APPLIED COMPUTATIONAL ELECTROMAGNETICS SOCIETY (ACES)

NEWSLETTER

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NEWSLETTER ARTICLES AND VOLUNTEERS WELCOME

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

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OFFICER'S REPORTS

PRESIDENT'S MESSAGE

As many of you have already experienced, a certainty (in addition to death and taxes) of sufficiently long employment involving Computational Electromagnetics is that your domicile decays and falls down while you are away from home working all the time. Pat Adler has been forced to shake my tree vigorously for this inexcusably late report because my years of procrastination have caught up with me, and my wife has all of this old dog's spare time right now committed to learning some new tricks of the home repair and maintenance variety. This, together with a fairly heavy travel schedule, has made me pretty hard to find lately. I expect to get back to some regular time for ACES by mid-August, and should be readily accessible for ACES business again by then. Meanwhile, I ask for your patience and sympathy if you are emailing or calling about ACES, and I promise to get back to you as quickly as possible.

The ACES 2000 conference was a big hit, and congratulations to everybody on the conference team for their success. The California weather was exceptionally cooperative this year, and I have only heard positive feedback from those who attended the conference. Leo Kempel (kempel@egr.msu.edu) is leading the charge forward "into the fog" toward ACES 2001, and he would like to hear from you if you are interesting in becoming more actively or deeply involved in the conference event. For all the latest conference buzz, check out the ACES web site (aces.ee.olemiss.edu), where a 2001 Conference Short Course Survey is currently available and the most comprehensive conference info can be found at all times.

You may be interested to know of some recent changes and appointments. The terms of three ACES Directors expired in March – Andreas Cangellaris, Ray Perez, and Norio Takahashi. On behalf of ACES, our collective thanks are extended to them for their three years of service on the Board of Directors. Osama Mohammed, Tapan Sarkar, and Masanori Koshiba are the newly elected replacements and, similarly, congratulations and a warm welcome are extended to these new Directors.

Andy Peterson (Peterson@ee.gatech.edu) started work as the new Publications Chair in March, and is interested in new Special Issue projects and other initiatives to move ACES Publications to the next level. I urge you to contact Andy if you have ideas and/or ambitions in the Publications area. Melinda Piket-May has agreed to assume the duties of Finance Chair, Pat Foster is new Awards Chair, and Doug Werner is new Conference Chair. We gratefully acknowledge the time and outstanding service rendered to ACES by the outgoing chairs (Bob Bevensee of Conference, Andy Peterson of Finance, and John Brauer of Awards).

Finally, it is both relevant and a pleasure to report that the financial report received at the March Board of Directors meeting shows ACES to be on solid ground and continuing to gain strength, thanks in large part to the continuing popularity and success of the annual conference in Monterey. The Directors are actively looking into ways of enhancing the value of ACES membership, and the possibility of actually reducing membership dues in the future – more later!

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SECRETARY'S REPORT

ACES BOARD OF DIRECTORS MEETING MINUTES

1. The Applied Computational Electromagnetics Society's Board of Directors held its annual symposium business meeting 20 March 2000 at 1730, in 101A Spanagel Hall, Naval Postgraduate School. The meeting was called to order by President W. Perry Wheless, Jr.

Present were Directors Perry Wheless, Norio Takahashi, Andreas Cangellaris, Bruce Archambeault, Jianming Jin, Allen Glisson, Robert Bevensee, Ray Perez, Ping Werner, Doug Werner, Leo Kempel, Atef Elsherbeni, Keith lysiak, and ACES Executive Officer Richard W. Adler.

2. The first order of business was to elect the officers of the Board of Directors. The following were elected for a two year term:

President	W. Perry Wheless. Jr.
Vice President	Eric Michielssen
Treasurer	Allen Glisson
Secretary	Bruce Archambeault

3. Corrections were made to the October BOD meeting as follows:
— change item #3 to show the membership at 400
— change item #6 to 'conference schedule' from 'conference handbook'

A motion to adopt these changes was made by Alan Glisson and seconded by Andreas Cangellaris. A vote was taken, and the motion passed.

4. Andrew Peterson presented the budget report.
a) membership is down from previous year, but dues are up. This is due to late renewals and a number of military voucher forms taking a long time to pay
b) publication expenses are down from last year. This is mostly due to better prices and a reduced volume of pages.
c) The reserves are healthy.
d) conference costs were down mostly because we did the refreshments ourselves. Printing costs were also down.

A motion to accept the budget report was made by Allen Glisson and seconded by Andreas Cangellaris. A vote was taken, and the motion passed.

5. The Publications committee meeting and dinner the previous evening cost appx. $170. Perry suggested ACES might pay for this. A motion to pay the $170 for the publications committee dinner was made by Alan Glisson and seconded by Andreas Cangellaris. A vote was taken, and the motion passed.

6. Perry Wheless commented that the ACES web page was in great shape. Atef Elsherbeni reviewed the two year proposal he made last year. He pointed out that many expenses are not covered by ACES. The budget for last year was $10k, and this year the request is for $13k. Some details were provided:
a) the web server is a 550 MHz Pentium III with a 10 Gbit hard disk
b) the development machine is not funded by ACES
c) detailed list of new requests totaling $13k
d) 1998 and 1999 journal articles are on the server, and 1997 will be soon
e) 75 conference articles are on-line
f) the cost for a vendor to organize all pdf files and provide 500 CDs is $15 each

A motion to continue funding the on-line web effort was made by Alan Glisson and seconded by Andreas Cangellaris. A vote was taken, and the motion passed.

7. Ray Perez started a discussion about conference papers being available for download to ACES members only. Others can only download abstracts. Ray Perez, Atef Elsherbeni and the publications chairman are to prepare a member survey before May and recommend a program to convert newsletter to softcopy.

A discussion about requiring electronic submission for conference and to request electronic submission for journal articles (although not required) followed. A motion to approve this was made by Alan Glisson and seconded by Andreas Cangellaris. A vote was taken, and the motion passed.

8. Appointments
   a) Andy Peterson accepted interim chairman of the Publications Committee
   b) Doug Werner accepted the Conference Committee chairmanship

9. There was a discussion to have the 2001 conference provide both paper and CD publications. Possibly all ACES members could have the proceedings CD mailed to them after the conference. All conference attendees get the hard copy and can purchase the CD if desired.

a) Ray Perez reported that of the 10 hardcopy responses to the survey, all were positive for electronic publishing of the newsletter.

10. Doug Werner reported on the 2000 Conference.
   a) 140 papers were submitted
   b) APS 2000 is in competition with lots of other conferences
   c) Registration as of Tuesday was 170
   d) The Copyright and Author Registration should be with the electronic submittals

11. ACES 2001 conference chair will be Leo Kempel
   — Keith Lysiak will be publicity chair
   — Ed Rothwell will be co-chair
   — Tim Holzhimer will be vendor chair
   — We need someone to organize the short courses

12. A discussion about changing the conference schedule to match the Naval Postgraduate School's academic calendar was presented. Perry Wheless will do a survey of the BOD after the conference

13. There was a discussion about reducing membership dues. It was decided there would be no change now, and we need to have at least one special issue per year.

14. The following BoD directors' terms ended for Andreas Cangellaris, Ray Perez, and Norio Takahashi. They were thanked for their help, support, and efforts to ACES.

15. The following BOD directors terms will be over in 2001; Bruce Archambeault, Tony Brown, Eric Michielssen

16. The student winner of the Best Student Paper Contest will receive a $200 check; and she/he and the advisor will be entitled to a free conference registration and a free short course registration for 2001. Specifics of the student paper contest will be reviewed by the conference committee, and further changes may result from their study conclusions and recommendations.

17. Committee Appointments:
Membership and Chapter Development: Anthony Brown, UK, and Guiseppe Pelosi, Italy

18. A request to sponsor the FEM and Microwave Engineering Conference was discussed. A motion was made by Allen Glisson and seconded by Andreas Cangellaris. A vote was taken, and the motion passed.

A motion to adjourn was made at 7:40 PM.

Respectfully submitted
Bruce Archambeault

ACES ANNUAL MEMBERS MEETING MINUTES

The meeting was held on 7:50 AM on Tuesday, 21 March 2000.

The Financial Report was read. Allen Glisson moved to accept the report and the motion was seconded by Andrew Peterson. A vote was taken and the motion passed.

The meeting was adjourned at 7:57 AM.

