking error. Version 4.1 has compiled this and a 2500 segment code correctly, so we are forced to use the new version. At least v. 4.1 for the PowerMac and v. 3.4 for 68k Macs are set up to coexist in the same MPW folder. Previous versions would not do this unless you modified some file names and script files. The new versions still would not allow access to the makefile builders for both 68k and PowerMacs until I got a new script file from Absoft.

Some information on Mac FORTRAN was put on nec-list@ee.ubc.CA by Bernie Bernstein. He reports that the Language Systems FORTRAN compiler for the PowerMac compiles 4 to 5 times faster than the Absoft compiler and requires much less memory. However, the code that it produces is not as fast as that from the Absoft compiler. He also reports that the Motorola FORTRAN compiler has been released. He has not tried it, but said that it apparently does not support the Mac toolbox. That would probably be OK for NEC, since the toolbox is only used for dialog boxes for opening files. However, we did get plotting routines working with the toolbox (with help from Absoft) and they are impressively fast on the PowerMacs.

COMPUTATIONAL ELECTROMAGNETICS USING THE RESEARCH AND ENGINEERING FRAMEWORK AS A BACKBONE

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COMPUTATIONAL ELECTROMAGNETICS (CEM)

In today's application of this technology numerous formulations are implemented in stand-alone codes. Such formulations include the method of moments, the uniform theory of diffraction, and finite elements and finite differences. Furthermore, individual application areas (e.g., antenna/platform integration, radar cross section) quite often have developed their own set of duplicative stand-alone codes implementing one or more of the various CEM formulations. What results is a proliferation of codes, such as is shown in figure 1.

There are many disadvantages to this situation. These include:

- Increased maintenance requirements since they must be accomplished for each individual code;
- Difficulty in generating an electromagnetic model which spans more than one technique because of the significantly different input data languages for each code;
- Extensive learning curves because of the unique nature of each code, even for codes which implement the same CEM formulation or which are used for the same application area; and
- Difficulty in incorporating new technology (physics, mathematics, computer science) into existing codes because most codes were not developed using current software practices, or they do not have adequate documentation to facilitate the insertion or replacement of new code.

As a consequence, the CEM code state-of-the-art unnecessarily lags behind what is available from the scientific research communities, and it lags behind that which is needed to model and analyze the emerging DoD and consumer product system and device technology.

A study of figure 1 reveals that many coding similarities can be exploited and, by doing so the CEM community will be able to do its job more smartly. All analyses begin with a description of the physical geometry of the system to be analyzed. This can be a computer file, drawings, or measurements made on the structure and entered into a "database" of some sort.

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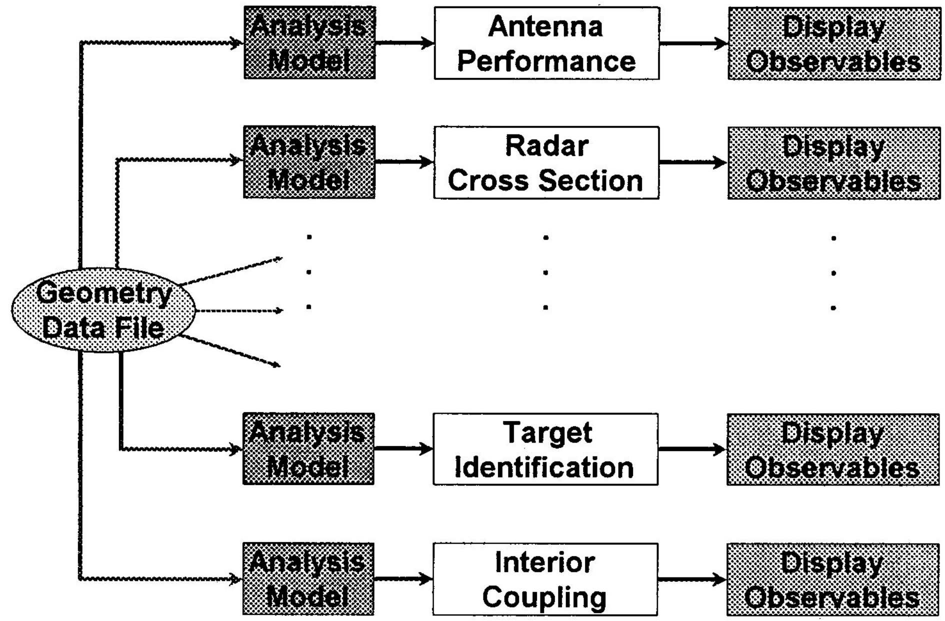

Figure 1. Current CEM Application Practice

All codes have an input function which translates the physical geometry into an electromagnetic model which is understandable to the computational engine. Almost always it is an exercise of the imagination and insight of the analyst to manually develop the model. This is done using some graphics-based tool, of which there are nearly as many as there are CEM codes, thereby doubling the impact of the disadvantages mentioned above.

Once the analysis has been performed using whatever is the computational engine of choice, the output data is reduced and presented in various forms for study. These include tables, plots, histograms, mappings on the structure surface, and many other specialized forms more suited to the problem at hand. A variety of packages, both in-house developed and commercial off-the-shelf (COTS), is available to accomplish this data presentation.

Is there not a more efficient way in which to configure CEM codes and perform CEM analyses such that the inefficiencies are reduced?

THE ELECTROMAGNETIC MODELING AND SIMULATION ENVIRONMENT FOR SYSTEMS (EMSES)

Figure 2 graphically depicts in a simplified form a proposed workspace in which CEM codes can be more efficiently used, maintained, and enhanced. In this scheme the functions of geometry database access, electromagnetic model generation and display, and output data reduction and display are resident in modules that are common within some computational framework. The only codes that are unique are those associated with the computational engine.

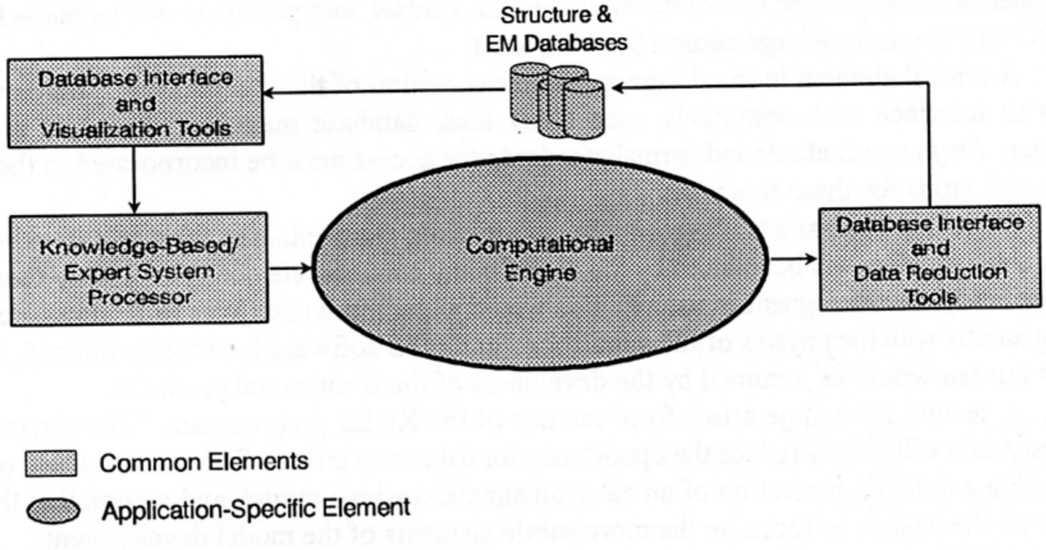

Figure 2. EMSES Concept

It should be noted that the block in figure 1 which indicates the development of the analysis model has been divided into two separate functions in figure 2. The first operation deals with the interface to the database (as available to the analyst, appropriate to the device under analysis, and not determined by or dependent on the CEM computational environment) and the visualization of the geometry on a graphics display device. Then, this is operated on by the analyst using a knowledge-based, expert system (KBES) preprocessor. This preprocessor would generate a first approximation to the analysis model based upon its rule base, the geometry being considered, and information pertinent to the analysis. The analyst would have the option of refining this tentative model to more closely reflect the requirements of the particular analysis and in light of the analyst's previous modeling experience.

Once the analyst has determined that the model satisfies the requirements, the preprocessor then generates the necessary input data file in the language appropriate to the computational engine being used. It is anticipated that each computational engine would be wrappered by a preprocessor and a postprocessor. These are nothing more than translators or filters which allow the KBES to output a standard language file and the data reduction tool package to input a standard format for subsequent processing and display. Wrappers will be developed in accordance with the computational environment's specifications, ideally using software tools provided in the environment.

The other major common element in figure 2 is the postprocessing package which accomplishes the interface with the system electromagnetic database and provides the suite of tools which accomplish data reduction and display in a variety of styles and formats. Several existing

CEM-specific packages can be interfaced within this toolset, and provision will be made for the use of a COTS software package chosen by the analyst.

A critical element in the design and implementation of the computational framework is the ability to interface with commonly used CAD files, database managers, and data presentation packages. *De facto* standards and formal standards for access must be incorporated in the design of framework stubs for these functions.

There are several advantages to be gained from the implementation of this scheme. The maintenance problem will be reduced by the use of many common elements, especially those that are not specifically electromagnetic in nature. This would allow more resources to be focussed on issues dealing strictly with the physics of the simulation. If COTS software is properly utilized, then much of that burden would be assumed by the developers of the commercial products.

A second advantage arises from the use of the KBES preprocessor. The direct access of geometry data will greatly reduce the opportunity for data entry errors. The extensive use of rule sets will enable a quicker generation of an electromagnetic analysis model, and automating the process will allow the analyst to focus on the more subtle elements of the model development.

A third advantage is the benefit to those who are developing new codes. They will not need to be concerned with the design and generation of code to perform the functions which would be performed by these "standard" packages. Once again, they will be able to concentrate their efforts on the code which implements the CEM formulations. If additional capabilities need to be added to the pre- and post-processors, then they can be incorporated by those performing the support function for these elements of the computational framework.

There are also disadvantages associated with this approach. The primary one is that it really does not address fully the issue of the code disadvantages enumerated earlier. To address this broad area one must work toward a full implementation of the concept of a library of electromagnetic processes as described in other papers [1-2]. This is a long-term issue which must receive a careful treatment from the perspectives of electromagnetics, mathematics, and computer hardware and software.

A second disadvantage is the large number of issues (some unfortunately political) which must be taken into account in the implementation of even this simpler form of the EMSES concept, a form which considers the framework as the interface of several separate, stand-alone CEM codes. Those involving the three components of input processing, the computational engine, and output processing are shown in table 1. Table 2 shows some of the issues which must be considered from the system level.

Table 1 refers to the CEM formulation level, that is, those issues that are related to an individual CEM formulation, tool (e.g., NEC, GEMACS, or the BSC), and application area. Table 2 addresses that level which can be considered to be analogous to the operating system of a computer. The environment must operate on a number of different computer platforms, including workstations, mainframes, distributed processors, and massively parallel processors. It must accommodate a variety of software packages, some of which are user developed and some of which are COTS. Each of the elements deserve a careful design process in itself, followed by a very thorough integration test to ensure compatibility among all the elements and with the various CEM tools now available.

INPUT PROCESSING	COMPUTATIONAL ENGINE	OUTPUT PROCESSING
Database Interface Generate Model User Interface Visualization Physical Measurements Electrical Parameters Modeling Elements Data Formats	Method of Moments Uniform Theory of Diffraction Finite Differences Finite Elements Physical Optics Physical Theory of Diffraction Characteristic Modes Time & Frequency Domains Hybridization	Database Interface User Interface Data Reduction Data Formats Types of Observables Presentation Formats Hardware Interfaces

Table 1. Processing Considerations for EMSES Design

Computer Platforms	Security	Maintenance and Testing
Programming Languages	Proprietary Data	Distribution
Operating Systems	Documentation	Configuration Control
Graphics Standards	Training	Transportability
Database Standards	Assets in Place	Administration and Control

Table 2. System Considerations for EMSES Design

This daunting task need not be started from scratch. There are a number of environments that are now in the process of being developed. One of them, the Microwave and Millimeter-Wave Advanced Computational Environment (MMACE), has a definite potential for providing the underlying foundation for the EMSES framework.

MICROWAVE AND MILLIMETER-WAVE ADVANCED COMPUTATIONAL ENVIRONMENT (MMACE)

Scientific and engineering research and design codes are typically FORTRAN-based having vastly different data structures and I/O, and unique and inconsistent user interfaces resulting in long learning curves and difficulties in usage. This is true of the codes implementing the various computational electromagnetics (CEM) formulations. The result is numerous opportunities for the introduction of human error in the design process because of usage difficulties, the required manual manipulation of data from code to code, the inability to graphically view the geometry, and the inability to ensure that all design codes are working with valid input data. This situation also adversely affects the cost (human and software/hardware) of maintaining and extending design capabilities as the computing environment and design needs advance. The Microwave & Millimeter-wave Advanced Computational Environment (MMACE) program is directed at the development of a coherent heterogeneous design environment which will address the above problems. Although MMACE is focused on vacuum electronics design, the underlying Research and Engineering

Framework (REF) is open in the sense of accommodating a broad range of scientific and engineering disciplines, including computational electromagnetics. The first phase of the MMACE program concentrated on development of a detailed understanding of vacuum electronics design requirements and on construction of a prototype design environment. The purpose of building a prototype was to begin experimenting with some of the concepts proposed for improving the design process with the intent being to provide an open and extensible testbed for ideas. However, it was important that this not be an intellectual exercise, but that the system be usable, provide immediate benefit, and provide a foundation for future development and testing.

The MMACE prototype is meant to be a first step in the long process of creating an extensive standard framework supporting design, simulation, and analysis for a broad range of research and engineering fields. It was released to the US microwave tube industry in January 1994 for evaluation and testing. The lessons learned from development and usage of the prototype are being applied to refine the direction of the Phase II framework development.

Figure 3 depicts the structure of the MMACE system. At the center is the Research and Engineering Framework. The REF acts as the switchboard and buffer between the user and the other environments. The REF core provides technology to support:

- ease of use;
- sharing of common design data;
- "master" geometry shared by all design tools and able to directly drive analysis codes;
- an advanced, standardized numerical grid generation capability;
- smooth, two-way integration of codes at different dimensionalities;
- flexible, interactive, and visual connection of tools and data flows;
- configuration and management of small group design data; and,
- transparent networking and distributed processing.

The actual analysis codes form the Computational Modules Environment. Although the MMACE program is testing the REF using analysis tools related to the design of power tubes, any group of codes with similar requirements could be easily "plugged" into the REF to take advantage of the core technologies.

Additional information regarding the MMACE program and the tools associated with it can be found on the Electromagnetics Bulletin Board World Wide Web Server, which can be accessed from Mosaic using URL *http://mmace.nrl.navy.mil:8080*.

EMSES AND MMACE

The MMACE schematic shown in figure 3 is modified and reproduced in figure 4 to show how its core Research and Engineering Framework (REF) can be used as the foundation for a CEM computational environment. Note that all the elements remain unchanged, with the exception that the vacuum electronics design toolset has been replaced by an electromagnetics toolset. The toolset consists of a number of CEM codes which implement the formulations shown in the figure. For example, method of moments is the engine of the NEC-MOM developed by the Navy, GEMACS developed by the Air Force, CARLOS developed by McDonnell-Douglas for the Electromagnetic

Code Consortium (EMCC), or any specific MOM code with which an analyst is familiar. The same applies to any of the other formulations shown in figure 4. Furthermore, still other existing and emerging formulations and their implementing codes can be easily added to this toolset.

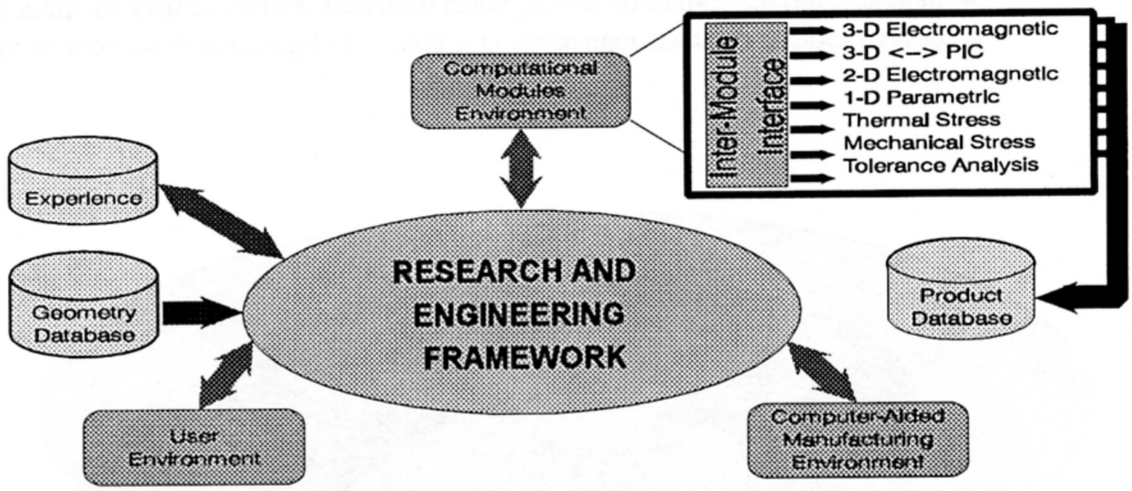

Figure 3. MMACE Conceptual Diagram

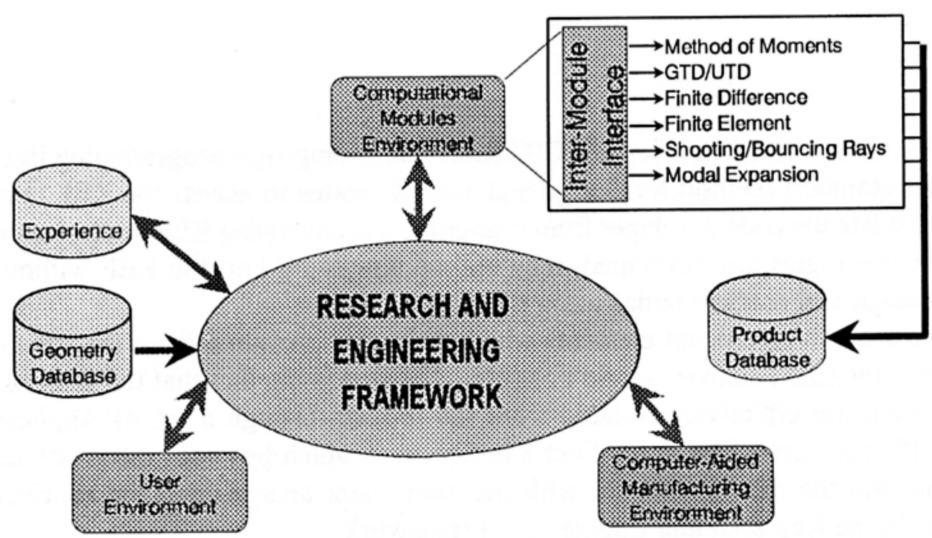

Figure 4. CEM Toolset in MMACE Framework

Figure 5 shows the internals of the REF core. It consists of:

- tools and utilities for storing, managing and visualizing design data including geometry;
- modules to support network communications and links to other environments (e.g., manufacturing); and,
- module interface standards which, when followed, allow the user to "plug in" the desired toolset, database management system, configuration management system, etc.