Respectively submitted

Bruce Archambeault

PERMANENT STANDING COMMITTEES OF ACES INC.

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COMMITTEE REPORTS

AWARDS COMMITTEE

At the 2000 16th Annual Review of Progress in Applied Computational Electromagnetics, an Awards Banquet was held to recognize ACES members and conference attendees who have made special contributions to the Society. President W. Perry Wheless, Jr. presented two Service Awards, along with awards to authors of three special categories papers.

THE 2000 VALUED SERVICE AWARD was presented to Randy Haupt, for his dedicated leadership as Chairman of the 15th Annual Review of Progress.

THE 2000 MAINSTAY AWARD was presented to Masanori Koshiba for organizing the recent highly successful ACES-Japan Short Course.

THE 1999 OUTSTANDING JOURNAL PAPER AWARD was presented to Daniel Reuster, Gary Thiele, and Paul Eloe, for their paper, "On the Iteration of Surface Currents and the Magnetic Field Integral Equation" published in Vol. 14, No. 3, November 1999, pp. 76-83, ACES Journal.

THE $200 BEST STUDENT PAPER CONTEST AWARD was presented to Alkim Akyurtlu for her paper on "A New FDTD Scheme to Model Chiral Media", which appears on pages 181-188 in Volume 1 of the conference Proceedings. Her co-authors were D.H. Werner and K. Aydin. Along with the check, a free Conference Registration and a free Short Course Registration for C2001 is included.

THE WINNERS OF THE $500. BEST (PROCEEDINGS) PAPER AWARD was presented to Jun Fan, James Knighten, J. Drewniak, and Hao Shi, from the University of Missouri at Rolla. The title of this paper is "An MPIE-based Circuit Extraction Technique and its Applications on Power Bus Modeling in High-Speed Digital Designs, and appears on pages 130-137 in Volume 1 of the Proceedings.

Please join us in thanking these winners for their outstanding service. The opinions of all members are solicited in selecting all award winners, so feel free to send in your nominations for 2001.

Pat Foster, Awards Chair

NOMINATIONS COMMITTEE

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' STATEMENT (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.

DIRECTORS-AT-LARGE

Bruce Archambeault	2001	Allen W. Glisson	2002	Masanori Koshiba	2003
Anthony Brown	2001	Guiseppe Pelosi	2002	Osama Mohammed	2003
Eric Michielssen	2001	Perry Wheless, Jr.	2002	Tapan Sarkar	2003

Adalbert Konrad, Nominations Chair

PUBLICATIONS COMMITTEE

Greetings from the Publications Committee! At the March conference, I allowed Perry Wheless to twist my arm and traded the Chairship of Finance for Publications. Best of luck to Melinda Piket-May, who is assuming the helm at Finance!!

In my view, the ACES Journal and Newsletter have improved in both appearance and content during the past few years. What has not improved, and what I feel needs attention, is the circulation of these publications. Since circulation is essentially limited to membership (libraries count as institutional members), the gradual decline of our membership rolls is of continuing concern. One idea for invigorating the subscription rate of our publications is a renewed activity in Special Issues. I propose to target niche topics associated with CEM in an attempt to raise the visibility of ACES publications, and I look forward to reader suggestions as to what topics are appropriate and consistent with that goal. If anyone has other ideas, please send them along!

Due to the efforts of Atef Elsherbeni, ACES has an active web site and is exploring ways to use the site to assist in our publications activities. Not only will it provide an electronic archive for pubs, it will serve as a means for electronic submission and review of manuscripts. I encourage you to visit the site (http://aces.ee.olemiss.edu) and get up to speed on these features!

As always, if any of you are interested in becoming more active in the area of publications (coordinating a column for the Newsletter, serving as an Editor or Associate Editor, etc.) please let me know and I will do what I can to facilitate a role for you to play.

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MODELER'S NOTES

Gerald J. Burke

There is not much to report for Modeler's Notes for this issue, but since it did not appear in the last issue I will write something. A couple of NEC-4 bug problems came up, but it turned out they were old ones resurfacing. John Grebenkemper found that the radiation pattern angles in his single precision NEC-4 were not being incremented. This was related to a problem reported in the July 1998 Newsletter in which the angles were not reset in a frequency loop. The fix reported in that newsletter was correct, but I made the change wrong in our single precision version for Windows. The double precision NEC4D was changed correctly. The correct section of code from the main program, 14 lines before the statement `54 CONTINUE`, is:

```
IF (IFLOW.EQ.7) THEN
   IF (IXTYP.GT.0.AND.IXTYP.LT.4) GO TO 54
   PSOR1(NSCINC)=THSAVE
   PSOR2(NSCINC)=PHSAVE
```

In our NEC4S for Windows the second IF statement was put after the next two statements restoring values of PSOR1 and PSOR2, so that the values were restored every time around the loop and the angles could not change. There were probably a reasonable number of codes that went out that way, but maybe not too many people use the single precision code.

Another problem was reported by Mike Morgan at NPS who was computing near magnetic fields over ground and getting isolated glitches in the field values. This is an old problem that results from the code switching between lookup tables, model fitting and asymptotic approximations in the evaluation of the field over ground. The switching may result in small jumps in the near electric field, but these become spikes in magnetic field when H is evaluated from a finite difference approximation of the curl of E. One of the worst glitches occurs when the interaction distance via ground is one wavelength, and that is what was causing Mike's problems. This turns out to be caused by an attempt to reduce computation time by switching from integrating over the segments to using a point-source approximation. The way to turn off the switch and eliminate the glitch was reported in the July 1997 Newsletter. However, I had forgotten about that until rediscovering the problem. The fix is to comment out four lines from subroutine EFLD as shown below.

```
IF(IPERF.EQ.2)THEN
          IMAGF=1
          RHO=SQRT(XIJ*XIJ+YIJ*YIJ)*GSCAL
          ZIJS=ZIJ*GSCAL
C         IF((ZIJS.GT.ZZMX1).OR.(-ZIJS.GT.ZPMX1).OR.(RHO.GT.RHMX1))THEN
C            IMAGF=2
C            GO TO 17
C         END IF
       END IF
```

I will make this change in the distribution code, since the increase in computation time should be minor, and it can cause serious problems in near H. A couple of other people besides Mike Morgan, have called with H field glitch problems in the past year and I did not think of this particular case. There are other sources of glitches in H, especially at low frequencies, so I hope to get a chance to set up a direct evaluation of H over ground. The time for evaluating Sommerfeld integrals should not be too significant. For E it is about five times longer than for the table lookup, but the present H evaluation needs six E evaluations for the central difference approximation of $\nabla \times E$, so it should come out about even.

In previous Newsletters I have mentioned problems in compiling NEC and its plotting programs with the DEC/Compaq Visual Fortran V. 6 compiler. Originally using DVF V. 5 it went very smoothly. The only problem that we encountered with the V. 5 compiled code was that on some Windows NT systems the text that the program wrote to the screen was invisible. That happened on a whole classroom of computers at the short course at the 1999 ACES Conference at NPS. According to Compaq Support that is a known bug, and the display can be made visible by changing the "number of colors" setting in Windows.

I updated to DVF V. 6 to fix the invisible writing problem but encountered some new quirks. If you entered data on the screen and then tried to delete it, the right half of each character was left on the screen. The entry was really deleted, but they wrote black rectangles over the characters on the screen and got the position wrong. Apparently that only happened when compiling as a "Standard Graphics Application," but that is used with or without graphics so that files can be opened with a dialog box. That was a new problem to Compaq, but they provided a source code work-around and later an object code patch called QWGFWND.OBJ. This fixed the delete problem but turned the cursor into an hourglass, which was a bit disconcerting since on the Mac that often means the system is hung. Compaq has since released V. 6.1 of their compiler, which is a free upgrade for V 6.0 users. It fixes the cursor problem but not the delete problem, which will be fixed in a future release, so you still have to link with QWGFWND.OBJ for Standard Graphics Applications.

Another problem with the V. 6 compiler was that the "clear screen" command missed a strip of about 3/16 inch at the bottom of my screen. Compaq Support guessed that it might be related to the graphics driver, although it did not happen with the V. 5 compiler. They are probably right, since it does not happen when I run the code on a different Windows 98 system instead of my Windows 95 system at home. However, I have not yet taken their suggestion of updating the Matrox Millennium driver to see if that fixes it under Windows 95.

Since the fun with DVF seems to be winding down, I just upgraded my Fortran compiler for the Mac to Absoft Pro Fortran 6.2. It requires OS 9 or at least part of the OS 9 library, so I updated the operating system. However, ATM (needed for *Textures* TeX) is not compatible with OS 9, and Acrobat Reader will not run, which makes it hard to read the documentation. For ATM, you need to update to V. 4.5 and then install an OS 9 update from Adobe. The compiler compiles NEC-4 as a single file and fairly quickly. With the old compiler you had to split NEC into separate subroutine files to avoid overflowing compiler tables. I have not yet gotten it to link with the routine to open files with a dialog box. When that is solved I will try the LAPACK routines optimized for the G4 "Velocity Engine" and see how they compare with the results that Tom Wallace reported in the last issue.

Anyone who can contribute material on modeling-related issues for future newsletters is encouraged to contact our editor Ray Perez or Jerry Burke, Lawrence Livermore National Lab., P.O. Box 808,L-154, phone: 925-422-8414, FAX: 925-423-3144, email: burke2@llnl.gov.

TECHNICAL FEATURE ARTICLE

POWER CONSIDERATIONS FOR THE MEASUREMENT OF NONLINEAR EFFECTS

Donald D. Weiner[1]
Andrew L. Drozd[2]

ABSTRACT

This is the second in a series of articles which explores the analysis and modeling of nonlinear behaviors in circuits, devices, and receiver systems. Discrete and quasi-discrete methods can be developed to readily analyze complex nonlinearities from elemental formulations such as the weakly nonlinear series. The topics discussed are quite general and have application to such diverse areas as automatic control, broadcasting, cable television, communications, EMC, electronic devices, instrumentation, signal processing, and systems theory.

INTRODUCTION

The previous paper in this series discussed the nonlinear effects of intermodulation, spurious responses, desensitization, cross modulation, and gain compression/expansion [1]. In the laboratory the severity of these effects is typically evaluated through the use of power measurements. Unfortunately, this approach can be very confusing to the uninitiated because there are several different definitions of power and power gain. In particular, one may speak of either average power, available power and/or exchangeable power. These, in turn, are used in definitions of average power gain, insertion power gain, transducer power gain, and exchangeable power gain. Although the gains become identical in certain specialized situations, they are not, in general, equivalent. Consequently, the various powers and power gains cannot be used interchangeably.

INSTANTANEOUS, AVERAGE, AND COMPLEX POWERS

Recall that the complex voltage

$$E = |E|\, e^{j\alpha}$$

is associated with the sinusoidal voltage

$$e(t) = |E| \cos(2\pi ft + \alpha)$$

while the complex current

$$I = |I|\, e^{j\beta}$$

is associated with the sinusoidal current

$$i(t) = |I| \cos(2\pi ft + \beta).$$

Let e(t) and i(t) be the sinusoidal voltage and current associated with the linear network having the impedance Z, as shown in figure 1. By definition, the instantaneous power delivered to the linear network is the product of the voltage across the input terminal pair multiplied by the current entering the positive terminal. Specifically, the instantaneous power is

$$p(t) = e(t)\, i(t) =$$

$$1/2\, |E||I| \cos(\alpha - \beta) + 1/2\, |E||I| \cos[2\pi(2f)t + \alpha + \beta].$$

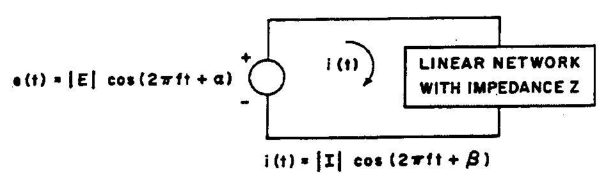

Figure 1. A linear network in the sinusoidal steady state with impedance Z.

Thus, in the sinusoidal steady state, the instantaneous power consists of a constant plus a sinusoid at twice the frequency of the voltage and current.

In many applications, the time average of the instantaneous power is of interest. This is referred to as the average power and is given by

$$p_{ave} = 1/2\, |E||I| \cos(\alpha - \beta) =$$

$$1/2\, \mathrm{Re}\{EI^*\} = 1/2\, \mathrm{Re}\{E^*I\}$$

where Re{•} denotes the operation of taking the real part of the argument. The average power is seen to depend upon both the magnitudes and angles of the complex voltage and current. The complex power is defined to be

$$p_{com} = 1/2\, EI^*$$

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such that

$$p_{ave} = \mathrm{Re}\,\{p_{com}\}.$$

The concept of complex power is extremely useful in deriving various expressions for the average power delivered to a network. For example, let the linear network of Figure 1 have the impedance

$$Z = R + jX$$

where R is the resistance and X is the reactance of the impedance. From Ohm's law,

$$E = IZ.$$

Use of Ohm's law in the expression for the complex power yields

$$p_{ave} = 1/2\ |I|^2\ \mathrm{Re}\{Z\} = 1/2\ |I|^2R.$$

Although the latter result looks very much like the instantaneous power formula for a resistor, we must remember that I is a complex current and that R is the real part of a complex impedance.

The admittance of the linear network of Figure 1 is

$$Y = 1/Z = G + jB$$

where G is the conductance and B is the susceptance of the admittance. Using the fact that

$$I = EY,$$

the average power can, also, be expressed as

$$p_{ave} = 1/2\ |E|^2\ \mathrm{Re}\{Y\} = 1/2\ |E|^2G.$$

However, in general, the conductance of a network does not equal the reciprocal of its resistance. The correct relationship between G and R is

$$G = R/(R^2 + X^2).$$

Consequently,

$$p_{ave} = 1/2\ |E|^2\ R/(R^2 + X^2).$$

For the special case in which the impedance is real, Z = R and X = 0, the average power for this very special situation is

$$p_{ave} = 1/2\ |E|^2/R$$

and is valid only for those networks whose impedances are purely resistive.

AVERAGE POWER GAIN

A 2-port network, as shown in Figure 2, is commonly used in the transmission of power from a signal source to a load. Because the network is assumed to be operating in the sinusoidal steady state, circuit variables are indicated as complex voltages and currents. It is assumed that no independent sources are contained within the 2-port. The signal source driving the 2-port is modeled as a Thevenin equivalent circuit consisting of the ideal voltage source E_S in series with the generator impedance Z_S. The linear load is designated by the impedance Z_L. The behavior of the linear 2-port is completely characterized by the open-circuit impedance parameters according to the equations

$$E_1 = Z_{11}I_1 + Z_{12}I_2$$

$$E_2 = Z_{21}I_1 + Z_{22}I_2.$$

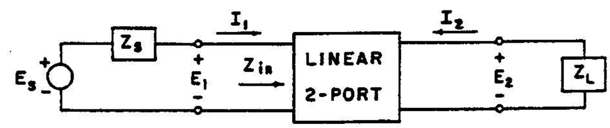


Figure 2. A loaded linear 2-port in the sinusoidal steady state.

Of interest is the amount of power delivered to the load as compared to that generated by the source. Although we have assumed that no independent generators are contained within the 2-port, controlled sources are allowed. Therefore, power gains greater than unity are possible. For passive 2-ports, where the gain is less than unity, the power loss is defined to be the reciprocal of the power gain.

With reference to Figure 2, let

p_L = average power delivered to the load from the 2-port,
p_S = average power delivered to the 2-port from the source,
g_P = average power gain of the loaded 2-port.
The average power gain is defined as

$$g_P = p_L/p_S.$$

To pursue the topic further, let the load and input impedances of the 2-port be given by

$$Z_L = R_L + jX_L$$

$$Z_{in} = R_{in} + jX_{in}.$$

From our discussion of average power, it follows that

$$g_P = |I_2/I_1|^2 R_L/R_{in}.$$

g_P is seen to be the product of the magnitude squared of the 2-port current gain times the ratio of the real part of the load impedance to the real part of the input impedance.

An interesting relation results when the 2-port and load are entirely resistive. With

$$X_L = X_{in} = 0,$$

R_L and R_{in} can be expressed as

$$R_L = -E_2/I_2 \text{ and } R_{in} = E_1/I_1.$$

In addition, the current gain for a resistive network is purely real. Therefore, the expression for the average power gain reduces to

$$g_P = (E_2/E_1)(-I_2/I_1).$$

This is recognized as the product of the loaded 2-port voltage and current gains. The minus sign is associated with I_2 since $-I_2$ represents the current flowing from the 2-port into the load. In general, the above expression for g_P is not valid when the 2-port and/or load contain energy storage elements.