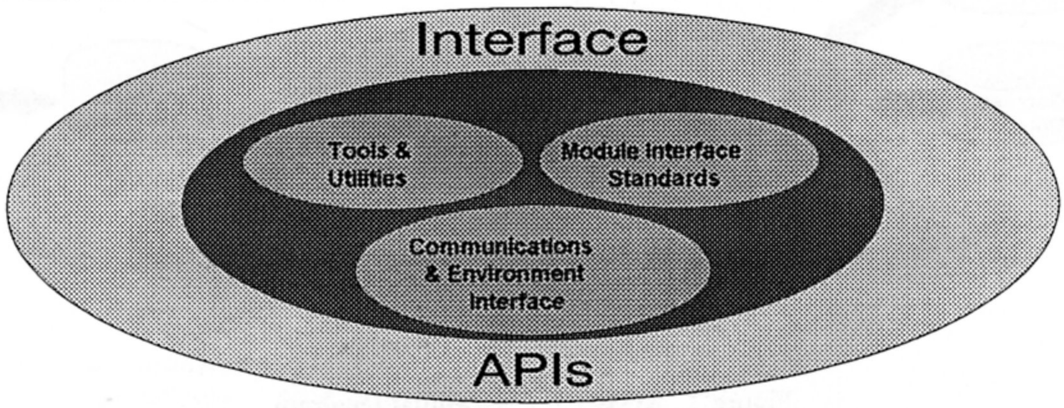

Figure 5. Research & Engineering Framework Elements

Surrounding these is a layer of FORTRAN and C language programming interfaces (APIs) which allow a standard method for design and analysis codes to access the REF core technology. These APIs insulate the code developer from changes to the underlying REF implementation. As new and better technologies are developed, they will be integrated into the REF without any need to change the design and analysis codes using the APIs.

There are two significant elements which make this transition from the vacuum electronics design world to the CEM analysis world possible. The first is the fact that the foundational tools of the environment are effectively isolated from the toolset through a set of Application Program Interface (API) routines. This is in effect a buffer zone which provides the interface between the analysis tools and the data associated with the item under analysis and the routines and utilities available within the Research and Engineering Framework.

The second is the fact that all interfaces are governed by a set of specifications which define formats for data, sequences of data, and operational interfaces with the framework. Tool and utility developers can thus be able to design interfaces which will match the requirements imposed by the

REF. The REF will have software routines which will provide templates and dialog boxes for the development of a user-specific API in accordance with the standards.

A further advantage is the fact that legacy codes need not be modified prior to interface with the REF. A code wrapper, supplied as part of the REF core technology, will perform all the input and output translations required to interface the analysis code with the core of the environment. Through a dialog process with the CEM code developer the necessary API will be generated in a nearly automatic fashion. This will ease the transition from the code operating in its current stand-alone mode to one that is interfaced with the REF.

Figure 6 shows conceptually the interface of an existing stand-alone code with the REF. Although the GEMACS code is identified in the figure, the process is not unique to this code. Any existing code can be treated in the same way.

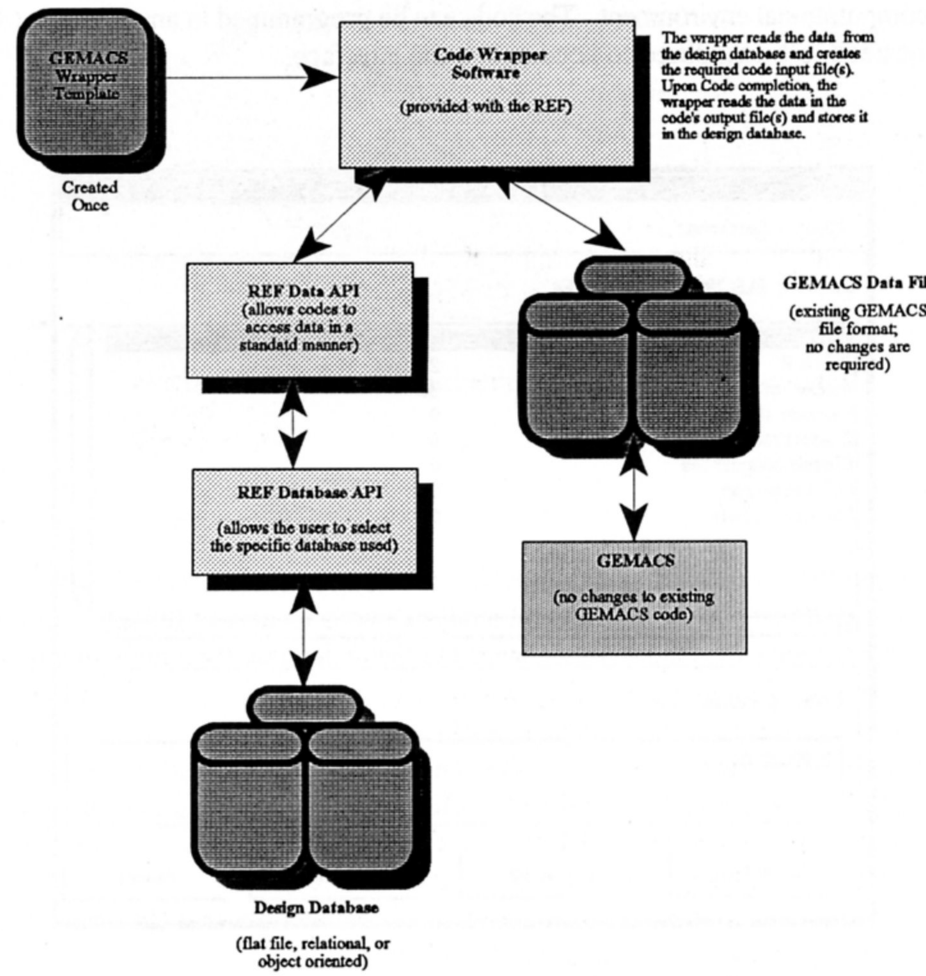

Figure 6. Code Interface with the REF

One of the initial steps performed in the process of interfacing a code with the REF is the development of the code wrapper, which was mentioned previously. This wrapper describes the data requirements of the code, including property names, data types, data formats, data values, default data values, valid data ranges, etc. This facilitates the extraction of data from an internal REF database. Figure 7 is an example of a code wrapper computer screen.

The output files from the code are also defined in order that pertinent analysis data may be located in the files, extracted properly, and stored in the internal database, using property names that are recognized by the analysis code.

For existing CEM codes the input and output data files are simply temporary files which are deleted when the job has been completed and the output data for the object under analysis have been transferred to the REF database.

The following observations demonstrate the power and versatility offered by the REF. First, as has already been stated, the REF provides a mechanism for the inclusion of existing codes within an advanced computational environment. The code can be programmed in any higher order language, and it does not need to be modified to accomplish the interface.

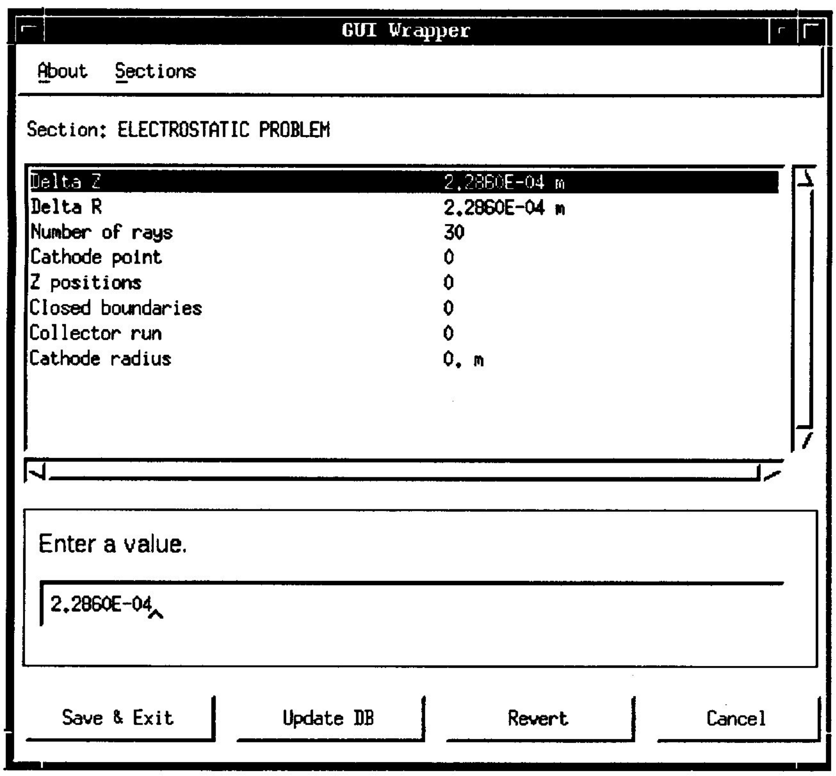

Figure 7. Code Wrapper Computer Screen

Second, the interface between the REF and the analysis code is a code wrapper which describes all properties of the data used by the analysis code. During the analysis process the data are stored in an internal database which is common to all the codes that are interfaced with the REF. Aliasing is allowed to facilitate use of data that is common to more than one code, but not necessarily identified with the same name in each of the codes which access that variable. Differences in units of measurement or other item descriptors can be translated within the code wrapper.

Phase II of the MMACE program will develop a computer-aided wrapper generator. Depending on the amount of data required or generated by the analysis code, it can be a laborious manual process to develop the wrapper for a specific analysis code. Software tools will ease the input of data and the grouping of like fields of data for eventual use by the analysis code.

Third, the CEM community can gradually move from the current "stove pipe" way of handling codes (see figure 1) to the framework method depicted in figure 4. Existing codes which have a long life potential can be first interfaced to the framework using the wrapper, and their own input and output methodology, as shown in figure 6. As a second step, the code's input and output handling routines can be eventually stripped away and use made of the REF-resident tools to accomplish these tasks.

This two-step process is illustrated in figure 8, again using the GEMACS code as an example. In the first phase GEMACS remains as an entity. As far as the internal structure and computational processes are concerned, the use of the REF code wrapper is invisible to the flow of the program. The input and output files are the same as when the code is used in a stand-alone mode. Also, all internal links within the GEMACS modules are utilized as if the code were being executed in a stand-alone mode.

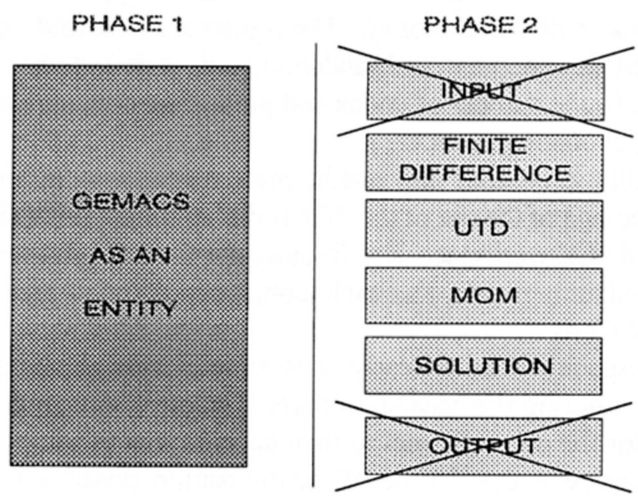

Figure 8. GEMACS Integration with the REF

However, in the second phase the code is broken into its various modules. For GEMACS this is a relatively simple matter since the code is really six independent programs tied together either manually or through a software batch process which executes the appropriate modules during the analysis run.

In phase 2 GEMACS would exist as six individual codes, contributing the initial population of the toolset shown in figure 4. Once this is accomplished, the GEMACS input and output modules would be discarded. Direct hooks to the REF database through a REF API would be developed to generate the input data file for a CEM analysis. A second set of hooks to the REF output visualization toolset would replace the functions provided by the current GEMACS output module. The temporary GEMACS input and output files, shown in figure 6, would not be generated in this case, as they would be in phase 1.

The fourth and final point follows from the third. New CEM codes being generated would be designed for insertion into the REF. The data input and output requirements would be developed in a manner consistent with the capabilities and rules of the REF. Any requirements for geometry generation, visualization, and the like would be met by hooks and calls to the tools and utilities within the REF. Thus, the code developer would focus on the electromagnetic issues, giving considerably less attention to the supporting issues, important though they may be.

Advantages of Adopting the REF As a CEM Environment

One advantage of adopting the REF as a CEM environment has already been cited: the code developer is able to focus more on the electromagnetics and less on developing a user-friendly graphical user interface (GUI). In fact, in the past most CEM codes developed by CEM researchers and engineers have presented formidable walls to potential users approaching them for the first time—or for the nth time, for that matter.

It is only recently that user-friendly GUIs have started to appear. These have been developed at great expense in terms of time and money. The results are two-fold: the resources were drawn away from CEM research, development, and validation; and, each code has its own GUI, duplicating the resource expenditure by the number of codes and presenting the user community with a different "appearance and behavior" for each code.

The walls are still there—they may just be prettier and more expensive.

The second advantage of the use of the REF is that it will present a single, consistent look and feel to the user. The MMACE workspace, also founded on the REF, is shown in figure 9. From this initial screen are accessed all the tools, and each icon, when pressed, presents a seven-choice menu which is identical for all icons.

The CEM workspace will be similarly constructed. This will be done using software tools which will be developed during the MMACE phase II effort. Thus, individual users will have the capability to further tailor the control panel to their specific modes of operation and personal taste.

The third advantage is the degree of customization possible in implementing the CEM workspace on an individual workstation. The REF and the various tools and utilities associated with it are effectively isolated from the user's codes, including the graphics engine. This is the result of the use of wrappers, both code and file, and the use of standards whenever possible, such as the IGES standard implemented in many commonly used graphics packages.

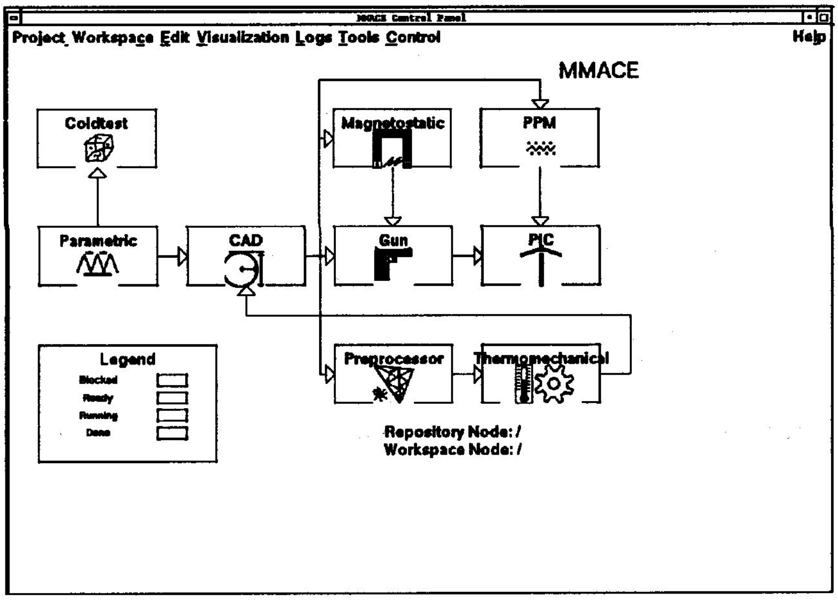

Figure 9. MMACE Control Panel

The user is then able to work with AutoCAD, BRLCAD, ACAD, CADRA, or any other CAD package, once the appropriate API has been developed. The translation will be further simplified through the use of the DT_NURBS package being developed by Boeing Computer Services under contract to the Navy David Taylor Model Basin (NSWC).