An alternate expression for the average power gain is obtained by utilizing the load and input admittance of the 2-port which are given by

$$Y_L = 1/Z_L = G_L + jB_L$$

$$Y_{in} = 1/Z_{in} = G_{in} + jB_{in}.$$

Then

$$g_P = |E_2/E_1|^2 G_L/G_{in}.$$

In this form g_P is the product of the magnitude squared of the 2-port voltage gain times the ratio of the real part of the load admittance to the real part of the input admittance.

Finally, in terms of the 2-port open-circuit impedance parameters, it can be shown that

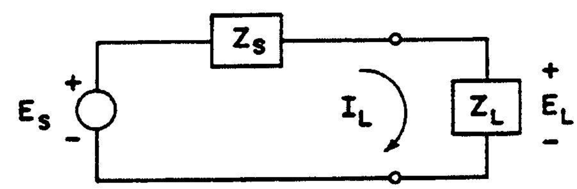
$$g_P = |Z_{21}/(Z_{22} + Z_L)|^2 R_L/\text{Re}\ \{Z_{11} - Z_{12}Z_{21}/(Z_{22} + Z_L)\}.$$

Observe that the source impedance does not appear in this expression. This is a consequence of both the 2-port current gain and the input impedance being independent of Z_S. In this sense, average power gain is a figure of merit for the loaded 2-port irrespective of the generator exciting the network.

INSERTION POWER GAIN

The 2-port of Figure 2 serves as an interface for the power transfer from source to load. Another way of evaluating 2-port performance is to compare average powers delivered to the load with and without the 2-port in place. For this purpose, consider the circuit of Figure 3 in which the load is connected directly to the source. Let $\tilde{p}_L$ denote the average power delivered to the load without the 2-port in place.

Figure 3. Circuit of Figure 2 with 2-port



removed.

With reference to Figure 3,

$$\tilde{p}_L = 1/2\ |I_L|^2 R_L = 1/2\ |E_S/(Z_S + Z_L)|^2 R_L.$$

On the other hand, with respect to Figure 2,

$$p_L = 1/2\ |E_2|^2 R_L/(R_L^2 + X_L^2) = 1/2\ |E_2|^2 R_L/|Z_L|^2.$$

The insertion power gain is defined to be

$$g_I = p_L/\tilde{p}_L.$$

It follows that

$$g_I = |E_2/E_S|^2\ |(Z_S + Z_L)/Z_L|^2.$$

In many applications, it is common to have identical generator and load impedances. The insertion gain then reduces to

$$g_I = 4\ |E_2/E_S|^2.$$

Observe that the voltage E_2 is associated with the circuit of Figure 2 while the voltage E_S is associated with that of Figure 3. Nevertheless, since the generator is the same in both circuits, it is possible to interpret the ratio E_2/E_S as a voltage gain involving the loaded 2-port of Figure 2.

In terms of the 2-port open-circuit impedance parameters, it can be shown that

$$g_I = |Z_{21}(Z_S + Z_L)/[(Z_{11} + Z_S)(Z_{22} + Z_L) - Z_{12}Z_{21}]|^2.$$

Note that g_I does depend upon the source impedance Z_S whereas the average power gain, g_P, does not. Therefore, given the parameters characterizing the 2-port and load, it is possible to maximize the insertion power gain with respect to the generator impedance. This is not true for the average power gain.

The average and insertion power gains differ in a rather significant way. In the definition of average power gain one of the powers is dissipated in the load while the other is dissipated at the input to the 2-port. However, in the definition of insertion power gain, both powers are dissipated in the load. From an experimental point of view, measurement of average power gain involves a single circuit. On the other hand, measurement of insertion power gain requires that the load be removed from the 2-port and connected to the source. This may, or may not, be convenient depending upon the application.

AVAILABLE POWER GAIN

In this section we introduce the concept of available power. Consider the situation shown in Figure 3 in which the generator is connected directly to the load. Let the load and generator impedances be given by

$$Z_S = R_S + jX_S$$

$$Z_L = R_L + jX_L.$$

Assume the generator impedance is fixed, but that the load impedance is to be chosen so as to maximize the average power delivered to the load. Obviously, the choice

$$Z_L = -R_S - jX_S$$

results in infinite power flow. To eliminate this physically unrealizable solution from consideration, we require both the load and generator impedances to correspond to passive networks. This, in turn, requires that both R_S and R_L be non-negative. The unrealizable solution is no longer acceptable since a positive value of R_S results in a negative value of R_L which is prohibited. The maximum power delivered by the generator is now finite and is defined to be the generator's available power, denoted by p_A.

It can be shown that the maximum power is delivered to the load when

$$Z_L = R_S - jX_S$$

which is the complex conjugate of the source impedance. This choice for Z_L is referred to as a conjugate impedance match. It follows that the available power of the generator is

$$p_A = |E_S|^2 / 8R_S.$$

It is important to emphasize that available power is a property of the generator alone. A generator may be characterized in terms of its available power whether or not it is connected to a conjugate-matched load. Of course, when the load is conjugate matched, the average power delivered by the generator is its available power. Otherwise, a lesser amount is delivered. Available power is an extremely convenient concept because it can be determined independent of the network to which the generator is connected.

We now introduce the concept of available power gain. With reference to Figure 2, let

p_{AO} = available power from the 2-port
p_{AS} = available power from the generator
g_A = available power gain of the loaded 2-port.

The available power gain is then defined as

$$g_A = p_{AO}/p_{AS}.$$

In terms of the 2-port open-circuit impedance parameters, it can be shown that

$$g_A = |Z_{21}/(Z_{11} + Z_S)|^2 R_S/\mathrm{Re}\,\{Z_{22} - Z_{12}Z_{21}/(Z_{11} + Z_S)\}.$$

A comparison of this expression with the corresponding expression for the average power gain is enlightening. The two expressions are identical in form. In fact, g_A can be obtained from g_P by interchanging Z_{22} with Z_{11} and Z_L with Z_S. Whereas g_P is independent of Z_S, g_A is independent of Z_L. To put it another way, whereas average power gain depends only on that portion of the network to the right of the generator, available power gain depends only on that portion to the left of the load.

A word of caution is in order. Available power gain must not be confused with actual power flow in the network. Available power flows from the source only when the 2-port input impedance is conjugate matched to the generator. Similarly, available power flows from the 2-port only when the load is conjugate matched to the 2-port output impedance. In general, these conditions are not met and available powers do not flow. Nevertheless, available power gain may be used to characterize a loaded 2-port even under unmatched conditions. This is often done when it is desirable to have a figure of merit independent of the load. When conjugate-matched conditions are achieved, then available power gain does equal actual power gain.

TRANSDUCER POWER GAIN

Yet another power gain is the transducer power gain, denoted by g_T. This is a ratio of the average power delivered to the load to the available power from the source. In terms of our previous notation,

$$g_T = p_L/p_{AS}.$$

It follows that

$$g_T = 4\ |E_2/E_S|^2\ R_S R_L/|Z_L|^2.$$

Thus, g_T involves the same voltage ratio as was encountered in our discussion of insertion power gain.

In terms of the 2-port open-circuit impedance parameters, it can be shown that

$$g_T = 4\ |Z_{21}|^2\ R_S R_L/|(Z_{11} + Z_S)(Z_{22} + Z_L) - Z_{12}Z_{21}|^2.$$

Interestingly enough, g_T depends upon both Z_S and Z_L.

Transducer power gain is a commonly used figure of merit for transducers because it relates the average power actually dissipated in a load to the maximum power capable of being delivered by the source.

EXCHANGEABLE POWER GAIN

Exchangeable power is closely related in concept to available power. In some applications the real part of the generator impedance is negative, as occurs with networks employing tunnel diodes. In this section the concept of available power is generalized to accommodate negative resistance sources.

Once again, consider the circuit of Figure 3. Assume the real part of the generator impedance is negative. To avoid the possibility of infinite average power flow, it is necessary that the real part of the load impedance, also, be negative. Note that negative values of R_L imply average power being returned by the load to the source. Assume the generator impedance is fixed and that both R_S and R_L are negative. The problem is to choose the load impedance so as to maximize the average power returned to the source. This maximum, denoted by P_E, is defined to be the exchangeable power of the generator.

It can be shown that the maximum power is returned to the source when, as shown before,

$$Z_L = R_S - jX_S.$$

The exchangeable power of the generator is then given by

$$p_E = |E_S|^2\ /\ 8R_S.$$

Observe that this is negative because of the negative sign associated with R_S.

The definition of exchangeable power gain is analogous to that for available power gain. With reference to Figure 2, let

p_{E0} = exchangeable power from the 2-port,
p_{ES} = exchangeable power from the generator,
g_E = exchangeable power gain of the loaded 2-port.

The exchangeable power gain is defined to be

$$g_E = p_{E0}/p_{ES}.$$

COMPARISON OF g_P, g_I, g_A, AND g_T

In general, there is no straightforward relationship between the expressions for the various power gains. Even so, certain observations can be made. From the definition of available power, we conclude

$$p_S \le p_{AS},\ \tilde{p}_L \le p_{AS},\ p_L \le p_{A0}.$$

It follows that

$$g_P \ge g_T,\ g_I \ge g_T,\ g_A \ge g_T.$$

We see that transducer gain acts as a lower bound for the other gains.

Various gains become equivalent under certain conjugate-matched conditions.
separate formulations. This discussion builds upon the previous article in this series which addressed the nonlinear effects of intermodulation, desensitization,

cross modulation, spurious responses, and gain compression/expansion. In the next article in this series, we will discuss in greater depth the various nonlinear modes and mechanisms that may arise in practical systems and components that incorporate nonlinear devices. A later article will present new findings of research and development to add nonlinear analysis and prediction capabilities to existing computational electromagnetics tools which determine detailed interference rejection requirements as part of an automated EMC assessment methodology for large, complex systems.

Table 1. Relations Between Power Gains Under Conjugate-Matched Conditions.

Conjugate-Matched Conditions	Equal Powers	Equal Power Gains
$Z_{in} = Z_S^*$	$p_S = p_{AS}$	$g_P = g_T$
$Z_L = Z_S^*$	$\tilde{p}_L = p_{AS}$	$g_I = g_T$
$Z_L = Z_{TH}^*$	$p_L = p_{AO}$	$g_A = g_T$
$Z_{in} = Z_S^*, Z_L = Z_S^*$	$p_S = p_{AS}, \tilde{p}_L = p_{AS}$	$g_P = g_I = g_T$
$Z_{in} = Z_S^*, Z_L = Z_{TH}^*$	$p_S = p_{AS}, p_L = p_{AO}$	$g_P = g_A = g_T$
$Z_S = Z_{TH}, Z_L = Z_S^* = Z_{TH}^*$	$\tilde{p}_L = p_{AS}, p_L = p_{AO}$	$g_I = g_A = g_T$
$Z_S = Z_{TH}, Z_L = Z_S^* = Z_{TH}^*, Z_{in} = Z_S^*$	$p_S = p_{AS} = \tilde{p}_L, p_L = p_{AO}$	$g_P = g_I = g_A = g_T$

Let

Z_{in} = loaded 2-port input impedance

Z_{TH} = Thevenin impedance seen looking into the 2-port output with the generator connected to the input.

Relations between power gains under conjugate-matched conditions are given in Table 1.

In general, equivalence between the power gains should not be assumed without checking to see whether the appropriate conjugate-matched conditions listed in the table are satisfied.

SUMMARY

In this article we have discussed the concepts of average power, available power and/or exchangeable power. These, in turn, are used in definitions of average power gain, insertion power gain, transducer power gain, and exchangeable power gain. Although the gains become identical in certain specialized situations, it was demonstrated that they are not, in general, equivalent. Consequently, the various powers and power gains cannot be used interchangeably which leads to a set of

REFERENCES

[1] D. D. Weiner, A. Drozd, Kurt V. Sunderland & Irina Popitich., "Analysis and Modeling of Weakly Nonlinear Systems", Newsletter Technical Features Article for the Applied Computational Electromagnetics Society, Vol. 15, No. 1, ISSN 1056-9170, pp. 9-15, March 2000.

Tutorial editor's note:
While many readers are very familiar with one modeling technique or another, they may not be familiar with all modeling techniques. The next few tutorial articles will provide a very basic introduction to each of the modeling techniques in turn. It is hoped that by minimizing the math, the reader will be able to achieve an intuitive understanding of the modeling technique, and then may be interested to pursue the topic further.

INTRODUCTION TO THE FINITE-DIFFERENCE TIME-DOMAIN TECHNIQUE

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Background

The Finite-Difference Time-Domain (FDTD) technique is a direct solution method for Maxwell's time-dependent curl equations. FDTD is straightforward to implement on a computer, and possibly more importantly, it is simple to understand and use, while being extremely powerful. It was originally introduced by Kane Yee in 1966 [1] but was not widely used until after 1975 when Talfove published the first validated FDTD models of sinusoidal electromagnetic wave penetration into a three-dimensional metal cavity [2]. Since 1986, FDTD has been extensively used for a wide range of applications involving electromagnetics.

The Basics of FDTD

As stated above, FDTD is based directly on Maxwell's curl equations (1).

$$\begin{aligned} \nabla \times H &= J + \frac{\partial D}{\partial t} \\ \nabla \times E &= -\frac{\partial B}{\partial t} \end{aligned} \qquad (1)$$

Although the FDTD equations are fully capable of handling materials with volume dielectric displacement currents (J in 1), the application of interest in this introduction is either free space or perfect electrical conductor (PEC). Therefore, since there is no dielectric material and this J term will be always zero in these materials, it is removed to make the explanation of FDTD clearer.

One-dimensional FDTD

The three dimensional Maxwell's equations can be converted to a one-dimensional case to introduce FDTD in (2).

$$\begin{aligned} \frac{dH}{dx} &= \frac{dD}{dt} \\ \frac{dE}{dx} &= \frac{dB}{dt} \end{aligned} \qquad (2)$$

Simply stated, the spatial change in magnetic field at a point in space is equal to the change in the electric flux density with time; and the spatial change in electric field at a point in space is equal to the change in the magnetic flux density with time (ignoring the Permittivity and permeability for a moment). Converting these differential equations into central-difference equations (3)(4), creates a simple set of linear equations. When the terms are rearranged,

$$\frac{H_i^n - H_{i-1}^n}{\Delta x} = \varepsilon \frac{E_i^{n+1} - E_i^n}{\Delta t} \qquad (3)$$

$$\frac{E_i^n - E_{i-1}^n}{\Delta x} = \mu \frac{H_i^{n+1} - H_i^n}{\Delta t} \qquad (4)$$

where: i denotes position in space and n denotes time. The equations now provide a solution to the 'new' value of the electric field, based upon the 'old' value of the electric field at that same point, and the difference of the magnetic fields on either side of it (5). The 'new' value of the magnetic field, based upon the 'old' value of the magnetic field at that same point, and the difference of the electric fields on either side of it (6).

$$E_i^{n+1} = E_i^n + \frac{\Delta t}{\Delta x \varepsilon}\left(H_i^n - H_{i-1}^n\right) \qquad (5)$$

$$H_i^{n+1} = H_i^n + \frac{\Delta t}{\Delta x \mu}\left(E_i^n - E_{i-1}^n\right) \qquad (6)$$

In order to maintain accuracy when converting from differential equations to central difference equations, the spatial position where the central difference is taken is the center between the alternate positions. Figure 1 shows the alternating positions of the electric and magnetic fields in space. The distances between adjacent electric or magnetic field points must be small, that is, the