DT_NURBS is a mathematical library for spline geometry. This ARPA-, Navy-, Army-, and MMACE-sponsored program is a library of over 350 routines coded in FORTRAN-77. It provides engineering analysis applications with the capability to read and write CAD geometry files. It enables various engineering disciplines to establish a common geometry engine across many otherwise loosely related analysis programs.

Figure 10 shows the flow of data that occurs when one employs the DT_NURBS package to access the geometrical description of a structure from a CAD file. One begins with a CAD package of choice, which generally provides an IGES file to some extent in terms of the modeling elements available. An NSWC-provided batch program filters the set of NURBS entities to the subset supported by the NASA IGES specification. This subset is then operated on using the subroutines provided in the DT_NURBS package to transform the NURBS entities into a set of geometry elements which can be treated by the analysis code. Routines are being developed which can also generate a mesh substructure for use in finite element and finite difference codes.

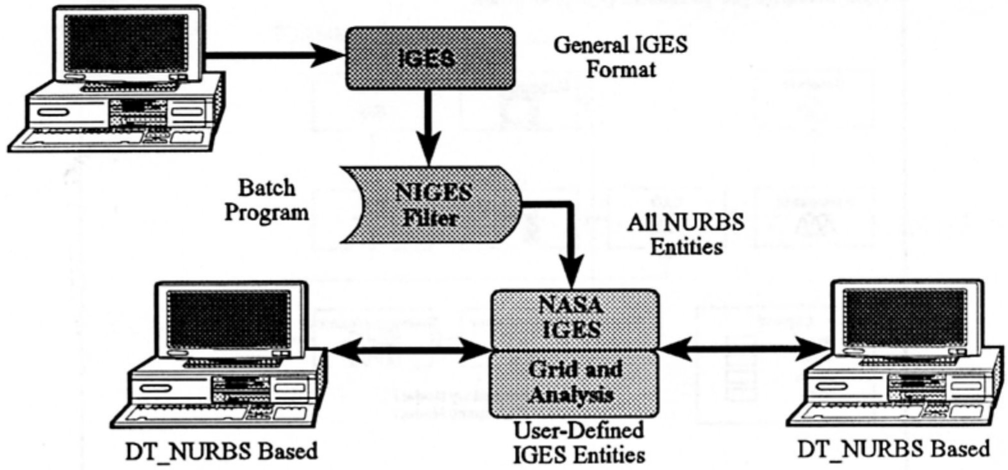

Figure 10. DT_NURBS Data Flow

A copy of the DT_NURBS package and its documentation can be obtained by contacting Dr. Robert Ames at the David Taylor Naval Basin via e-mail at ames@oasys.dt.navy.mil.

In addition, even though the distribution version of the CEM workspace contains codes derived from GEMACS, the individual user will be able to replace the GEMACS MOM code with NEC-MOM, CARLOS-3D, or any other MOM code of preference. It will require modification of the wrappers, but the existing wrapper can be used as a template.

Being able to use one's own suite of codes will reduce learning time and maintenance time. There will be an initial start-up cost in learning to use the REF and to incorporate one's tools of preference. However, it will be quickly amortized since a site will no longer need to develop and maintain its own input and output visualization tools, as well as a library of utilities and file transfer routines.

Reduced long-term maintenance cost is another advantage achieved by the presence of both a large user community and the contractors involved with the development of the REF. The large user community will more quickly uncover any bugs, anomalies, and inconsistencies present in the released version of the framework. Each problem will be isolated and "corrected" by the phase II contractor team for the NRL MMACE development program. The code will thereby become a stable, reliable framework more quickly.

One final advantage of using the REF is the capability it will have for distributed computing. The user will access the framework at a local workstation. Geometry data can be downloaded from a central location (e.g., government program office, integration contractor, airframe manufacturer) where the configuration control data is stored, maintained, and updated.

Individual CEM codes can be remotely accessed and executed, possibly on a remote massively parallel processor. The electromagnetic model would be developed on the local workstation once the geometry file had been downloaded. The model would then be uploaded to the computing site. The resultant analysis data would then be downloaded to the local workstation for further processing or the development of appropriate forms of display (e.g., X-Y plots of coupling vs. separation distance or polar plots of far-field patterns).

Selected data can then be uploaded to the central system facility for storage with the geometry data, from where it can be sent to other engineering groups for their use as they pursue their individual responsibilities for the development of the final product.

Thus the project will benefit in terms of money and schedule. The latest data will be available to all members of all the engineering teams associated with the project, including those concerned with logistics, test, and overall program direction. Trade-off studies among several competing concepts can be quickly and more thoroughly performed. Surprises and rework at system integration time can be minimized. Perhaps a product with unexpected capabilities will result.

SUMMARY

It has become apparent that a more efficient methodology must be developed and implemented for the performance of CEM analysis. This can be extended to engineering analyses of all disciplines. The optimum solution is one which provides a common computational framework across as many engineering disciplines as possible. The framework can also provide a common starting point over which strict configuration control is maintained. The framework will be invisible to the using communities, allowing the developers and analysts the freedom to focus on the application at hand rather than on learning the language and intricacies of the framework.

This paper has proposed the use of the MMACE-supported Research and Engineering Framework (REF) as the backbone of such an across-the-board computational environment. Although the demonstration has been in terms of computational electromagnetics (CEM), it can by simple extension and replacement of appropriate terms be applied to other engineering disciplines, such as computational fluid dynamics, thermal analysis, and circuit design.

The advantages and arguments cited for the use of the REF over other existing frameworks can be assumed by the proponents of those various frameworks. However, the one argument that none of the other frameworks may be able to claim to the degree to which it can be applied to the REF is the extreme effort that is being made to isolate the framework functions from the various applications. The isolation is being achieved through the use of code wrappers and file wrappers. This allows users to insert tools with which they are familiar or which are more appropriate for the particular requirements of the analysis at hand. This may add a start-up cost and it may add a computational overhead, but in the long run it results in greater efficiency during product development, increased efficiency of the analysts, increased cross fertilization among the engineering disciplines, and lower maintenance costs for the computational framework.

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The Practical CEMist

- practical topics in communications -

Perry Wheless, K4CWW

It will be March again and time for another ACES Symposium before you know it! We are hopeful of having enough papers for a session at the 1996 conference devoted to amateur radio antenna design and analysis. You are probably receiving this issue of the *ACES Newsletter* sometime after October 20, the deadline for paper summaries to be received by Richard Gordon, the 1996 Technical Program Chair. If the kind hand of fate has delivered your *Newsletter* early and you have interest in contributing a paper, please get in touch with me or Richard Gordon ASAP. In any event, check the conference program when it comes out to see if the session 'makes' and, if it does, by all means plan to attend and support this session at ACES '96 in Monterey.

Please note that the conference will be held during the traditional **third week in March**. Some early publicity shows the fourth week, but that has been changed! The new dates appear elsewhere in this *Newsletter*. Just be sure you have them straight before you buy your plane tickets, and we will eagerly await seeing you at the Naval Postgraduate School for ACES '96.

Jim Breakall (WA3FET) and Rudy Anders (AA2HT) have urged some social event for amateurs at the conference, but it fell through the cracks in 1995 and we don't want that to happen again in 1996. It is tricky to guarantee that any given evening during the conference week will be free, so I would like to know how many of you plan on arrival in Monterey by Sunday preceding the Monday conference kickoff. Please let me know if that Sunday would be either a good or bad time for you to participate in a meal and social period.

You may have noticed that the ACES membership renewal form now asks you to give your amateur radio call sign, if applicable. This year's renewal cycle is only partially complete, but at least thirty-six ACES members reported having amateur radio calls so far. The list includes: K3CXZ, AA2HT, G0GSF, VK2KRB, DJ2ZS, VK7AW, W0UN, K6ZTA, KB8I, WA5GKM, K1POO, VK1BRH, ZL2BDV, N3MPV, AH6NX, W7LMX, W7EL, ND9I, N6BRF, KC6Q, WA8DDS, SM7OUB, K7SWP, W0LJP, N8VZQ, K8GIH, W6JTH, W5LLE, WA4KBY, WB9AIA, K6FRY, VK4KAB, N4UFP, WI6X, and NX7U.

I sent a letter to known 'hamz in ACES' earlier (in September). If you did not receive a mailing from me, it indicates that either you are presently unknown to me as active in amateur radio, or there is some problem with the address I have for you. In either event, I hope you will contact me at your earliest convenience so that you can take your rightful place on our list.

The article for this installment of THE PRACTICAL CEMIST draws your attention to the feature in some contemporary wire antenna analysis programs of *real* conductor types. Q factor, which can take into account ohmic as well as radiation losses, provides an important description of wire antennas with imperfect conductors. High Q microwave resonators were analyzed manually, with graphical aids, in past years. Now, the computer-based determination of Q factor is more convenient, accurate, and reliable. However, the low Q factor resonances associated with HF/VHF wire antennas call for extension of the present techniques (based on assumptions strictly valid for $Q \geq 100$). The reader is invited to develop and submit refinements appropriate to low Q resonances for future publication in this column.

The submission of papers for this regular column in the *ACES Newsletter* is both invited and encouraged. Please send your contributed article to: Dr. W. Perry Wheless, Jr., P.O. Box 11134, Tuscaloosa, AL 35486-3008. You may reach me for discussions at telephone number 205-348-1757 or by Internet e-mail at wwheless@ua1vm.ua.edu.

Wire Antennas with Real Conductors

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Abstract - Some computer programs for the analysis and design of wire antennas now take into account the effect of real conductors. Ohmic loss in an antenna structure increases its apparent bandwidth. Q factor is most appropriate for the characterization of bandwidth, and numerical determination of Q factor is discussed in this paper. HF/VHF wire antennas exhibit low Q resonances that may require extension of microwave techniques, which are based on assumptions strictly valid only for high Q resonances.

I. Introduction

EARLIER computer programs for the analysis of wire antennas generally assumed idealized, PEC (perfect electric conductor) wires. This caused occasional misinterpretations among users. For example, these codes may deceptively suggest that certain broadband structures, such as the open-sleeve antenna with overlapping resonances, achieve impressive bandwidths with no sacrifice of radiated field strength.

Recently, analysis codes are starting to provide for wires of finite conductivity. The inclusion of real conductor types in the programs is a welcome addition, one which will help antenna designers keep their feet on solid ground. First principles of physics impose an unavoidable relationship between bandwidth and loss. It is educational and entertaining to experiment with the radiation resistance behavior resulting from various wire configurations (i.e., current distributions), but one must always be mindful of the presence and influence of ohmic losses.

Bandwidth with respect to wire antenna input impedance is evidenced by standing-wave ratio (SWR) change over a range of frequencies so, for example, one may consider 2:1 SWR bandwidth. Practical radio communicators frequently speak and think in terms of SWR bandwidth. The more common engineering convention, however, is the use of Q factor. There is a direct correspondence between the two, of course, as they both describe the same physical phenomenon. The discussion here is in terms of Q factor.

II. Objective

UNLOADED Q, denoted Q_0, is an inherent property of a resonant device, whether microwave cavity resonator or HF wire antenna, which describes bandwidth of the resonator in isolation. When the resonant device is coupled to an external circuit, the Q factor becomes loaded Q, denoted Q_L, and the resonant frequency changes to the loaded resonant frequency, f_L. The isolated resonant device, when coupled to the outside, becomes a resonant system, which includes the coupling means and an external circuit.

The main objective of this paper is to demonstrate one or two procedures for characterization of the resonant frequency and unloaded Q factor from wire antenna feed-point impedances calculated at a few discrete frequencies about resonance. Application of a wire antenna analysis program twice - once with PEC wires and again with a real conductor type - allows a meaningful appreciation of the effect of imperfect conductors by comparison.

In the case of an open, radiating antenna with negligible losses, one may choose the term 'radiation Q' for discussion in lieu of 'unloaded Q'. In the analytical procedures below, the notation Q_0 will be used throughout without distinction. The formulations are such that power dissipated in the antenna structure itself (i.e., ohmic loss) is included in the reported values of unloaded Q, Q_0.

III. Model Fitting to Impedances

An iterative procedure can be used for least squares estimation, from NEC (or similar) calculated feed-point impedances, of the parameters specifying a lumped-element circuit equivalent to a single wire antenna resonance. The equivalent circuit involves $M = 5$ parameters, and N impedances are calculated at discrete frequencies about resonance.

The equivalent circuit model follows that of [1], and is shown in Figure 1. R_e is negligible, and the parameterized input impedance is

$$Z = jX_e + j2R_cQ_e\delta + \frac{\kappa R_c}{1 + jQ_0\chi} \qquad (1)$$

where κ is the coupling coefficient, Q_0 is the unloaded Q factor, $\delta = \dfrac{\omega - \omega_0}{\omega_0}$, and $\chi = \dfrac{\omega}{\omega_0} - \dfrac{\omega_0}{\omega}$. The circuit model is thus described in terms of five parameters: $X_e, Q_e, \omega_0, \kappa$, and Q_0.

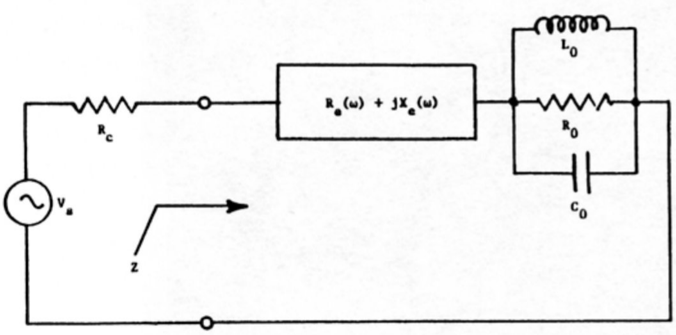

Fig. 1. Equivalent circuit valid near resonance ω_0.

The procedure for the determination of the numerical values of these circuit parameters is described in detail in [2], and will be only summarized here. The procedure is an iterative method for least-squares estimation, from an overdetermined system of impedances at discrete frequencies, of the parameters above. The five parameters are made elements of a 5×1 column vector $|a\rangle$. The referenced paper discusses how a 5×1 vector $|\beta\rangle$ and 5×5 matrix $\underset{\sim}{\alpha}$ are formed so that the iterative corrections to $|a\rangle$ are found by the calculation $|\Delta a\rangle = \underset{\sim}{\alpha}^{-1} |\beta\rangle$. Iteration continues until convergence to a user-specified ϵ is achieved. Error estimates on the final parameter values are provided by this method.

To illustrate this procedure, consider a 'voltage-fed' full-wave dipole of overall length 67 feet (20.43 m) in free space and in the vicinity of 14 MHz. Two sets of impedances were calculated using NEC-WIN Basic [3], first specifying 'perfect' 16 AWG wire and then specifying 'steel' wire of the same size. In both cases, impedances were calculated at 0.2 MHz intervals from 10 to 18 MHz, so the overdetermined system for least-squares solution in this case involves some 41 impedance data points.

Once the five parameters discussed above are determined by the numerical procedure, the circuit model clearly provides a continuous prediction of impedance versus frequency through Eq. 1. The result of directly fitting the model to the impedances calculated with NEC-WIN Basic for PEC conductors is shown in Figure 2. The graphical outcome of fitting the model to the impedances from the corresponding NEC-WIN run with steel wire selected is omitted, because its appearance is very similar to that for PEC wire. The numerical results are in Table I.

Table I. Least-squares model fit to NEC impedances.

	PEC	Steel
X_e	6.3 ± 85.6	-5.0 ± 79.8
Q_0	11.8 ± 0.23	11.1 ± 0.19
f_0 (MHz)	14.078 ± 0.006	14.054 ± 0.005
κ	1.13 ± 0.054	1.07 ± 0.049
Q_e	0.072 ± 0.058	0.079 ± 0.055

The unloaded Q drops from 11.8 to 11.1 when the dipole wire is changed from 'perfect' type to 'steel' type. The bandwidth of the steel dipole is approximately 6% greater than its PEC counterpart but, unfortunately, this arises from power dissipated in the antenna structure by ohmic loss.

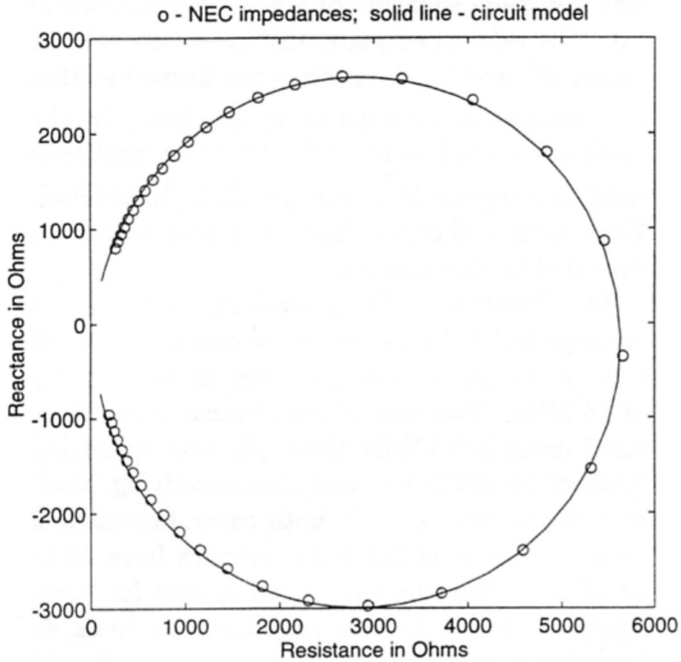

Fig. 2. Model fit to NEC impedances.