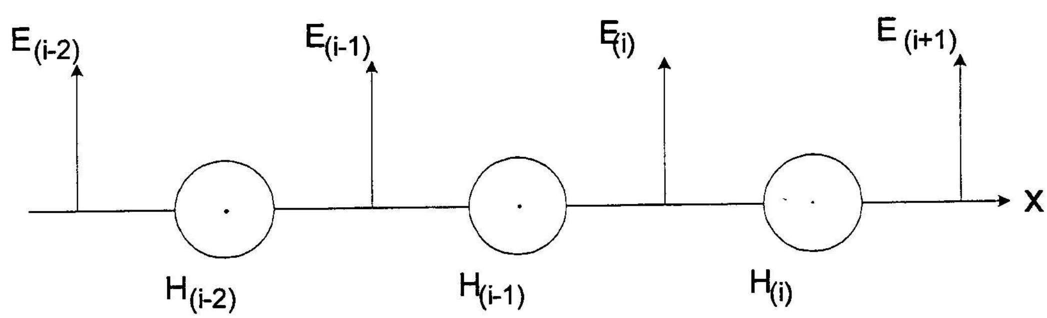

Figure 1 One Dimensional FDTD Grid

fields must not vary rapidly between adjacent points. A normal rule-of-thumb is to require the FDTD grid size to be no larger than 1/10th of the shortest wavelength. For cases where extreme accuracy is required, grid sizes of 1/20th the shortest wavelength are typically used. Since the electric and magnetic fields are not found at exactly the same point, the time when they are calculated must be slightly different as well. Typically, the time calculation for the electric fields is specified at +/- ½ time increment. So equations (5) and (6) become:

$$E_i^{n+1/2} = E_i^{n-1/2} + \frac{\Delta t}{\Delta x \varepsilon}\left(H_i^n - H_{i-1}^n\right) \qquad (7)$$

$$H_i^{n+1} = H_i^n + \frac{\Delta t}{\Delta x \mu}\left(E_i^{n+1/2} - E_{i-1}^{n+1/2}\right) \qquad (8)$$

Two-dimensional FDTD

When the above formulation is expanded into two dimensions, there will be three fields to solve, for example; either E_x, E_y, and H_z or H_x, H_y, and E_z. A typical two dimensional FDTD grid is shown for the TM case in Figure 2. (Although the TM case is shown here, the TE case can also be used.) Note the same basic conditions apply, that is, the 'new' value of the magnetic field (H_z, where 'z' is out of the page) is dependent on the 'old' magnetic field at that point, and the difference between the E_x components on either side of it, and the difference between the E_y components on either side of it.

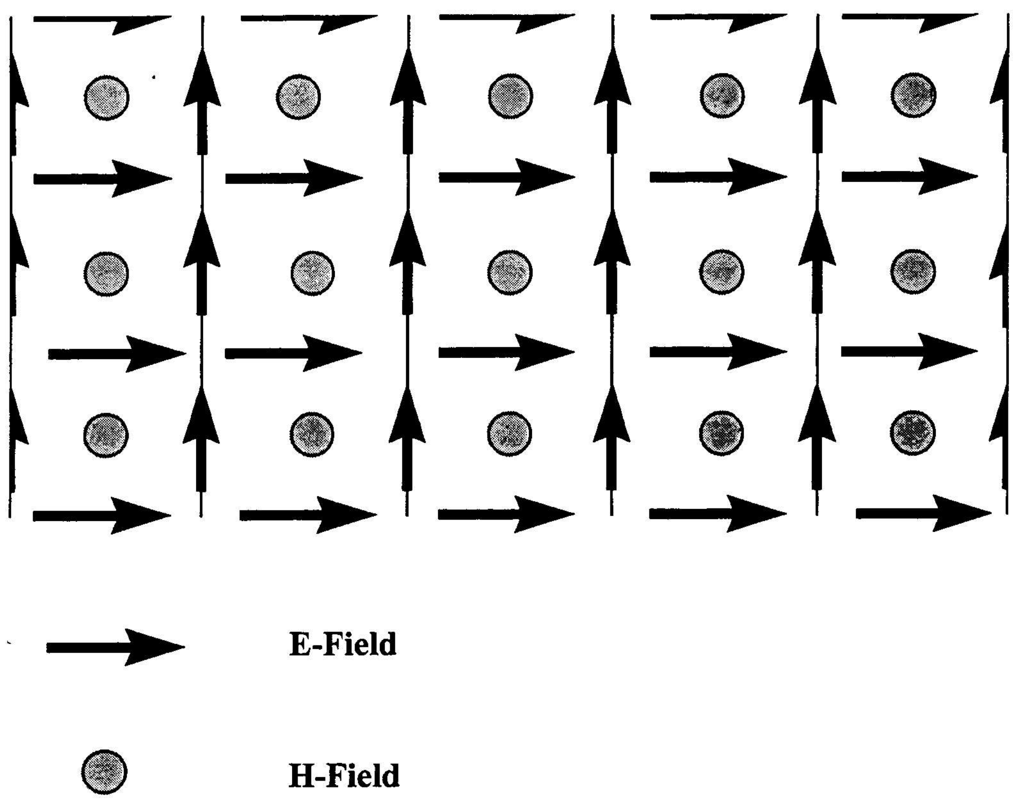

Figure 2 Two Dimensional FDTD Grid

As before, the 'new' value of the electric field is found from the 'old' value of electric field at that point, and the difference between the magnetic fields on either side of it. The FDTD two-dimensional equations are given in (9), (10) and (11) for this case, but the opposite case with E_z, H_x, and H_y can easily be found as well.

$$E_{x(i,j)}^{n+1/2} = E_{x(i,j)}^{n-1/2} + \frac{\Delta t}{\Delta x \varepsilon}\left(H_{z(i,j)}^n - H_{z(i,j-1)}^n\right) \qquad (9)$$

$$E^{n+1/2}_{y(i,j)} = E^{n-1/2}_{y(i,j)} + \frac{\Delta t}{\Delta y \varepsilon}\left(H^{n}_{z(i,j)} - H^{n}_{z(i-1,j)}\right) \quad (10)$$

$$H^{n+1}_{z(i,j)} = H^{n}_{z(i,j)} + \frac{\Delta t}{\Delta y \mu}\left(E^{n+1/2}_{x(i,j)} - E^{n+1/2}_{x(i,j-1)}\right) - \frac{\Delta t}{\Delta x \mu}\left(E^{n+1/2}_{y(i,j)} - E^{n+1/2}_{y(i-1,j)}\right) \quad (11)$$

Three Dimensional FDTD

The above FDTD expressions can be expanded into full three dimensions. Although difficult to draw many 3-D cells, a single cell is shown in Figure 3. The three electric fields are found along the axis, while the magnetic fields are found in the center of the face of the cube. Although the example in Figure 3 is a cube, the shape of the FDTD cell can be any rectangular shape. Typically, the aspect ratio of the 3-D FDTD cell is kept to 3:1 or less.

The basic operation of FDTD remains the same as earlier, that is, the 'new' electric field E_x is found from the 'old' electric field E_x, and the difference between adjacent H_z fields and the difference between adjacent H_y fields. Similar descriptions apply for the other five fields. Equations (12) through (17) show the simplified three dimensional FDTD equations.

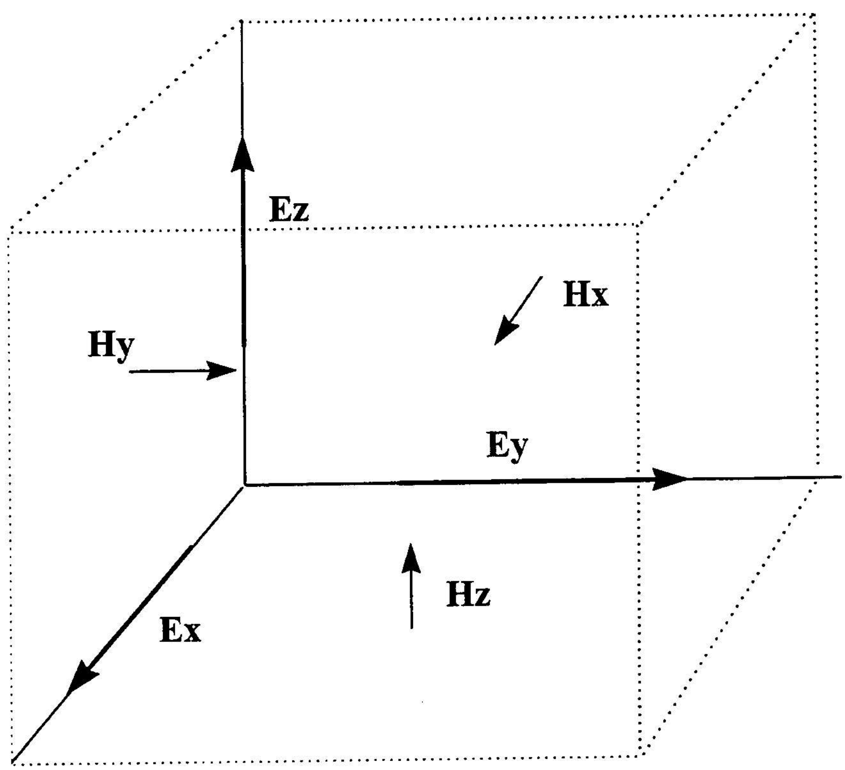

Figure 3 Three Dimensional FDTD Cell

$$H^{n+1}_{x(i,j,k)} = H^{n}_{x(i,j,k)} + \frac{\Delta t}{\Delta z \mu}\left(E^{n+1/2}_{y(i,j,k)} - E^{n+1/2}_{y(i,j,k-1)}\right) - \frac{\Delta t}{\Delta y \mu}\left(E^{n+1/2}_{z(i,j,k)} - E^{n+1/2}_{z(i,j-1,k)}\right) \quad (12)$$

$$H^{n+1}_{y(i,j,k)} = H^{n}_{y(i,j,k)} + \frac{\Delta t}{\Delta x \mu}\left(E^{n+1/2}_{z(i,j,k)} - E^{n+1/2}_{z(i-1,j,k)}\right) - \frac{\Delta t}{\Delta z \mu}\left(E^{n+1/2}_{x(i,j,k)} - E^{n+1/2}_{x(i,j,k-1)}\right) \quad (13)$$

$$H_{z(i,j,k)}^{n+1} = H_{z(i,j,k)}^{n} + \frac{\Delta t}{\Delta y \mu}\left(E_{x(i,j,k)}^{n+1/2} - E_{x(i,j-1,k)}^{n+1/2}\right) - \frac{\Delta t}{\Delta x \mu}\left(E_{y(i,j,k)}^{n+1/2} - E_{y(i-1,j,k)}^{n+1/2}\right) \quad (14)$$

$$E_{x(i,j,k)}^{n+1/2} = E_{x(i,j,k)}^{n-1/2} + \frac{\Delta t}{\Delta y \varepsilon}\left(H_{z(i,j,k)}^{n} - H_{z(i,j-1,k)}^{n}\right) - \frac{\Delta t}{\Delta z \varepsilon}\left(H_{y(i,j,k)}^{n} - H_{y(i,j,k-1)}^{n}\right) \quad (15)$$

$$E_{y(i,j,k)}^{n+1/2} = E_{y(i,j,k)}^{n-1/2} + \frac{\Delta t}{\Delta z \varepsilon}\left(H_{x(i,j,k)}^{n} - H_{x(i,j,k-1)}^{n}\right) - \frac{\Delta t}{\Delta x \varepsilon}\left(H_{z(i,j,k)}^{n} - H_{z(i-1,j,k)}^{n}\right) \quad (16)$$

$$E_{z(i,j,k)}^{n+1/2} = E_{z(i,j,k)}^{n-1/2} + \frac{\Delta t}{\Delta x \varepsilon}\left(H_{y(i,j,k)}^{n} - H_{y(i-1,j,k)}^{n}\right) - \frac{\Delta t}{\Delta y \varepsilon}\left(H_{x(i,j,k)}^{n} - H_{x(i,j-1,k)}^{n}\right) \quad (17)$$

Stability Conditions

The size of the FDTD cell must be set to be electrically small (small compared to the shortest wavelength) to satisfy the linearity assumption when converting from the derivative to central-difference equations. However, the size of the time step must also be small to ensure stability. The limit of the size of the time steps is given by the Courant Stability Condition [3] in equation (18).

$$\Delta t \le \frac{1}{c\sqrt{\frac{1}{\Delta x^2} + \frac{1}{\Delta y^2} + \frac{1}{\Delta z^2}}} \quad (18)$$

Since equation (18) uses *c* (the speed of light) it sets the maximum time step for free space. For applications where the speed of propagation is less than c (for example in a dielectric), then the slowest propagation velocity must be used instead of c in (18).

Absorbing Radiation Boundary Conditions

One important use of FDTD is in geometries with 'open' regions where the spatial domain of the computed field is unbounded in one or more coordinate directions. Clearly, no computer can store an unlimited amount of data, and so the computational domain must be limited in size. Although the computational domain must be large enough to enclose all the physical features of the model, it must be small enough to allow reasonable calculation times.

Since there is a limit to the size of the computational domain, there must be some boundary condition to ensure there is no non-physical (and unwanted) artificial reflection from the edge of the computational domain. As is seen in Figure 1 and equations (5) and (6), the 'new' electric field value at any point is found by using the 'old' electric field values, and the difference between the magnetic fields *on either side of that point*. However, at the edge of the computational domain there is no magnetic field outside the electric field on the boundary, and some other means to find this outer electric field is needed. It is not correct to simply set fields outside the computational domain to zero, since this would effectively place a perfect electrical or magnetic conductor along the boundary (depending upon which field is set to zero). Therefore, an absorbing boundary condition (ABC) is needed to simulate free space to approaching fields to ensure they do not reflect from the edge of the computational domain.

The simplest ABC would be to force the far-field relationship in (19) and find the outer electric fields

$$E = 120\pi \; H \quad (19)$$

from the magnetic fields and the wave impedance of free space. This is not practical, since the knowledge of the magnetic fields is not known at the point where the electric field is needed because both the E and H fields (and the times they are determined) are offset. Some other means to find the outer value of the electric field is needed.

Quite a number of different ABC's have been proposed in the literature. Some are better than others, that is, less error (apparent artificial reflection), and some take more processing resources than others. The most popular ABC's in use today are the Mur [4], Higdon [5], and Liao [6] ABC's. Comparisons between these and others can be found in the references [3][7], and are not be repeated here. All these ABC's operate correctly when the approaching fields are 'far-fields', that is, the relationship between E and H is as shown in equation (19). The amount of error (unwanted reflection) increases dramatically when the far-field condition is not maintained.[1] This results in so-called 'white space' around the model elements of interest to ensure the far-field condition is met. The white space is typically 1/6th lambda at the longest wavelength of interest (since this distance is commonly used as the distance where the far-field condition exists between the E and H fields). This results in much larger computational domain (to include the necessary white space) than is required for simply the model itself.

Creating Solid Models

Creating a solid material in the FDTD computational space is simply a matter of assigning ε, μ, and σ. It is very common to use perfect conducting conductors (PEC) in FDTD models, since the actual conductivity of the metal is often not of major concern in the simulation.[2]

FDTD Software

There are a number of commercial FDTD software packages (ranging in price from a few thousand dollars to many tens of thousands of dollars), as well as free FDTD code available from universities. The major difference is usually the computer platform the code runs on, and the ease of using the graphical user interface (GUI). As always, the potential FDTD user is encouraged to carefully weigh the advantages of one code against another for their particular application.

Summary

The FDTD technique is a fairly simple and straightforward technique, but very powerful nevertheless. FDTD solve Maxwell's curl equations by converting them to central difference equations, and solving the electric field and magnetic fields in leap-frog fashion.

The above description is intended to provide a general review of the FDTD technique. The reader is encouraged to use [3] and [7] for further reading of the FDTD technique.

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[1] All ABC's require the far-field relationship between the electric field and the magnetic field except the recently developed ABC's by Ramahi [8][9] and Berenger [10].

[2] For example, in the case of shielding of a metal enclosure or plate with apertures, the conductivity of the metal is not important, and PEC is sufficient.

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INDEX TO COMPUTER CODE REFERENCES FOR VOL. 14 (1999) OF THE *ACES JOURNAL* AND THE *ACES NEWSLETTER*

This computer code index is usually updated annually and published in the second issue of each volume of the *ACES Newsletter.*

LEGEND:

AJ	ACES Journal
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SI	Special Issue
*	Pre- or postprocessor for another computational electromagnetics code
**	Administrative reference only: no technical discussion (This designation and the index do not include bibliographic references)
Page No.	The first page of each paper in which the indicated code or technique is discussed

NOTE: the inclusion of any computer code in this index does not guarantee that the code is available to the general ACES membership. Where the authors do not give their code a specific name, the computational method used is cited in the index. The codes in this index may not all be general-purpose codes with extensive user-oriented features – some may only be suitable for specific applications. While every effort has been made to be as accurate and comprehensive as possible, it is perhaps inevitable that there will be errors and/or omissions. We apologize in advance for any inconvenience or embarrassment caused by these.