IV. Comparative Results

A more sophisticated approach to the computer-based characterization of resonant frequency, unloaded Q factor, and the coupling coefficient is contained in the monograph **Q Factor** by Darko Kajfez [4], which is accompanied by a useful diskette of software including the program *Qzero*. An important caveat to the reader is that Kajfez is concerned with Q factors of **at least 100**, and his preface explicitly states that 'The higher the Q, the more accurate are the assumptions used for the development of the data fitting procedure.' While the bounds on the reliability of his procedure for Q factors on the order of 10 are unknown at this time, experimentation with *Qzero* in the context of wire antenna resonances has produced some interesting results which are directly relevant here.

Application of program *Qzero* to the same dipole described earlier, of 'perfect' wire construction, produces the graphical summary shown in Figure 3. The numerical results for both PEC and steel wire are in Table II.

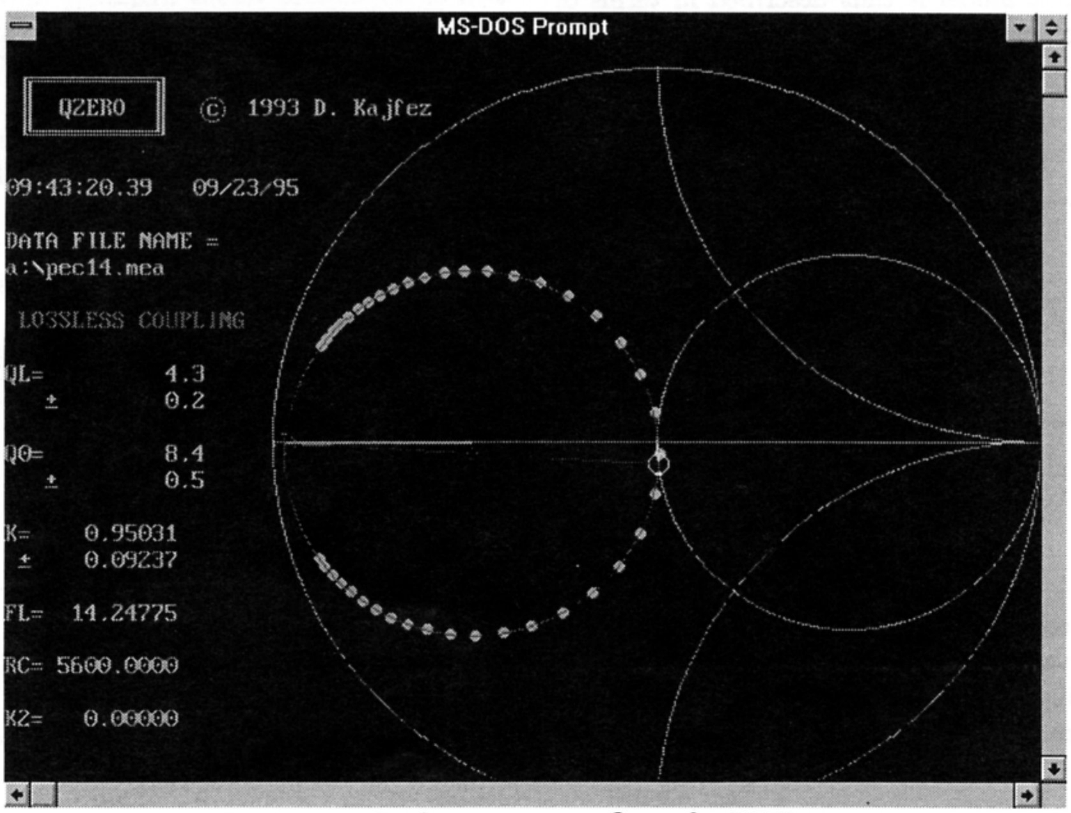

Fig. 3. Results from program *Qzero* for PEC dipole.

Table II. Results from program QZERO [4].

	PEC	Steel
Q_0	8.4 ± 0.5	7.8 ± 0.5
f_0 (MHz)	14.247	14.167
κ	0.934 ± 0.09	0.922 ± 0.093

V. Discussion of Results

ALTHOUGH the data processing methods for the two procedures discussed above differ, they share a common assumption of a coupling mechanism which is independent of frequency. For the iterative least-squares procedure, the model fit to NEC impedances shown in Figure 2 is actually after a rotation of the model curve (the solid line plot in Figure 2) clockwise by 4.7°. Without this rotation, the model prediction is displaced outside the circle defined by the discrete NEC impedances at low frequencies, and inside the circle at high frequencies. This likely suggests that the coupling mechanism for this low Q resonance is really frequency-dependent. In his book, Kajfez [4] warns that the assumption of a frequency-independent coupling mechanism becomes suspect for Q factors below approximately 100.

According to one practical handbook [5], one may expect dipole Q values to vary from 'about 14 for a length/diameter of 25,000 to about 8 for a ratio of 1250.' For the example case here, the length/diameter ratio is on the order of 16,000. Another ARRL publication [6], however, suggests that dipole Q's as low as 5 may be expected on occasion. Thus, it is difficult to conclude which estimations of unloaded Q here - those from Table I or from Table II - are more reliable. **In fact, a conclusion simply is not appropriate in the context of these low Q resonances,** as both analysis procedures are admittedly applied beyond the range of validity for their underlying assumptions.

VI. Conclusion

THE *good news* is twofold. First, the provision of contemporary wire antenna analysis codes for real conductors is an important feature, and designers should routinely employ this capability. Second, it is feasible to implement a computer-based procedure for the precise determination of Q factor for wire antenna resonances. Comparison of the analysis results for PEC wires against the corresponding results for real wires can provide valuable information about power dissipation in the subject antenna structure.

The *bad news* is that the historical need for analysis of Q factors has been in the realm of microwave resonators, with Q factors of hundreds to thousands. Some extensions of the present art will be required to allow accurate and reliable characterization of wire antenna resonances with Q factors on the order 5 to 20.

Here, all losses associated with the antenna structure itself have been lumped into Q_0. Altogether, four Q factors are useful for the characterization of resonant systems - loaded Q factor (Q_L), unloaded Q factor (Q_0), radiation Q (Q_{rad}), and external Q (Q_{ext}). If the maximum stored energy is $W_{\max} = W_e + W_m$, and the resonator conductor losses (plus dielectric losses, if relevant) are in the power $P_0 = P_{cond} + P_{diel}$, then unloaded Q is defined by

$$Q_0 = \frac{\omega W_{\max}}{P_0} \tag{2}$$

with the understanding that the expression is always evaluated at the resonant frequency. The external Q factor (Q_{ext}) accounts for power dissipated in the external circuit, where we are working with a resonant device which is coupled to an external circuit, either inductively or capacitively, with coupling coefficient κ relating the power dissipated in the external circuit to that dissipated in the resonant device by

$$Q_{ext} = \frac{Q_0}{\kappa}. \tag{3}$$

Referring to Fig. 3, for perfect wire, $Q_0 = 8.4$, $\kappa = 0.95$, and $Q_L = 4.3$ are reported. Eq. 3 says

Q_{ext} = 8.84 in this case, and the fundamental relation

$$\frac{1}{Q_L} = \frac{1}{Q_0} + \frac{1}{Q_{ext}} \tag{4}$$

is satisfied, with Q_0 amounting to 'radiation Q' since the antenna power is all radiated.

The corresponding figure for the steel wire analysis is not shown, but the numerical results from *Qzero* are $Q_0 = 7.8$, $\kappa = 0.922$, and $Q_L = 4.0$. In that case, it is more appropriate to introduce an explicit radiation Q,

$$Q_{rad} = \frac{\omega W_{\max}}{P_{rad}} \tag{5}$$

so that Eq. 4 becomes

$$\frac{1}{Q_L} = \frac{1}{Q_0} + \frac{1}{Q_{rad}} + \frac{1}{Q_{ext}}. \tag{6}$$

The formal distinction between Q_0 and Q_{rad} was not made in this paper, but the point of these remarks is that such a distinction is practical and will be developed further in a future article.

Readers who are seriously interested in this subject are urged to acquire and study reference [4], where the theoretical foundations are carefully worked out.

VII. Acknowledgment

The author wishes to gratefully acknowledge the active assistance of Professor Darko Kajfez with the least-squares procedure, as well as constructive conversations regarding his recent work on Q factor determination.

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Tutorial Article

It is usually desirable in computational electromagnetics to obtain the greatest amount of information possible with the least number of computations. Typically, observables such as radiation patterns, input impedance, transient signatures etc. are desired over all space, wide bandwidths, or long time windows depending upon the needed quantity. However, obtaining all the desired information directly from computations based on first principle models can be time-consuming and costly, if possible at all. Model Based Parameter Estimation (MBPE) is an alternative, more computationally efficient means of obtaining information between and beyond those points that are computed from a first principles model. MBPE differs from straight-forward curve fitting in that the fitting model is based on the underlying physics and mathematics of the first principles model.

I am pleased that Dr. Edmund Miller has contributed the tutorial, "Model based parameter estimation in electromagnetics: I - Background and theoretical Development " to this issue of the *ACES Newsletter*. This article is Part I of a three part series on MBPE that Dr. Miller has contributed. Part II, "Model based parameter estimation in electromagnetics: II - Applications to EM observables" will appear in the March 1996 issue of the *ACES Newsletter*, and Part III, "Model based parameter estimation in electromagnetics: III - Applications to EM integral equations " will appear in the November 1995 issue of the ACES journal.

Dr. Miller needs little introduction in the electromagnetics community. Which is good, because he is presently lecturing in England as this manuscript goes to press, and I have forgotten to request a biographical sketch from him! He has made many fundamental contributions to computational electromagnetics, and many contributions to the organization and continued growth of ACES. In his "retirement" he continues to serve the electromagnetics community as a teacher and scholar.

If you have ideas or suggestions for future tutorial articles, would like to contribute a tutorial article to the newsletter, or have comments on past articles, please feel free to contact me:

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I would greatly welcome suggestions and contributions.

MODEL-BASED PARAMETER ESTIMATION IN ELECTROMAGNETICS: I--Background and Theoretical Development

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0. ABSTRACT

Electromagnetics, as is true of other scientific disciplines, utilizes solution tools that range from first-principles analytical formulations to approximate, engineering estimates. One goal in essentially all problem solving, but especially for design applications where specific performance is desired, is that of expending only enough solution effort to obtain a quantitative result commensurate with the problem requirements. Included among the possible approaches for achieving such a goal is model-based parameter estimation (MBPE).

In using MBPE, the requirement of obtaining all samples of desired observables (impedance, gain, RCS, etc.) from a first-principles model (FPM) or from measured data (MD) is circumvented by instead using a reduced-order, physically-based approximation of the sampled data called a fitting model (FM). One application of a FM is for interpolating between, or extrapolating from, samples of FPM or MD observables to reduce the number of samples that are needed. Another is to use a FM in FPM computations by replacing needed quantities by simpler analytical approximations to reduce the computational cost of the FPM itself. In either application, the FM can greatly reduce the effort needed to obtain the desired information. In addition, the FMs can be more suitable for design and optimization purposes than the usual numerical data that comes from a FPM or MD because MBPE descriptions can normally be handled directly analytically rather than via operations on the numerical samples.

This article provides a background and motivation for using MBPE in electromagnetics, focusing on the use of FMs that are described by exponential and pole series and how data samples obtained from various kinds of sampling procedures can be used to quantify such models, i.e., to determine numerical values for their coefficients. The application of MBPE to various kinds of electromagnetic observables will be demonstrated in a second article [Miller (1996)]. The concluding article in this series will discuss how MBPE might be used to improve the efficiency of FPMs based on frequency-domain integral equations [Miller (1995)].

1.0 BACKGROUND AND MOTIVATION

Most EM phenomena, whether observed in the time domain or the frequency domain, or as a function of angle or location, are not of interest at one or a few discrete times, frequencies, angles or locations but, instead, require essentially continuous representation over some specified observation interval. Even with increasingly sophisticated instrumentation or computer models, determining EM observables to sufficient resolution as a function of the relevant observation variable can be expensive as well as potentially error-prone. A typical problem now involves determining time- or frequency-domain responses over bandwidths that are no longer just a few percent or a few megahertz, but might extend over frequency ranges that are 10:1 or more in relative bandwidth or multiple gigahertz bandwidths in absolute terms. Resolving a response that contains many high-Q, closely spaced resonances requires slow sweeping rates or long observation times experimentally, or an excessive number of frequency samples or time steps computationally.

A computational basis for solving most problems in physics and engineering derives from first-principles mathematical descriptions, or first-principles models (FPMs), of the applicable physics. Depending on the problem parameters and information needed, various physically and mathematically based approximations are almost invariably used subsequently in obtaining numerical results. The approximations can arise from: making appropriate changes in the physical problem (e.g., replacing a solid, metallic surface with a wire mesh); exploiting commonly-encountered problem attributes (such as the thin-wire approximation); or taking advantage of asymptotic trends (as in geometrical optics and the geometrical theory of diffraction). Such approximations can yield acceptable results for a wide variety of problems, and might be regarded as being derived or "second-generation" FPMs, i.e., they originate from a FPM to which they have a well-defined relationship rather than directly from Maxwell's equations themselves.

Another category of models might also be recognized which are further removed from the FPM on which they are based, but are instead related to some variable associated with that model or to some observable that FPM produces. For example, consider that when wideband, impulsive excitation is used, the electromagnetic (EM) late-time response can be rigorously represented as an exponential series in time whose exponents are generally complex, problem-dependent, parameters. Such a representation of the late-time response can be directly deduced from the FPM [Cordaro and Davis (1980)] and forms the basis for the Singularity Expansion Method (SEM) [Baum (1976)]. Generally, however, the exponential series otherwise retains no information specific to the problem being modeled in terms of that problem's physical or electrical characteristics, except to the extent that the parameters of the series might imply them. The value of such a model is that it can substantially reduce the effective rank of the problem description insofar as the amount of information needed to represent the observables that exhibit such behavior are concerned. Thus, rather than needing the complete problem description required to compute the NxN interaction coefficients of an integral-equation (IE) FPM, this reduced-rank, parametric description can be substituted for the FPM in appropriate applications. Beyond that, knowledge that such a parametric model properly describes physical observables for general problems in appropriate circumstances suggests that the parametric model may be applied to other data derived from the same kind of physics, whatever process produced that data. Such reduced-order models, whose analytic form is known and which are therefore completely specified when their associated parameters are quantitatively determined, provide the basis for "model-based, parameter estimation" (MBPE).

Although MBPE is discussed here specifically with respect to some representative EM applications, it should be appreciated that it is a very general procedure that is applicable to essentially any process, physical or otherwise, for which a reduced-order, parametric model can be deduced. Also, it must be noted that MBPE is not "curve fitting" in the sense that term is normally used, which also can involve finding the parameters of some function which is fit to the available data. The essential difference between MBPE and curve fitting is that the former uses a fitting model (FM) based on the problem physics, while the latter need not do so, which is why MBPE might be characterized as "smart" curve fitting. When curve fitting includes the goal of finding the correct FM for the process that generated the given data, this approach can also be described as "system identification." It's worth emphasizing that MBPE is not limited to physical processes but forms the basis for variously named analytical procedures {e.g., Kummer's method, Richardson extrapolation and Romberg quadrature [Ralston 1965)]} whose purpose is to speed the numerical convergence of mathematical representations involving integrals and infinite sums and wherein an integrand function or a sequence of partial sums can be regarded as a generalized "signal." In discussing a non-linear procedure he developed for a similar purpose, Shanks (1955) referred to such phenomena as "physical" and "mathematical" transients. In essence, any process that uses or produces a sequence or set of samples is a candidate for MBPE using as a FM whatever mathematical representation is appropriate as a reduced-order representation of that process.

This article begins by discussing two FMs, given by exponential and pole series, respectively, that are widely encountered in electromagnetics, related to each other by the Laplace transform, and which arise whenever wave-equation problems are encountered. As a paradigm for the general problem we use the time-frequency transform pair for this purpose and first illustrate use of MBPE for time waveforms that are described by sums of complex exponentials. As an example of the other half of the transform pair, we then consider using MBPE for spectral estimation, not, however, as is most-often the case in signal processing where time samples provide the starting data, but instead when the original data comes from sampling in frequency. It is demonstrated that the parameters of the time-frequency models can be obtained numerically either from samples of the process being modeled, which we call "function sampling," or alternatively from samples of the time or frequency derivatives of the process, which we call "derivative sampling." Finally, we consider some other aspects of MBPE and preview two companion articles [Miller (1995) and (1996)], referred to herein as RII and RIII, respectively, in a concluding discussion. These companion articles include a variety of sample applications from the perspective of using MBPE to reduce the complexity of FPM computations and to minimize the number of FPM samples that are needed to represent the EM observables that these models produce.

2.0 WAVEFORM-DOMAIN AND SPECTRAL-DOMAIN MODELING

Although Maxwell's equations are usually first encountered in the time domain, traditionally most subsequent analysis has been done in the frequency domain, a situation, however, that is changing with the increasing computer power available. Each of these two domains describes the same physical problem in principle, and either is capable of providing the sought-for EM behavior. Therefore, it seems reasonable to seek solutions in that domain and using that formulation best suited to the problem at hand, wherein medium or object non-linearity, dispersivity, inhomogeneity, anisotropy, etc. might become the determining factor in choosing a model. Whatever approach is taken to solve a problem, it must be noted that once the problem's sampled observables have been obtained it is often necessary to examine their behavior in the transform domain, i.e., in a domain other than that in which the solution has been developed.

The two most frequently used domains for formulating and solving EM problems, and for many other physical phenomena as well, are the time domain (TD) and frequency domain (FD), for which generic descriptions are given by exponential and pole series respectively, which together form a Laplace-transform pair. More generally, it should be noted that the same transform relationship exists between other observable pairs that are also described by exponential and pole series as listed in Table I. Thus, we use the terms "waveform domain" (WD) and "spectral domain" (SD) respectively for phenomena that are described by exponential series and pole series as a generalization of their more specific and familiar TD and FD forms.

Series comprised of complex exponentials or complex poles can represent the kinds of wave-equation solutions that arise in EM and similar physical phenomena basically because they occur in solutions to the kinds of differential equations that describe the problem physics and thus might be regarded as natural "basis" functions for developing numerical solutions to such problems. One example of this is demonstrated by SEM [Schelkunoff and Friis (1952), Baum (1976)] which exploits the fact that the EM behavior of conducting objects is described in part by complex-frequency resonances, or poles, which are located away from the $i\omega$ axis in the complex-frequency plane. While the pole terms alone do not describe the entire response, a summation of poles can be a good approximation to that response and provides a concise way to represent it. For our purpose, which is to develop more efficient FPMs as well as to reduce the number of samples of those observables that are needed in the first place, the FM need not be exact so long as it provides an acceptably accurate and parsimonious representation. This is an important distinction of a FM as compared with what is required of a FPM.

The generic WD and SD FMs can be expressed as

$$f(x) = f_p(x) + f_{np}(x) = \Sigma R_\alpha \exp(s_\alpha x) + f_{np}(x),\ \alpha = 1, \ldots, P \quad (1)$$

and

$$F(X) = F_p(X) + F_{np}(X) = \Sigma R_\alpha/(X - s_\alpha) + F_{np}(X),\ \alpha = 1, \ldots, P \quad (2)$$

where "x" represents the WD independent variable and "X" is its SD, or transformed, counterpart. For the time-frequency transform pair, x would be the time variable t and X would be the complex frequency s, in general, but for most purposes would be limited to radian frequency $i\omega$. Note that while the exponential or pole series contributions, designated respectively by $f_p(x)$ and $F_p(X)$, represent what we might call the "resonant" response, there is in general a non-pole component as well, denoted by $f_{np}(x)$ and $F_{np}(X)$. The additional non-pole term in the waveform and spectral models is included to account for the fact that the complete response is not entirely described by the exponential or pole terms, except for the former at "late" times when an object is no longer excited by an incident field. During the time when the exciting field continues to affect the object's response, i.e., the "early" time, the response is not purely resonant. Thus, the transient response can be divided into the early-time, or driven, response and the late-time, or source-free response, distinguished in time by the transition from driven to source-free conditions. Although a similar observation can be made for the FD, there the separation between the driven and source-free behavior is not so clearly demarcated because essentially all frequency components are affected by both time behaviors.

The FM parameters, the complex resonances (or poles), "s_α," and the modal amplitudes (or residues), "R_α," (or their rational-function counterparts which occur as polynomial coefficients) are quantified by fitting samples of the relevant observable or phenomenon to the desired model, $f_p(x)$ or $F_p(X)$. Once these parameters are available from one domain, they can be used to obtain the observable they represent in the transform domain as well. The various FMs and their estimated parameters provide a mathematically-concise and physically insightful way to characterize electromagnetic and other wave-equation phenomena. The relationship between the model parameters and observables in the two domains are illustrated conceptually in Fig. 1.

Concluding then that exponential and pole series can be appropriate FMs for EM applications of MBPE, there remain the important questions of determining how FPMs should be sampled and

TABLE I
VARIOUS WAVEFORM- AND SPECTRAL-DOMAIN FIELD EXPRESSIONS

DOMAIN	VARIABLES		MODEL	PARAMETERS	COMMENTS
WAVEFORM (Exponential)	x	g	f(x)	Residues	
SPECTRAL (Pole)	X		F(X)	Poles	
			Model 1a--Time/Frequency		
Time	t	--	$f(t) = \Sigma M_\alpha \exp(s_\alpha t)$	M_α = amplitude of αth mode.	
Complex Frequency	s		$F(s) = \Sigma M_\alpha/(s - s_\alpha)$	$s_\alpha = i\omega_\alpha - \sigma_\alpha$ = complex resonance frequency of αth mode.	Provides a way to find SEM poles from transient or spectral data [Poggio et al. (1978)].
			Model 1b--Frequency/Time		
Frequency	ω	--	$f(s) = \Sigma M_\alpha \exp(t_\alpha s)$	M_α = amplitude of αth mode.	
Time	t		$F(t) = \Sigma M_\alpha/(t - t_\alpha)$	t_α = peak time for a waveform	Superposition of sinusoidal signals.
			Model 2a--Frequency/Space		
Frequency	ω	i/c	$f(\omega) = \Sigma S_\alpha \exp(g\omega R_\alpha)$	S_α = amplitude of αth source.	
Space	R		$F(R) = \Sigma S_\alpha/[g(R - R_\alpha)]$	R_α = position of αth source along line of view.	Related to profiling or inverse imaging along viewing direction [Miller and Lager (1982)].
			Model 2b--Space/Frequency		
Space	R	i/c	$f(R) = \Sigma S_\alpha \exp(gR\omega_\alpha)$	S_α = amplitude of αth source.	
Frequency			$F(\omega) = \Sigma S_\alpha/[g(\omega - \omega_\alpha)]$	ω_α = frequency of αth source.	Observing field of multi-frequency sources over a range of spatial locations.
			Model 3a--Space/Angle		
Space	x	ik	$f(x) = \Sigma P_\alpha \exp\{gx[\cos(\varphi_\alpha)]\}$	P_α = amplitude of αth incident plane wave.	
Angle	φ		$F(\phi) = \Sigma P_\alpha/\{g[\cos(\varphi)-\cos(\varphi_\alpha)]\}$	φ_α = incidence angle of αth plane wave with respect to line of observation.	A basis for direction finding [Benning (1969)].
			Model 3b--Angle/Space		
Angle	φ	ik	$f(\varphi) = \Sigma S_\alpha \exp[gx_\alpha \cos(\varphi)]$	S_α = amplitude of αth source.	
Space	x		$F(x) = \Sigma S_\alpha/[g(x-x_\alpha)]$	x_α = position of αth source along linear array.	Can be used for pattern synthesis and imaging linear source distributions [Miller and Lager (1983)].

how the FMs themselves can be most effectively employed. Each of these questions is considered in the following discussion. It must be noted that exponential- and pole-series models are not the only physics-based FMs that can be used in electromagnetics, but because of their ubiquitous presence in first-principles formulations and EM observables they are the ones we choose to emphasize here. Some other kinds of FMs are discussed in RII.

3.0 SAMPLING FIRST-PRINCIPLE MODELS AND OBSERVABLES IN THE WAVEFORM DOMAIN

It was mentioned above that function and derivative sampling of the FPM can be employed in both the WD and SD, a point to be expanded upon below. Furthermore, it will also be shown that a given FM can use essentially any combination of function and derivative samples so long as the sampling satisfies whatever constraints are imposed by the FM. For example, the simplest FM implementation in the WD requires that the sampling be done in uniform steps of the independent variable x. Aside from such constraints, the FM only needs enough information from the FPM or MD to permit the FM parameters to be computed to the needed accuracy. This means that there can be a great deal of latitude available in using MBPE, an advantage on the one hand but one presenting, at the same time, an array of options that somewhat obscures how to obtain the best results from its use on the other.

3.1 Waveform-domain function sampling

Our starting point for function sampling in the WD is provided by Prony's method, a procedure whose presence can be discerned in much of modern signal processing, even though developed originally 200 years ago [Prony (1795)]. Although other signal-processing approaches may yield better results, especially for noisy signals, e.g. pencil-of-function-based methods [Sarkar, et al. (1980), Sarkar and Pereira (1995)], all begin at the same starting point using a series of discrete data samples. Prony's method has the advantage of being so obviously and directly connected to the physical process being modeling and begins by assuming the availability of uniformly-spaced samples of an exponential series, i.e.,

$$f_i = f(x_i) = f(i\delta x) = \Sigma R_\alpha \exp(s_\alpha x_i) = \Sigma R_\alpha \exp(s_\alpha i\delta x),\ \alpha = 1 \text{ to } P,\ i = 0, \ldots, D - 1 \tag{3}$$

where δx is the sampling interval and there are a total of D samples. Rewriting Eq. (3), we get

$$f_i = \Sigma R_\alpha (X_\alpha)^i,\ i = 0, \ldots, D - 1 \tag{4a}$$

where

$$X_\alpha = \exp(s_\alpha \delta x). \tag{4b}$$

The equations represented by (3) and (4) above can be written explicitly as

$$f_0 = R_1 + R_2 + \cdots + R_P, \tag{5a}$$

$$f_1 = R_1X_1 + R_2X_2 + \cdots + R_PX_P, \tag{5b}$$

$$f_2 = R_1(X_1)^2 + R_2(X_2)^2 + \cdots + R_P(X_P)^2, \tag{5c}$$

$$\vdots$$

$$f_{D-1} = R_1(X_1)^{D-1} + R_2(X_2)^{D-1} + \cdots + R_P(X_P)^{D-1}, \tag{5d}$$

where a polynomial relationship is clearly shown by the pattern of the X_α terms in each succeeding equation. This form suggests that the X_α might satisfy a polynomial of the form

$$A(X) = a_0 + a_1X + a_2X^2 + \cdots + a_PX^P = (X - X_1)(X - X_2)\cdots(X - X_P) = \Pi(X - X_i) = 0, \tag{6}$$

which is known as the characteristic equation. Upon multiplying the equation for f_0 by a_0, that for f_1 by a_1, etc. through f_P and adding the resulting equations, we obtain

$$f_0a_0 + f_1a_1 + \cdots + f_Pa_P = 0 \tag{7a}$$

where $A(X_1) = A(X_2) = \cdots = A(X_P) = 0$ has been used. Repeating this sequence of operations by multiplying f_1 by a_0, f_2 by a_1, etc. leads to

$$f_1a_0 + f_2a_1 + \cdots + f_{P+1}a_P = 0. \tag{7b}$$

Continuing these steps to generate a total of D - P equations, we obtain

$$f_2a_0 + f_3a_1 + \cdots + f_{P+2}a_P = 0, \tag{7c}$$

$$\vdots$$

$$f_{D-P-1}a_0 + f_{D-P}a_1 + \cdots + f_{D-1}a_P = 0. \tag{7d}$$

Eq. (7) forms the basis for finding the coefficients of the characteristic equation. Note, however, that Eq. (7) is homogeneous and requires some additional information, or a constraint on the characteristic-equation coefficients, for the problem specification to be completed.

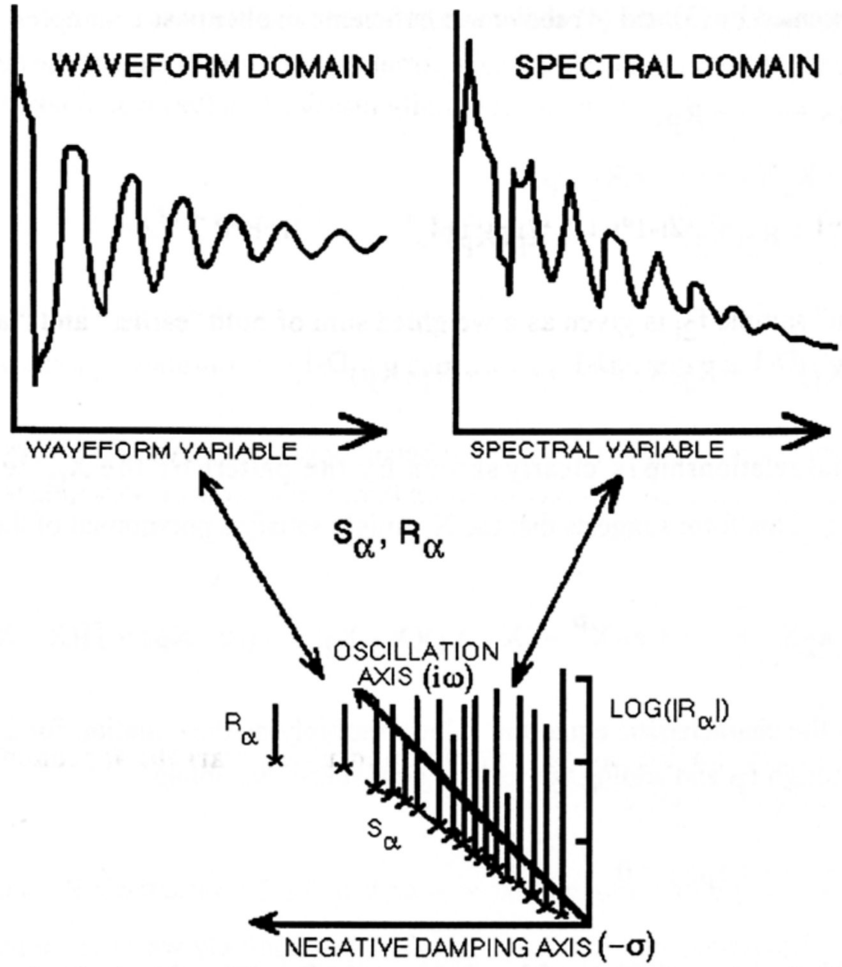


Figure 1. Illustrations of the relationship between the waveform-domain (left-hand plot) and spectral-domain (right-hand plot) observables and the two sets of parameters, R_α and s_α, common to them both (the bottom plot). The pole locations are shown in the quadrant II only, since they usually occur in complex conjugate pairs reflected about the damping axis and have only negative real parts.

Various approaches can be used for this purpose. Among the constraints that have been used are:

$\Sigma(a_i)^2 = 1, i = 0, 1, \ldots, P,$	(energy constraint)	(8a)
$a_P = 1,$	(linear-predictor constraint)	(8b)
$a_i = 1.$	(non-causal constraint)	(8c)

Prony's method employs $a_P = 1$, leading to the "linear-predictor" equation

$$f_i a_0 + f_{i+1} a_1 + \cdots + f_{i+P-1} a_{P-1} = -f_{i+P}, \tag{9a}$$

so-called because knowledge of the P predictor coefficients and the past P samples of the process that generated f(x) yields, or "predicts," the next sample as a weighted sum of the past P samples. If some intermediate coefficient were to be set to unity instead, then (9a) would take the form

$$f_i a_0 + f_{i+1} a_1 + \cdots + f_{2i-1} a_{i-1} + f_{2i+1} a_{i+1} + \cdots + f_{i+P} a_P = -f_{2i}, \tag{9b}$$

where the "present" sample f_{2i} is given as a weighted sum of both "earlier" and "later" samples. The energy constraint is not-often employed because it leads to a non-linear system of equations.

The parameters of the exponential-series FM now become accessible by first solving (7) with a_P = 1 for the remaining P predictor coefficients and then finding the roots of the characteristic equation for the X_α from which

$$s_\alpha = (1/\delta)\log(X_\alpha) \tag{10}$$

follows. The residues can be subsequently found using various approaches, the most straightforward of which is to solve Eq. (5) directly.

We note that the number of data samples must be such that $D \geq 2P$ since the P R_α's and P s_α's are needed to quantify the exponential model of Eq. (3). Alternatively we observe that there are P characteristic-equation coefficients to be found so that $D - P \geq P$ is required in Eq. (7). Since the value of P is itself often not known, some experimentation is usually needed to determine it. And while there is a definite lower limit to the number of data samples that are needed, it may happen that many more samples are available than this because a longer data record is available and/or the sampling interval is shorter than the minimum required. Such oversampling represents redundant information that might be used to reduce noise effects. Also, it should be appreciated that uniform sampling in x of f(x) is required for the estimation procedure outlined above to be realized. Non-uniform sampling in x would not produce the integer exponents in Eq. (5) that permits the polynomial of Eq. (6) to be utilized.

Many variations of the above procedure can be identified, differing most significantly in how the data samples might be "preprocessed" prior to accomplishing the actual parameter estimation. For example, rather than starting with the data samples themselves, auto-covariance estimates might be used instead, primarily as a means of reducing noise effects. Although the auto-covariance samples might be defined in various ways, for a finite-length data sequence they basically involve replacing the individual f_i's by sums of shifted, two-sample products of the data. An essentially equivalent outcome is achieved when the linear-predictor coefficients are solved for using a pseudo inverse, a process that also replaces data samples by product sums since the matrix then solved is a product of the original matrix by its transpose.

Other approaches for reducing noise effects are also available. If multiple measurements of the same process can be made and if the noise is uncorrelated, averaging will improve the signal-to-noise ratio of the data samples. Alternately, these separate data sets might be used to obtain multiple sets of parameter estimates which can then be averaged. The problem of dealing with noisy, uncertain and incomplete data is an important one that goes beyond the scope of this tutorial discussion. The applications for MBPE of primary concern in this presentation are those where noise is not usually a limiting factor. However, a FM can also be used to reduce the effects of noise or uncertainty in oversampled data as is illustrated in RII for WD FMs, and which is a goal of most signal processing.

3.2 Waveform-domain derivative sampling

Derivative sampling of WD data proceeds in a fashion analogous to that used for function sampling. We begin with the WD model of Eq. (1), i.e. the same FM is employed when using derivative sampling as when using function sampling, but rather than obtaining samples of the process as a function of x we instead use

$$f_n = d^n f(x)/dx^n = \Sigma R'_\alpha (s_\alpha)^n \exp(s_\alpha x),\ \alpha = 1, \ldots, P;\ n = 0, \ldots, D-1. \tag{11}$$

Making a change of variables like that done above for function sampling, we rewrite Eq. (11) as

$$f_n = \Sigma R_\alpha (s_\alpha)^n,\ \alpha = 1, \ldots, P;\ n = 0, \ldots, D-1 \tag{12}$$

where $R_\alpha = R'_\alpha \exp(s_\alpha x)$, i.e., the variation in f_n due to whatever value is used for the observation variable x is absorbed in the residue.

Upon explicitly writing some of the individual equations implicit in Eq. (12), we obtain

$$f_0 = R_1 + R_2 + \cdots + R_P,$$
$$f_1 = R_1(s_1)^1 + R_2(s_2)^1 + \cdots + R_P(s_P)^1$$
$$f_2 = R_1(s_1)^2 + R_2(s_2)^2 + \cdots + R_P(s_P)^2$$
$$\vdots$$

$$f_{D-1} = R_1(s_1)^{D-1} + R_2(s_2)^{D-1} + \cdots + R_P(s_P)^{D-1} \tag{13}$$

Again, a polynomial model is suggested by the terms appearing in Eq. (13). If we assume that

$$B(s) = b_0 + b_1 s + b_2 s^2 + \cdots b_P s^P = \Pi(s - s_\alpha),\ \alpha = 1, \ldots, P \tag{14}$$

and multiply the first equation in (13) by b_0, the second by b_1, etc., and add the resulting equations together we obtain

$$f_0 b_0 + f_1 b_1 + \cdots + f_P b_P = 0 \tag{15a}$$

since $B(s_1) = B(s_2) = \cdots = B(s_P) = 0$. Continuing this process by successively repeating these steps starting with the equation for f_1, f_2, f_3, et., we get

$$\begin{aligned}
&f_1 b_0 + f_2 b_1 + \cdots + f_{P+1} b_P = 0,\\
&f_2 b_0 + f_3 b_1 + \cdots + f_{P+2} b_P = 0,\\
&\cdot\\
&\cdot\\
&\cdot\\
&f_{D-P-2} b_0 + f_{D-P-1} b_1 + \cdots + f_{D-2} b_P = 0,\\
&f_{D-P-1} b_0 + f_{D-P} b_1 + \cdots + f_{D-1} b_P = 0.
\end{aligned} \tag{15b}$$

The set of equations in (15) provides a basis for solving for the coefficients of the B(s) polynomial. As found above for waveform-function sampling, Eq. (15) is also homogeneous in the B(s) coefficients, and thus requires additional information, or a constraint, to solve it. Lacking any convincing reason to do otherwise, we choose to set b_P to unity, which results in right-hand side entries for the successive equations of Eq. (15) given by $-f_{P+1}$, $-f_{P+2}$, . . ., $-f_{D-1}$, respectively. The resulting relationships then defined by Eq. (15) again take the form of a predictor equation, but where now P lower-order derivatives and the P predictor coefficients provide a derivative one order higher, although other coefficients constraints would lead to other relationships. Again, we need $D \geq 2P$ for a solution.

The above discussion shows that either function sampling or derivative sampling can be used to estimate the parameters of an exponential series. While function sampling might be achieved analytically, numerically or experimentally, derivative sampling is limited to the former two approaches as there seems to be no practical way of measuring, say, the time derivatives of a transient waveform. Derivative sampling is none-the-less relevant, as significant computational advantages can arise from using derivative information in the SD, as is discussed further in RIII. Although a similar advantage might also be found for using derivatives in WD modeling, our primary purpose in mentioning it here is to emphasize the symmetry that exists between function and derivative sampling and between the WD and SD.

3.3 Combining Waveform-Domain Function Sampling and Derivative Sampling

As a generalization of the above discussion, it should be noted that it is also possible to combine function and derivative sampling of WD data. Letting $f_{i,n}$ represent the n'th derivative of f(x) at $x = x_i$ leads to

$$f_{i,n} = \Sigma R_\alpha (s_\alpha)^n (X_\alpha)^i,\ \alpha = 1, \ldots, P;\ i = 0, \ldots, D;\ n = 0, \ldots, F \tag{16}$$

where it is assumed that there are F derivatives at each of d sample locations in x and where n = 0 is the usual function sample. The same number of derivatives is not needed at each sample point, but this restriction makes the development more straightforward. The following equations are then obtained,

$$
\begin{aligned}
f_{0,0} &= R_1 + R_2 + \cdots + R_P, \\
f_{0,1} &= R_1(s_1)^1 + R_2(s_2)^1 + \cdots + R_P(s_P)^1 \\
&\vdots \\
f_{0,F} &= R_1(s_1)^F + R_2(s_2)^F + \cdots + R_P(s_P)^F \\
f_{1,0} &= R_1(X_1)^1 + R_2(X_2)^1 + \cdots + R_P(X_P)^1 \\
f_{1,1} &= R_1(s_1)^1(X_1)^1 + R_2(s_2)^1(X_2)^1 + \cdots + R_P(s_P)^1(X_P)^1 \\
&\vdots \\
f_{1,F} &= R_1(s_1)^F(X_1)^1 + R_2(s_2)^F(X_2)^1 + \cdots + R_P(s_P)^F(X_1)^1 \\
&\vdots \\
f_{D,0} &= R_1(X_1)^D + R_2(X_2)^D + \cdots + R_P(X_P)^D, \\
f_{D,1} &= R_1(s_1)^1(X_1)^D + R_2(s_2)^1(X_2)^D + \cdots + R_P(s_P)^1(X_P)^D \\
&\vdots \\
f_{D,F} &= R_1(s_1)^F(X_1)^D + R_2(s_2)^F(X_2)^D + \cdots + R_P(s_P)^F(X_P)^D.
\end{aligned} \tag{17}
$$

The polynomial that might be used to reduce this system of equations is not obvious as was the case for function sampling or derivative sampling alone. At least two distinct choices might be considered, as

$$P_1(X,s) = A(X)B(s) \tag{18a}$$

and

$$P_2(X,s) = C(X,s) \tag{18b}$$

the difference being in the number of independent coefficients for given highest-order terms in X and s. The product form of (18a), for example, which might be viewed as a separation-of-variable form, yields ~ D + F coefficients whereas that of (18b) has instead ~ DF. Thus expanding $P_1(X,s)$ in terms of A(X) and B(s), we obtain

$$\begin{aligned} P_1(X,s) &= (a_0 + a_1X + a_2X^2 + \cdots + a_{D-1}X^{D-1} + a_DX^D)(b_0 + b_1s + b_2s^2 + \cdots \\ &\quad + b_{F-1}s^{F-1} + b_Fs^F) = a_0b_0 + a_1b_0X + a_0b_1s + a_2b_0X^2 + a_1b_1Xs + a_0b_2s^2 \\ &\quad + \cdots + a_Db_FX^Ds^F = 0 \end{aligned} \tag{19a}$$

where the cross-term (in X and s) coefficients involve various products and sums of the A(X) and B(s) coefficients for a total of D + F + 2 independent coefficients.

If, on the other hand, we use (18b), we would have

$$P_2(X,s) = c_{0,0} + c_{1,0}X + c_{0,1}s + c_{2,0}X^2 + c_{1,1}Xs + c_{0,2}s^2 + \cdots + c_{D,F}X^Ds^F = 0 \tag{19b}$$

which has a total of DF + 1 coefficients. Clearly, either approach leads to a linear system of equations for the polynomial coefficients, but without testing it's not obvious which of these, or possibly another, might be better.

Anticipating encountering a similar question in SD sampling where various combinations of function and derivative sampling can be used and have been found to be successful, some fundamental questions arise concerning the most effective sampling strategy in the WD. First, a minimum of 2P samples are required of a waveform signal having P pole terms whatever combination of function and derivative sampling has been used to generate them. Second, the sample spacing must satisfy the Nyquist rate for the highest-frequency component of the signal, i.e., $\delta \leq 1/2f_{max}$ where f_{max} is the maximum frequency, at least in a generalized sense, noting that this restriction is irrelevant when all samples are taken at a single frequency. Finally, the data samples must be spaced such as to obtain a polynomial model. A uniform spacing δ is generally used, but any sequence of sample spacings that are related as integer multiples of δ could be used in principle.

A critical question in considering alternate sampling strategies, both those that involve function sampling as well as function and derivative sampling, is "how much information is provided by each additional sample?" Some sampling strategies are clearly less effective. For example, if 2P samples were to be used but at an arbitrarily closer spacing $\delta' << 1/2f_{max}$, then the information provided by each sample shifts further to the right in the digits representing that sample's numerical value. This is because a given sample approaches the values of its neighbors in the limit $\delta' \rightarrow 0$ so that all sample values would eventually become identical. The consequence of decreasing sample spacing is that proportionately higher-accuracy samples would be needed to retain a fixed amount of information when performing function sampling.

It seems likely that derivative sampling might be subject to similar kinds of constraints. One of the first questions to consider might be how the information content of a derivative sample varies as a function of derivative order, and is there a highest "reasonable" order to use? Another question would be how the proper Nyquist rate changes with increasing derivative order. From a computational viewpoint it would also be important to know how function and derivative samples vary with respect to information content and whether there is some combination of both that maximizes the acquired information while minimizing the computer costs of evaluating the FPM.

Proceeding as before, successively multiplying the data-sample equations by the set of coefficients from the chosen polynomial model, equations in the characteristic-equation coefficients can be derived. For example, if we were to use the product-form polynomial in A(X) and B(s) as in Eq. (18a), equations like the following

$$a_0 b_0 f_{0,0} + a_1 b_0 f_{1,0} + a_0 b_1 f_{0,1} + \cdots a_{D-1} b_{F-1} f_{D-1,F-1} + f_{D,F} = 0 \qquad (19c)$$

would be obtained. Thus, the initial solution of the data matrix would provide intermediate results in terms of the coefficient products which could be subsequently solved for the individual coefficients themselves. We conclude our discussion of combined waveform function and derivative sampling at this point.

4.0 SAMPLING FIRST-PRINCIPLE MODELS AND OBSERVABLES IN THE SPECTRAL DOMAIN

The use of SD data for MBPE is conceptually similar to sampling in the WD, but because the FM for the former is the transform of the latter, significant differences arise, as is now shown where function and derivative sampling are respectively described.

4.1 Spectral-Domain Function Sampling

Spectral-domain function sampling begins with the FM given by Eq. (2) and assumes the availability of samples denoted by

$$F_i = F(X_i) = \Sigma R_\alpha/(X_i - s_\alpha) + F_{np}(X_i),\ \alpha = 1, \ldots, P \tag{20}$$

where, in contrast to waveform sampling, there is no requirement that the sampling points X_i be uniformly spaced. However, in contrast to waveform sampling, where the FM can be a purely exponential series at late times, spectral sampling can not avoid the presence of the non-pole term which is generally unknown. Fortunately, an approximation to the F_{np} term can be realized by generalizing the pole series in Eq. (20) [Miller and Burke (1991)]. First note that, as written, the pole series can be developed into a particular rational function where the denominator order exceeds that of the numerator by one. A general rational function has no specified connection between the orders of the polynomials which comprise it, however. The capability of a pole series to model resonances can be retained while changing the numerator-polynomial order relative to that of the denominator. Note that the special case of using rational functions as a model is also known as Padé approximation [Press et al. (1992)] when the function and derivative sampling is done at a single point although the term has been applied to the more general approach considered here [Kukulin et al. (1989)].

The possibility of using various numerator and denominator polynomial orders provides a way to approximate the effect of the non-pole term by simply increasing the order of the numerator polynomial. For example, an increase of one in the numerator order has the effect of representing F_{np} by a constant, which, when absorbed into the rational function, results in equal numerator and denominator orders. If F_{np} is represented by a constant and a term linear in X, this has the effect of making the numerator order one greater than the denominator. Thus, by varying the relative orders of the polynomials which comprise the FM, various models of the non-pole, SD contribution are implicitly included. This rather simple way of handling the non-pole contribution is not suitable for WD data unfortunately since the early-time, or driven, response contains essentially all frequency components whereas the above approach in the SD needs only to be effective over a relatively narrow frequency band.

Therefore, in general we use a SD FM given by

$$F(X) = \mathbb{N}(X)/\mathbb{D}(X) \tag{21a}$$

where

$$\mathbb{N}(X) = N_0 + N_1X + N_2X^2 + \cdots + N_nX^n, \tag{21b}$$

and $$\mathbb{D}(X) = D_0 + D_1X + D_2X^2 + \cdots + D_dX^d. \tag{21c}$$

The coefficients of the SD FM are also obtained from sampled values of the response. How this is done is easy to see by rewriting Eq. (21) as

$$F_i\mathbb{D}_i = \mathbb{N}_i,\ i = 0, \ldots, D - 1 \tag{22a}$$

where

$$F_i = F(X_i), \tag{22b}$$

$$\mathbb{D}_i = D_0 + D_1X_i + D_2(X_i)^2 + \cdots + D_d(X_i)^d, \tag{22c}$$

and $$\mathbb{N}_i = N_0 + N_1X_i + N_2(X_i)^2 + \cdots + N_n(X_i)^n. \tag{22d}$$

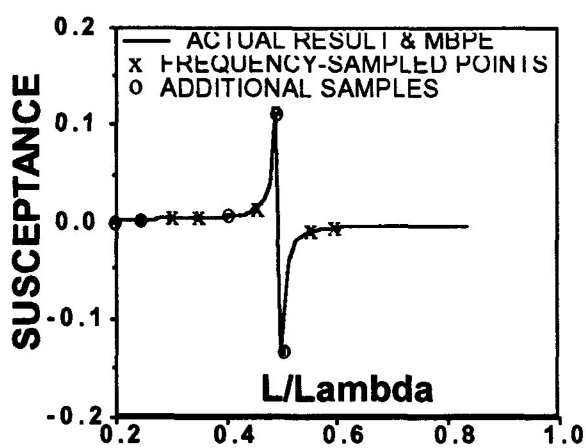
There are $d + n + 2$ unknown coefficients in the two polynomials $\mathbb{D}(X)$ and $\mathbb{N}(X)$, and as for the previous cases, a constraint or additional condition is needed to make the sampled equations inhomogeneous. Again, there is no unique choice for this constraint, but if we set $D_d = 1$, then the following equations result:

$$F_0D_0 + X_0F_0D_1 + \ldots + F_0(X_0)^{d-1}D_{d-1} - N_0 - X_0N_1 - \ldots - (X_0)^nN_n = -(X_0)^dF_0$$
$$F_1D_0 + X_1F_1D_1 + \ldots + F_1(X_1)^{d-1}D_{d-1} - N_0 - X_1N_1 - \ldots - (X_1)^nN_n = -(X_1)^dF_1$$
$$\vdots$$
$$F_{D-1}D_0 + X_{D-1}F_{D-1}D_1 + \ldots + F_{D-1}(X_{D-1})^{d-1}D_{d-1} - N_0 - X_{D-1}N_1 - \ldots$$
$$- (X_{D-1})^nN_n = -(X_{D-1})^dF_{D-1} \tag{23}$$

where $D \geq n + d + 1$ is again required. Note that the matrix coefficients are now comprised of a product of a data sample and the frequency at which the sample is taken raised to a power, in contrast to the time-domain situation where the data samples alone are the matrix coefficients. Also observe that the poles in the SD arise directly as the roots of $D(x)$ whereas in the WD, the poles are natural logarithms of the roots of the characteristic equation, Eq. (6). The exponentiation of the sampling frequencies suggests that large dynamic numerical ranges in the matrix coefficients may result if d and n are very large. One way to avoid this is to scale the frequency, so that, for example, if the sampling range is centered at 1 GHz, a scaling of 10^9 in the frequencies leads to nominal scaled values near unity. It is also possible to center the SD model about a frequency in the interval of interest so that terms like $(X_i - X_{ref})^n$ result. Combining scaling and translation similarly produces terms like $[(X_i - X_{ref})/X_{ref}]^n$. A simple example of using function sampling in the SD is shown in Fig. 2.

An over-sampled system, i.e., one where $D > n + d + 1$, can be handled in various ways, one of which is to employ a pseudo inverse for the solution. Another approach would be to employ overlapping windows of different data sets to compare performance of their respective FMs. A third would be to progressively increase the number of data samples while retaining the same number of FM coefficients and comparing the FM spectra to observe their trends.

Derivative sampling can also be performed in the SD, as can various combinations of function and derivative sampling. One result is that the samples can be spaced more widely when derivative information is available. A more important consideration is that in some circumstances a derivative sample can be obtained for a computation operation count that is of order $1/X_i$ of the first function sample alone. Thus, if a derivative sample provides information concerning the response equivalent to that provided by a function sample, an obvious computation advantage arises from using derivative samples to decrease the number of X_i sample points that are needed. Using derivative sampling in the context of a frequency-domain FPM is discussed more fully in RII.



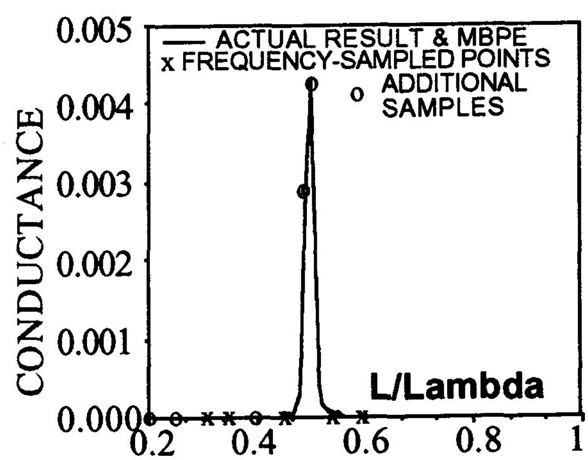


Figure 2. Example of applying rational-function FM to admittance results in the vicinity of the first resonance of a dipole antenna. The diamonds indicate the original FPM samples used for a single 5-coefficient FM computation (d = n = 2) while the additional points shown as squares are subsequent FPM samples added to test the accuracy of the FM.

4.2 Spectral-domain derivative sampling

The set of possible sampling strategies concludes with derivative sampling in the SD. In this case, we rewrite Eq. (21a) as

$$F(X)\mathbb{D}(X) = \mathbb{N}(X) \qquad \text{(21a)'}$$

which upon differentiating with respect to X t times leads to

$$F'\mathbb{D} + F\mathbb{D}' = \mathbb{N}'$$
$$F''\mathbb{D} + 2F'\mathbb{D}' + F\mathbb{D}'' = \mathbb{N}''$$
$$F'''\mathbb{D} + 3F''\mathbb{D}' + 3F'\mathbb{D}'' + F\mathbb{D}''' = \mathbb{N}'''$$
$$\vdots$$
$$F^{(t)}\mathbb{D} + tF^{(t-1)}\mathbb{D}^{(1)} + \ldots + C_{t,t-m}F^{(m)}\mathbb{D}^{(t-m)} + \ldots + F\mathbb{D}^{(t)} = \mathbb{N}^{(t)} \qquad (24)$$

where $C_{r,s} = r!/[s!(r-s)!]$ is the binomial coefficient and we require both that the $t + 1$ samples $F^{(0)}, \ldots, F^{(t)}$ be such that $t + 1 \geq n + d + 1$ and that one of the coefficients is set to unity for a unique solution to be obtained. Clearly these also become a linear system of equations for the polynomial coefficients which can be expressed in the form

$$N_0 = F_0$$
$$N_1 - F_0D_1 = F_1$$
$$N_2 - F_1D_1 - F_0D_2 - = F_2$$
$$\vdots$$
$$N_n - F_{n-1}D_1 - F_{n-2}D_2 - \ldots - F_{n-d}D_d = F_n$$
$$-F_{n+1}D_1 - F_nD_2 - \ldots - F_{n-d+1}D_d = F_{n+1}$$
$$\vdots$$
$$-F_{D-2}D_1 - F_{D-3}D_2 - \ldots - F_{D-d}D_d = F_{D-1} \quad (25)$$

where the substitution $X' = X - X_s$ has been made, with X_s the sampling frequency, so that all frequency-dependent terms drop out from the final equations and D_0 has been set to unity. The subscripted "F" quantities in the above matrix are $F_m = F^{(m)}/m!$. An example of using SD derivative sampling of a simulated transfer function is shown in Fig. 3.

A still more general sampling strategy can be employed where two or more frequencies are used at each of which a variable number of frequency derivatives can be independently specified. This leads to a more complicated system of equations [Miller and Burke (1991)], but which is numerically simply still another way to compute the FM parameters, i.e., the FM and the data used for its quantification in the SD can be independently specified. An example of two-frequency, function-derivative sampling is shown in Fig. 4. This simple test case is somewhat anomalous in that sampling at a single frequency has produced an apparently more accurate match between the simulated spectrum and the FM than when using two frequencies. The computational advantages of using SD derivative sampling from a CEM viewpoint are discussed in RIII.

5.0 DISCUSSION

The preceding discussion has focused on the original version of Prony's Method when applied to WD function sampling and its logical extension when WD derivative sampling can be done, as well as the corresponding SD variants of this basic approach. This was done because the basic Prony Method is so directly connected to the relevant wave-equation observables and it works very well when applied to data of sufficient accuracy. Some extensions of Prony's Method that enable better processing of noisy data than achievable using the original version are discussed by Kay and Marple (1981). Extensions like those described by Kay and Marple vary in detail, but include such features as exploiting oversampled data by employing a pseudo-inverse solution or "pre-filtering" the data by constructing auto covariance estimates from the data.

Other approaches are also applicable to both the WD and SD data and FMs discussed above for dealing with noisy data, one of which is the pencil-of-functions method [Sarkar et al. (1980)] and the matrix-pencil technique [Sarkar and Pereira (1995)]. The matrix-pencil technique is identical to the basic Prony Method when the number of data samples is twice the number of poles and the data is noise-free, but the two differ when the problem is over-determined (i.e., when more data samples are used) or when the data is noisy or of limited accuracy. In particular, the variance of the pole estimates is less for the matrix pencil than for the basic Prony version which can be a distinct advantage for attempting to extract EM parameters from available data.

However, it should be noted that the applications envisioned here and to be discussed in RII have as their primary goal not determining numerical values for the EM parameters represented by poles and residues, but rather developing a continuous WD or SD, low-order, analytical model for interpolating the available data. In that case, the match achieved between the FM and the data being modeled is a more relevant measure of performance than uncertainty in the EM parameters themselves. As discussed by Dudley and Goodman (1986), high-variance estimates of the poles and residues can result from noisy data even though that data is well-matched by the FM. Thus, accurate parameter estimates are not necessarily required for productive use to be made of a FM where the application is to represent EM observables rather than to obtain the EM parameters themselves. Furthermore, the data noise that arises from the limited numerical convergence of the FPM computation itself would generally be low enough that it should not invalidate the FM

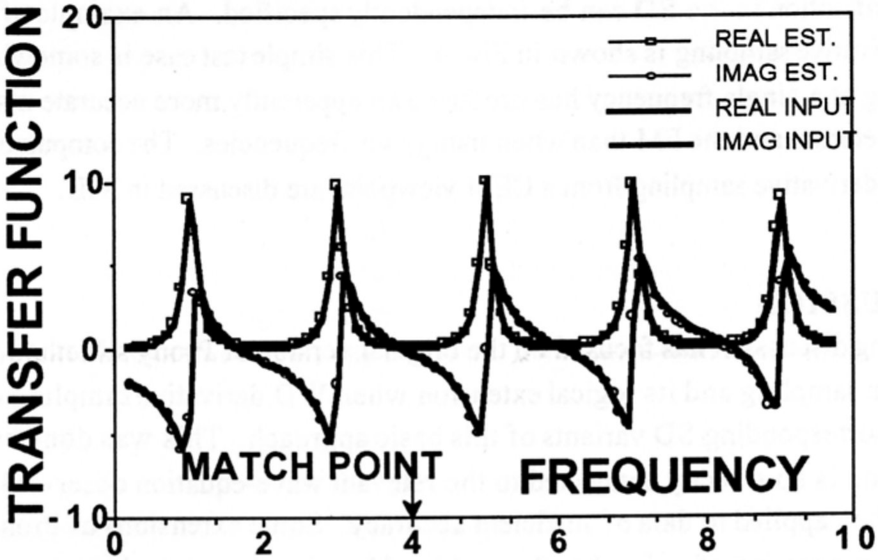


Figure 3. Results for a simulated spectrum having 10 poles (5 conjugate pairs) modeled with a 10-coefficient FM (n = 4 and d = 5) using one function sample and 9 derivative samples at the point shown by the arrow where the frequency = 4. The specified spectra are shown as lines without data points while the FM results are denoted by squares (the real part) and diamonds (the imaginary part) respectively.

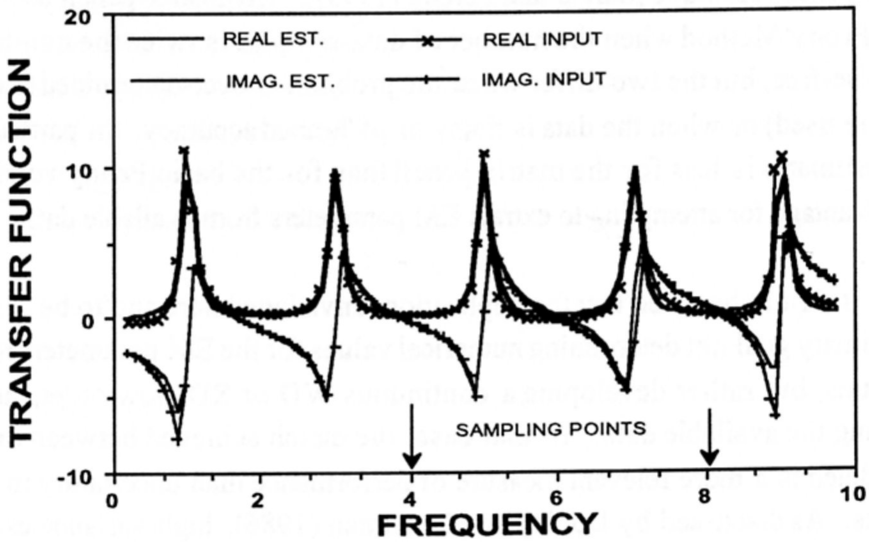


Figure 4. Results for the same 10-pole model as used in Fig. 2 but where the 10-coefficient FM (n = 4 and d = 5) is based on a function sample and the first four frequency derivatives at frequencies of 4 and 8, respectively. The data samples are indicated by lines having x's (the real part) and tickmarks (the imaginary part) while the FM estimates are shown as continuous lines.

computations. If the FPM data is not accurate enough to be modeled, the limitation is not one caused by MBPE.

These observations suggest another potential application of FMs in CEM. Any numerical computation is subject to various errors that might be characterized as numerical noise, as contrasted with the measurement noise that comes from experimental instrumentation. One of the applications of SD MBPE discussed in RII is filtering noisy frequency-domain measured data. This suggests the possibility that numerical noise may be similarly "filtered" from FPM computations, a procedure that would require oversampling of the FPM to obtain redundant information. Aside from increasing the accuracy of the final result beyond that provided by the FPM samples, this procedure also provides an opportunity of estimating the numerical accuracy of the FPM itself. One approach would be to compute the root-mean-square, or some other difference, between the FPM samples and the FM. While the oversampling required to do this might not be worth the extra cost for every application of a given FPM, this procedure could be used whenever a significantly different problem is modeled with it as a means of numerical validation.

6.0 CONCLUDING REMARKS

This article has provided a rationale and background for using model-based parameter estimation (MBPE) in electromagnetics. It was explained that MBPE involves use of a physically motivated, reduced-order, analytical description (the model) whose coefficients (the parameters) are obtained by fitting it to samples of the process being modeled. Two kinds of fitting models (FMs), consisting of exponential and pole series and which form a transform pair between the generalized waveform domain (WD) and the spectral domain (SD), respectively, were outlined, as suggested by the physics of wave-equation solutions. It was shown that the data used for FM quantification can separately involve samples of a process, which was referred to as function sampling, or derivatives of that process, referred to as derivative sampling, or combinations thereof.

In part II of this series [Miller (1995)], the application of MBPE to a variety of electromagnetic observables will be demonstrated. Part III [Miller (1996)] will discuss how the signal-processing concepts exemplified by MBPE might be used for developing more efficient numerical models in electromagnetics. The underlying basis for these applications is that the detailed, microscopic description of electromagnetic fields provided by Maxwell's Equations or FPMs naturally generate reduced-order macroscopic observables whose analytical description can be accurately achieved using much simpler FMs. A major benefit of such FMs is that they can reduce the number of samples needed to develop a continuous representation of a response, such as a frequency spectrum or radiation pattern. Another is that they can provide an approach for reducing the computational effort associated with the numerical evaluation of first-principles models (FPMs).

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EDITORIAL

Computational Electromagnetics and Wireless Personal Communications (Part IV)

In the past three editorials we have addressed system level issues concerning the role of computational electromagnetics in the wireless communication revolution. We have addressed general systems architecture (though we were very brief and this area needs to be expanded), antennas, and bioelectromagnetic effects. We address in this article the role that computational electromagnetics could have in the design of wireless communications microelectronics. It is here where wireless communications has had its greatest advances. As stated in the first article of this series, advances in microelectronics (specially deep sub-micron design) has made it possible to put whole functional blocks of communication systems into a single chip reducing tremendously the complexities and size of such systems. RF chips and digital signal processing chips are now a common occurrence.

Computational electromagnetics (CEM) can help in the "microelectronic" design of wireless personal communications systems in two ways: 1) by helping in the design of high speed circuits which are needed to quickly process the large amount of digital data originally obtained and handled by RF communications systems, and 2) by protecting the chips themselves from the adverse effects of electromagnetic radiation (this electromagnetic radiation will come at the intrasystems and intersystem level). In this article we will address the former. In the next issue (Part V) we will address the effects of electromagnetic energy coupling into microelectronics at the most fundamental levels.

CEM and High Speed Circuit Design: I remember many years ago when I was an undergraduate sitting in my first class in electromagnetic theory. The teacher, in an attempt to bring enthusiasm to this class that was known to bring pain to undergraduates, said: "for those of you who are in love with circuits and electronic design, I want to tell you that all electrical signals are electromagnetic in nature. The reason why your electronic designs work is because at low frequencies the physics of electromagnetic fields can be represented by lumped-parameters models; at high frequencies that is no longer true. You see guys, Kirchoff's laws come from electromagnetic theory under the assumption that the rate of change of electromagnetic fields is slow...we need to bring electromagnetic to electronics". After seeing the blank look on our faces, I thing he got discouraged too because we never saw an application of what he said.

It is estimated that interconnect delays account for 80% of the cycle time in digital circuits while only 20% account for the switching (set up time + hold time) of the gates. As frequency increases the interconnects (pins+traces+connectors) size becomes comparable in size with wavelength. In the past such interconnects were modeled as combinations of R's, C's, and L's. You can actually make complex ladder networks of these passive elements to understand such phenomena as ground bounce, reflections...etc. The problem is that it is very difficult to model and accurately account for so many parasitic effects---the lumped parameter models become too complex, not only to understand but also to analyze since it will also consume considerable amount of computing resources for such simulators as SPICE and the like. You can actually "see" these parasitic effects by looking at the characteristics of high speed circuits as shown in Table 1.

The following are high speed effects of circuits that you can "see" in your design.

1) The PCB or MCM only works at low frequencies
2) The PCB and MCM works only within a narrow frequency range
3) When you change vendor parts, it will not work as well or not work at all
4) Temperature changes make a big difference on your design
5) The design is peculiar to the type of connectors and parts you use
6) Small changes in power supply voltages can make a big difference
7) Touching or bringing the hand closer to the board can affect the performance
8) Adding bypass capacitors can cause significant changes in performance
9) The board radiates a lot and is sensitive to EMI
10) Things work fine alone. When you connect to system/other components, it won't work well or will not work at all.

Table 1. Diagnosing High Speed Circuits

Furthermore, as a rule of thumb, you may also start observing problems in high speed design when the propagation delay of the interconnects is about 25% of the rise and fall time. In order to diminish these problems multichip modules (MCM) are emerging as one of the most useful packaging technologies since the surface mount devices. With MCM you eliminate chip packaging with its associated parasitics and they also decrease the size and number of interconnects in a considerable manner. Problems with MCM, however, are the high density of crossover wiring, high resistivity of the connects and parasitics due to high wiring density.

At high frequencies, lumped parameter representations are no longer useful and the need arises for using computational electromagnetic methods. For years the design approach for boards was that described in Figure 1. This approach was only good when the interconnects parasitics could be modeled as lumped parameters. The conceptual design was followed by analysis of logic, timing, and critical paths. Analog and other simulations were performed to determine the logical/timing accuracy of the design. The logical design was converted to a physical layout. The physical layout tools were used to extract the parasitic which were then represented as lumped parasitics to account for interconnects loading. These lumped parasitics were dumped back into the simulators to check out the accuracy of the design. In high speed circuits detailed interconnect and signal integrity analysis is also required, as shown in Figure 2. For good signal integrity analysis CEM field solvers are needed, in addition to models of all the electronics and electrical network solvers. In the process of Figure 2 the layout database is scanned for interconnects, coupling structures and discontinuities. The CEM field solvers are then used to obtain a mathematical representation of the equivalent electrical circuits of physical structures on the model. All this information together with the electronic's library information of the semiconductor components is then dumped into the network solver for accurate simulation of the design.

The biggest challenge in CEM is to make the process described in Figure 2 as seamless as possible and this will require the development of "smart" CEM field solvers that would discern which paths would be the most critical for a particular design and skip the rest.

Reinaldo Perez
ACES Newsletter Chief Editor

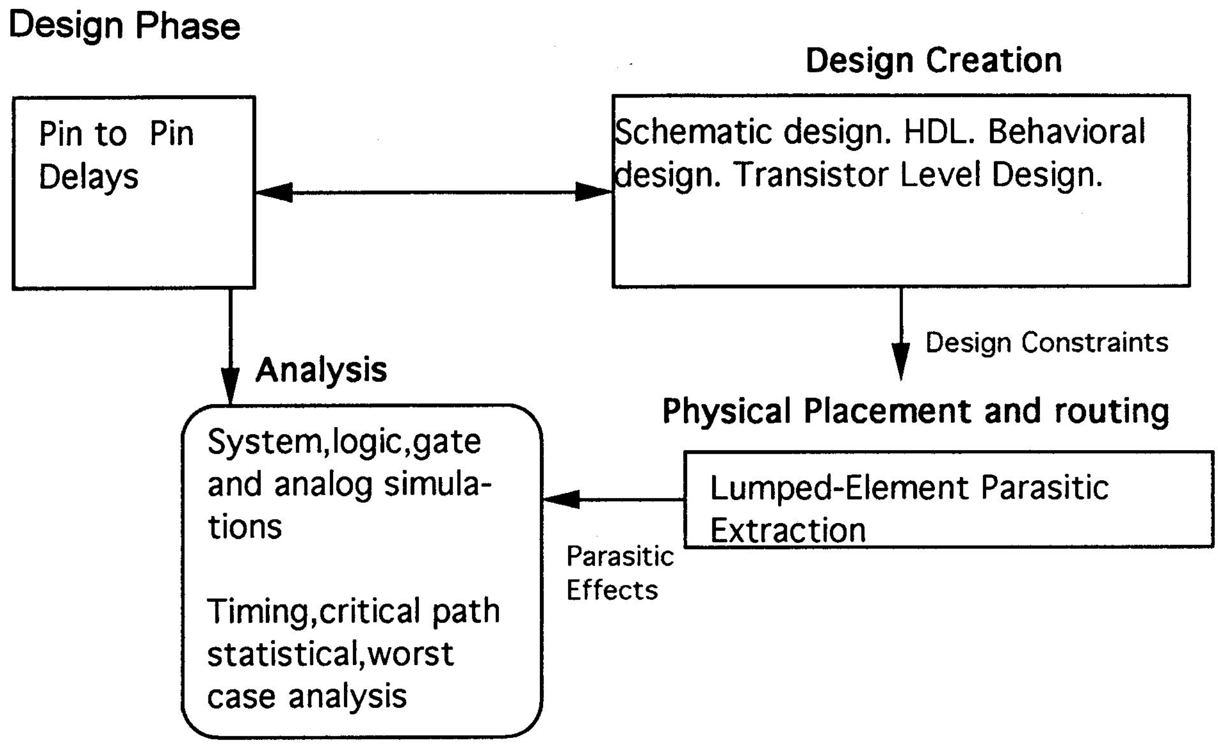

Figure 1. Classical Design Approach in Microelectronic Design

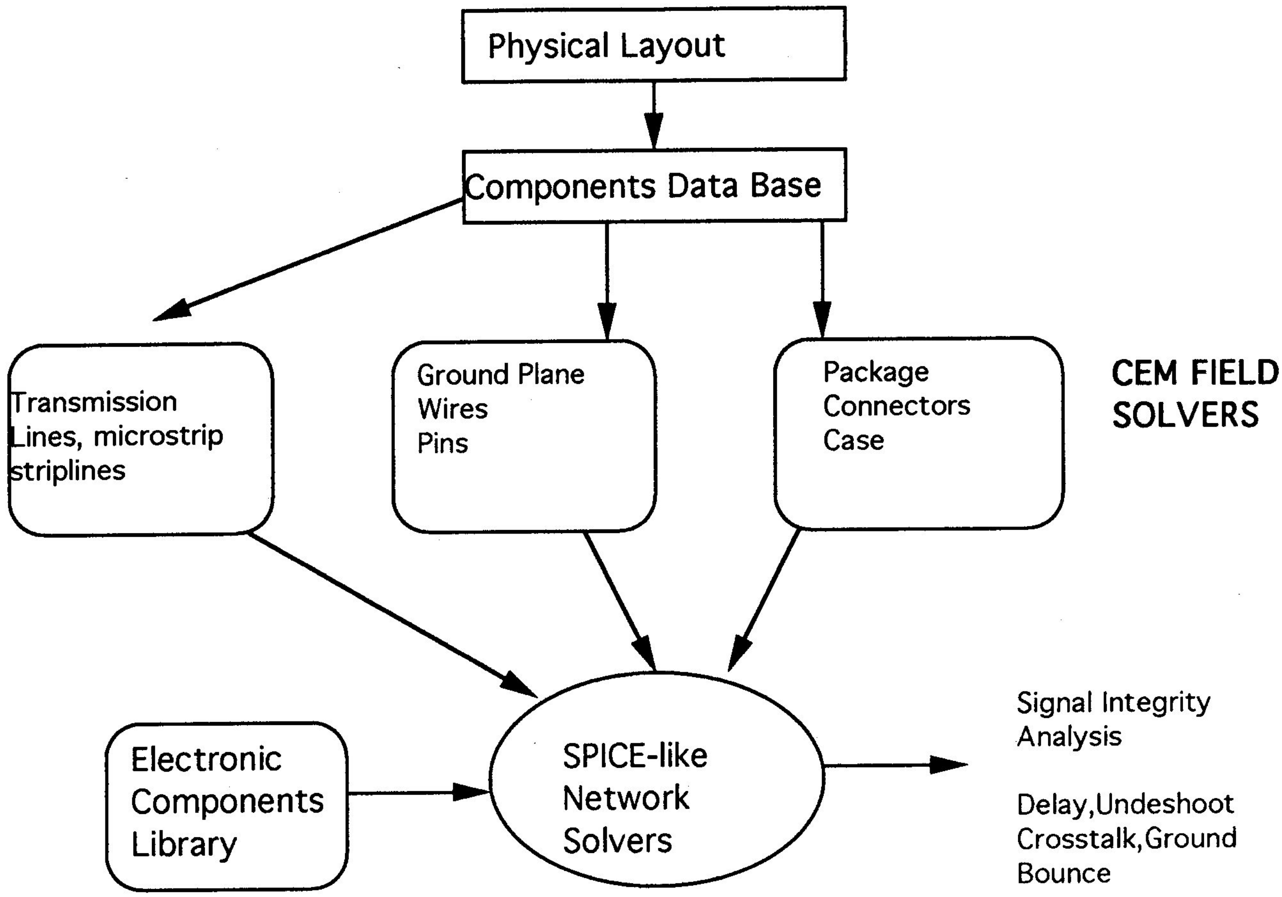

Figure 2. Usefulness of CEM in the Design of Microelectronics

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SHORT COURSES AT THE 12TH ANNUAL REVIEW OF PROGRESS IN APPLIED COMPUTATIONAL ELECTROMAGNETICS

The Applied Computational Electromagnetics Society (ACES) is pleased to announce eight short courses to be offered with its annual meeting of March 18 - 22, 1996. The short courses will be held on Monday and Saturday. Short Course Registration begins at 7:30 AM on Monday, 18 March. Registration by mail is suggested! [Note: Tuesday through Friday will be technical sessions and vendor exhibits], ACES has the right to cancel a course at any time with full refund. For further information contact Rob Lee, Short Course Chairman. (see complete address on page 69. Fee for a short course will be $90 for a half-day short course and $140 for a full-day course, if booked before Friday, March 1, 1996. **NOTE: Short Course attendance is NOT covered by the Symposium Registration Fee!**

COURSE INFORMATION

1. **AN APPLICATION ORIENTED INTRODUCTION TO THE NEC-BSC WORKBENCH**, Half-day, by Ron J. Marhefka and Lee W. Henderson, The Ohio State University. ,

The NEC-BSC Workbench is a windows based graphical user interface for creating and manipulating input files for NEC-BSC. The NEC-BSC input commands are displayed in an edit window, and the actual geometry is displayed in separate wireframe views. The user can also accomplish dialog box editing of commands. This Short Course will demonstrate the use of the Workbench with practical applications, using realistic geometries. The course will emphasize how the interaction with the Workbench allows the user to easily employ the full capabilities of the NEC-BSC.

2. **USING MODEL-BASED PARAMETER ESTIMATION TO INCREASE THE EFFICIENCY AND EFFECTIVENESS OF COMPUTATIONAL ELECTROMAGNETICS**, Half-day by Ed Miller,

Hidden beneath the mathematical detail associated with most electromagnetic analysis is the possibility of representing physical observables in simpler ways using reduced-order models. Knowledge of such models can be helpful in ways ranging from reducing the computer cost of achieving desired solutions to developing more compact representations of observables. The basis approach is to estimate unknown parameters of the models from sampled data, a process called "model-based parameter estimation" (MBPE). This lecture demonstrates some of the benefits that results, expanding on recent articles by the author in the ACES Journal and Newsletter.

3. **WAVELETS: THEORY, ALGORITHMS, AND APPLICATIONS**, Full-day, by Andrew K. Chan, Texas A&M

Wavelet Analysis is one of the most exciting topics to emerge from mathematical research that has a wide range of engineering applications. Because of its flexible time-frequency window, the wavelet transform complements the shortcomings of Fourier-based techniques. In signal processing applications, wavelets are used in speech compression, echo-cancellation, music processing, etc. Their applications in electromagnetic problems are relatively new. In particular, they have been applied for processing data from electromagnetic scattering and for matrix compression in solving some integral equations. This course is aimed at providing an overview of wavelet analysis along with algorithms and applications. The first part of the course will begin with a brief review of Fourier analysis and short-time Fourier analysis. Construction of orthonormal and semi-orthogonal spline wavelets based on the multi-resolution analysis will be discussed. The second part of the course is devoted fast algorithms and applications of wavelet analysis.

4. **CONFORMAL TIME DOMAIN NUMERICAL ELECTROMAGNETICS**, Full-day, by Kane Yee, Lockheed.

The workshop will provide a coherent account of the development of the finite difference time domain (FDTD) and its generalization in solving Maxwell's equations. The generalized FDTD, which is based on the surface-curve integral form of the Maxwell's equations, will be emphasized in the derivation of the numerical algorithms. The finite volume time domain (FVTD), which is based on the volume-surface integral forms of the Maxwell's equations, can be very convenient when unstructured grids are employed. Boundary condition simulation will be emphasized.

5. **ANTENNA PROPERTIES IN LINEAR AND NONLINEAR ENVIRONMENTS**, by Robert Bevensee, BOMA Enterprises.

General theorems for an antenna as a transmitter and as a receiver-scatterer in an electromagnetically linear environment will be reviewed and illustrated. A hypothesis about the best gain-bandwidth behavior possible within a given electrical working volume will be developed. For an antenna operating in an electrically linear environment, relations among transmitted powers at various frequencies with nonlinear control port loads will be derived via the Manley-Rowe Relations.

The difficulty of developing an upperbound as opposed to a lowerbound to the bistatic scattered power of an N-port antenna will be discussed. The approximate nature of the Optical Theorem will be demonstrated. For an antenna operating as a receiver in an electrically linear environment, relations among collected (extinction), scattered, and load powers with a nonlinear load will be derived via the Manly-Rowe Relations.

6. **FINITE ELEMENT FOR ELECTROMAGNETICS**, Full-day, by John Volakis, Univ of Michigan and John Brauer, MacNeal-Schwendler Corporation.

The course will develop and apply two-dimensional and three-dimensional finite elements, both nodal-based and edge-based. Local and global mesh truncation techniques will be examined including the new perfectly matched layer method. Applications will include antennas, scattering, microwave circuits, nonlinear magnetic apparatus, electronic packaging, and electromagnetic compatibility.

7. **APPLICATION OF MODERN ANALYTICAL AND HYBRID TOOLS FOR ANTENNA MODELING AND SYNTHESIS**, Half-day, by Roberto Rojas and Prabhakar Pathak, The Ohio State University.

8. **MATHEMATICAL SOFTWARE FOR COMPUTATIONAL ELECTROMAGNETICS**, Full-day, by Jovan Lebaric, The Naval Postgraduate School.

The objective would be to introduce **MATLAB** and **MATHCAD** software for research and teaching of computational electromagnetics. Each attendee will have a PC with **MATLAB** and **MATHCAD** software installed. Attendees would learn how to use the commercially available state-of-the-art mathematical software to solve static, transient and time-harmonic electromagnetic problems efficiently on a PC. The numerical techniques presented will be Method of Moments (MOM) and Finite Differences (FD). Attendees will be provided with sample solutions and working programs, but will be asked to extend or solve problems on their own.

THE APPLIED COMPUTATIONAL ELECTROMAGNETICS SOCIETY
12TH ANNUAL REVIEW OF PROGRESS
IN APPLIED COMPUTATIONAL ELECTROMAGNETICS

March 18-22, 1996
Naval Postgraduate School
Monterey, CA

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	BEFORE 3/1/96	3/1/96 TO 3/11/96	AFTER 3/11/96
ACES MEMBER	☐ $245	☐ $260	☐ $275
NON-MEMBER	☐ $285	☐ $300	☐ $315
STUDENT * (no proceedings)	☐ $115	☐ $115	☐ $115
RETIRED/UNEMPLOYED	☐ $150 (includes proc.)	☐ $150	☐ $150
BANQUET ☐ meat ☐ fish	☐ $ 30	☐ $ 30	☐ $ 30

Short Courses

Course Fees do NOT include attendance at the symposium. Short courses can be taken without attendance at symposium.
Fees for a half-day and full-day course are: $90 or $140, before 3/1/96; $100 or $150, 3/1/96-3/11/96; $110 or $160 after 3/11/96

WAVELETS: THEORY, ALGORITHMS, AND APPLICATIONS, by A.K. Chan Full-day	☐ $140	☐ $150	☐ $160
CONFORMAL TIME DOMAIN NUMERICAL ELECTROMAGNETICS, by K. Yee Full-day	☐ $140	☐ $150	☐ $160
ANTENNA PROPERTIES IN LINEAR AND NONLINEAR ENVIRONMENTS, by R. Bevensee Full-day	☐ $140	☐ $150	☐ $160
FINITE ELEMENT METHODS FOR ELECTROMAGNETICS, by J. Volakis and J. Brauer Full-day	☐ $140	☐ $150	☐ $160
MATHEMATICAL SOFTWARE FOR COMPUTATIONAL ELECTROMAGNETICS, by J. Lebaric Full-day,	☐ $140	☐ $150	☐ $160
AN APPLICATION ORIENTED INTRODUCTION TO THE NEC-BSC WORKBENCH by R. Marhefka and L. Henderson, Half-day	☐ $ 90	☐ $100	☐ $110
USING MODEL-BASED PARAMETER ESTIMATION TO INCREASE THE EFFICIENCY AND EFFECTIVENESS OF COMPUTATIONAL ELECTROMAGNETICS, by E. Miller, Half-day	☐ $ 90	☐ $100	☐ $110
APPLICATION OF MODERN ANALYTICAL AND HYBRID TOOLS FOR ANTENNA MODELING AND SYNTHESIS, by R. Rojas and P. Pathak, Half-day	☐ $ 90	☐ $100	☐ $110

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FIRESIDE LODGE (**) (1 star)
1131 10th St. Monterey, CA 93940
(408) 373-4172
Govt. Rate $55 (Tax: see below)
Conference Rate: $69 + 10% tax.

STAGECOACH MOTEL (**) (1 Star)
1111 10th St. Monterey, CA 93940
(408) 373-3632
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Conference rate: $49 S, $59D +10% tax

HYATT HOTEL & RESORT (**) (4 Star)
1 Old Golf Course Rd. Monterey, CA 93940
(408) 372-1234
Govt. Rate: $74 S, $101.50 D (Tax: see below)
Conference Rate: $105 S, $127 D + 10% tax.

HOLIDAY INN (**) (3 Star)
1000 Aguajito Rd. Monterey, CA 93940
(408) 373-6141
Govt. Rate $74 (Tax: see below)
Conference Rate: $89 + 10% tax

SUPER 8 MOTEL (2 Star)
2050 Fremont St. Monterey, CA. 93940
(408) 373-3081
Govt. & Conference Rate $39.60 Single + Tax
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Those attending on Govt. Travel Orders, can avoid paying CITY TAX if: (1) payment is made via government issued AMEX card; (2) have a copy of travel orders with them; and (3) have govt/military identification. Re Govt. rates: prevailing perdiem at time of arrival will be honored.

When you book a room mention that you are attending the "**ACES**" Conference, and ask for either government, or conference rates.

There is **NO CONFERENCE PARKING** at the Naval Postgraduate School or nearby streets, so we advise you to book a room within walking distance, or plan to use a taxi.

TO ALL AMATEUR RADIO OPERATORS IN ACES

PAPER SESSION AT NEXT ACES CONFERENCE

Perry Wheless is seeking authors for a paper session at the ACES '96 Symposium in Monterey (March 1996) on amateur radio antenna analysis and design. This is an opportunity for hams in ACES to combine their work and hobby, and to perpetuate the time-honored ACES tradition of interest in HF Antennas.

"We typically have 20-30 hams attend the annual symposium. I hope we can increase that number next year, and have a paper session in 1996. If I can play some role in that by coordinating a paper session for 1996, I would be pleased to get the ball rolling and then turn it over to others for continued development in the future.

Prospective authors should note from the Call for Papers that October 27 is the NEW deadline for summaries (300-500 words). FAX and e-mail submissions are possible, so I hope the deadline will pose no major problem to you. The final papers for the Proceedings are not required until early January 1996.

Contact me at your earliest convenience if you require additional information, or would like to discuss your idea for a paper in advance of submission. My telephone number is (205) 348-1757, FAX is (205) 348-6959 and e-mail is wwheless@ua1vm.ua.edu (note that the '1' in ua1mv is the numeral one and not the letter l).

It is intended that a social event will be organized for hams at ACES '96. An open invitation will be issued by way of the ACES Newsletter later, but you are encouraged to make known your interest in such an event, along with any suggestions for a preferred data and/or location.

I hope that you will be able to join us in Monterey next March. The ACES Symposium is a special event in a special place, and the atmosphere is always friendly. It is a pleasant and informative experience, which draws many regulars back to Monterey year after year! C U there".

Perry Wheless, 73 de K4CWW
University of Alabama
College of Engineering
317 Houser Hall, Box 870286
Tuscaloosa, AL 35487-0286

For information regarding ACES or to become a member in the Applied Computational Electromagnetics Society, contact Dr. Richard W. Adler. ECE Department, Code EC/AB, Naval Postgraduate School, 833 Dyer Rd, Rm. 437, Monterey, CA. 93943-5121, telephone (408) 646-1111, Fax: (408) 649-0300. E-mail: rwa@mcimail.com. You can subscribe to the Journal and become a member of ACES by completing and returning the form below.

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