Allen W. Glisson and Ahmed A. Kishk, Editors-in-Chief, *ACES Journal*
20 May 2000

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VOLUME 14, 1999

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17 May 2000

A USER-FRIENDLY COMPUTER CODE FOR RADIATED EMISSION AND SUSCEPTIBILITY ANALYSIS OF PRINTED CIRCUIT BOARDS

J. Carlsson and P-S. Kildal

A user-friendly computer code, PCB-MoM, that is intended to be used in EMC applications for predicting radiated emission and susceptibility of printed circuit boards (PCB) is presented. The formulation is based on an electric field integral equation (EFIE) expressed in the frequency domain. The EFIE is solved by the method of moments using two-dimensional pulse basis functions and one-dimensional pulse test functions. In order to incorporate dielectric material in the substrate a spectral domain formulation is used. The code has been validated by comparison with previously published results and results obtained by other methods and codes. **[Vol. 14, No. 1 (1999), pp. 1-8]**

A SPARSE ITERATIVE METHOD (SIM) FOR METHOD OF MOMENTS CALCULATIONS

A. P. C. Fourie, D. C. Nitch, and A. R. Clark

The Sparse Iterative Method (SIM) provides a faster solution to Method of Moments (MoM) matrix equations than does LU-decomposition with forward and back substitution. The SIM produces a solution with computational time proportional to N^2, as opposed to the N^3 time dependence associated with LU-decomposition. The SIM is implemented in an object oriented MoM program which is functionally equivalent to NEC2. In three examples, the SIM is shown to produce results as accurate as LU-decomposition. For dipoles, a flat wire grid and a generic three dimensional missile shape, the speed increase ranged from 3-30 times the speed of LU-decomposition; greater speed increases can be expected with electrically larger problems. The SIM requires no problem formulation changes, such as segment renumbering, and despite the fact that it is demonstrated for a wire MoM based on NEC2, it is general enough to be incorporated to any MoM formulation. **[Vol. 14, No. 1 (1999), pp. 9-16]**

RADIATION FROM 3D SOURCES IN THE PRESENCE OF 2D COMPOSITE OBJECTS OF ARBITRARY CROSS-SECTIONAL SHAPE

A. A. Kishk, P. Slättman, and P-S. Kildal

A method of moments code for computing the radiation from arbitrarily oriented narrow slots or straight dipoles in the vicinity of a two dimensional (2D) multi-region composite cylinder of arbitrary cross-sectional shape and infinite extension in the *z*-direction is developed. The dipoles and the slots are represented by known equivalent electric and magnetic current distributions, respectively. The finite extent of the sources is included by Fourier transforming the equivalent electric and magnetic source currents along the infinite cylinder axis. The computed radiation patterns are compared to results computed by other methods and measurements. Radiation patterns are predicted both in the elevation and azimuthal planes. **[Vol. 14, No. 1 (1999), pp. 17-24]**

THE NUMERICAL METHOD OF CHARACTERISTICS FOR ELECTROMAGNETICS

J.H. Beggs, D.L. Marcum, and S-L. Chan

The objective of this study is to explore the benefits of using the theory of characteristics to develop accurate and efficient numerical algorithms for Computational Electromagnetics. The present work adapts the numerical Method of Characterisitics (MOC) from Computational Fluid Dynamics to the one-dimensional Maxwell curl equations in the time domain. The relevant theory of characteristics is developed and the inverse matching method is used to develop two numerical algorithms based on different interpolation schemes in the initial data surface. Stability and dispersion for these algorithms are discussed. Results are given for one-dimensional model problems involving free space pulse propagation, scattering from perfect conductors and reflection/transmission for lossy dielectric materials. The model problems are designed to provide quantitative insight to both accuracy and efficiency for different classes of realistic application problems. The Finite-Difference Time-Domain (FDTD) method is used as a convenient reference algorithm for comparison. It is demonstrated that these algorithms have accuracy comparable to FDTD, but do not require staggered grid storage, which simplifies impedance boundary conditions and implementation on nonuniform grids. The theory of characteristics demonstrates a very natural outer boundary condition without nonreflecting approximations or matched layers. A dispersion enhanced version of the MOC is also developed which has phase errors 50-5000 times lower than FDTD. This approach appears promising for development of dispersion enhanced characteristic based schemes for two and three dimensional applications. **[Vol. 14, No. 2 (1999), pp. 25-36]**

VOLUME LOADING — A NEW PRINCIPLE FOR SMALL ANTENNAS

D.B. Miron

Volume loading means placing conductively enclosed volumes at the ends of the arms of a short linear antenna to increase and control the capacitance of the structure. This technique allows tuning the antenna to make it series-resonant at a design frequency for which it is electrically small. This paper describes the discovery and development of the concept, gives some computed examples to illustrate the relation between shape and performance, and describes some experimental work, including a 900 MHz cordless phone demonstration design. **[Vol. 14, No. 2 (1999), pp. 37-44]**

SHAPED REFLECTOR ANTENNA ANALYSIS BY GRAPHICAL PROCESSING METHODS

J. M. Rius, M. Vall-llossera, C. Salazar, and A. Cardama

This paper presents the application of GRECO code to obtain the radiation pattern of single reflector antennas. Shaped reflectors with arbitrary geometries are modelled by CAD software. The GRECO graphical processing technique is used in order to extract the relevant geometrical information from the CAD geometry database. Surface reflection and edge diffraction are respectively analyzed by Physical Optics approximation and Equivalent Edge Currents Method. **[Vol. 14, No. 2 (1999), pp. 45-51]**

VALIDATION AND DEMONSTRATION OF FREQUENCY APPROXIMATION METHODS FOR MODELING DISPERSIVE MEDIA IN FDTD

J. H. Beggs

Recently, digital signal processing techniques were used to design, analyze and implement discrete models of polarization dispersion for the Finite-Difference Time-Domain (FDTD) method. The goals of the present work are to illustrate the FDTD update equations for these techniques and to validate and demonstrate these techniques for one-dimensional problems involving reflections from dispersive dielectric half-spaces. Numerical results are compared with several other dispersive media FDTD implementations. **[Vol. 14, No. 2 (1999), pp. 52-58]**

SCATTERING FROM A PERFECT ELECTRIC CONDUCTOR CYLINDER WITH AN INHOMOGENEOUS COATING THICKNESS OF GYROELECTRIC CHIRAL MEDIUM: EXTENDED MODE MATCHING METHOD

D. Cheng and Y. M. M. Antar

Based on the eigenfunction expansion of electromagnetic waves in the gyroelectric chiral medium, an extended mode-matching method is developed to study the electromagnetic scattering of a perfect electric conductor (PEC) circular cylinder with an inhomogeneous coating thickness of gyroelectric chiral medium. Excellent convergence property of the bistatic echo width is numerically verified, which establishes the reliability and applicability of the present extended mode-matching method for the two-dimensional problem of gyroelectric chrial medium. **[Vol. 14, No. 2 (1999), pp. 59-66]**

IMPROVED MODELING OF SHARP ZONES IN RESONANT PROBLEMS WITH THE TLM METHOD

J. A. Morente, J. A. Portí, H. Magán, and O. Torres

Several new modified nodes based on the symmetrical condensed node with stubs are proposed to solve the difficulty of the TLM method in modeling problems highly dependent on frequency. These nodes avoid the problem of indirect modeling at critical points, such as bends and corners, thus providing more accurate results and flexibility in the modeling of conducting parts. The new nodes are applied to specific problems of rectangular waveguides loaded with rectangular irises of finite width to verify their capability to predict resonant phenomena. **[Vol. 14, No. 2 (1999), pp. 67-71]**

PREDICTING MOM ERROR CURRENTS BY INVERSE APPLICATION OF RESIDUAL E-FIELDS

A. P. C. Fourie, D. C. Nitch, and A. R. Clark

This paper presents a methodology to predict *a posteriori* the error associated with a Method-of-Moments solution. The discussion is limited to a one dimensional pulse basis function wire-based implementation, but is easily extended. A Formulation for Error Prediction based on the relationship between the error in the boundary conditions and the error in the solution is presented, and validated by an over-segmented problem. The formulation is then used in a normally-segmented solution to predict the error by means of a linear interpolation of the calculated current which results in a smoother boundary condition error. The results show that this normally-segmented methodology predicts the error current within 5% of the "accurate" error current obtained by a 20:1 oversegmentation of the problem. Further work needs to be performed to extend this to the multidimensional case, although no technical difficulties are expected with this. **[Vol. 14, No. 3 (1999), pp. 72-75]**

ON THE ITERATION OF SURFACE CURRENTS AND THE MAGNETIC FIELD INTEGRAL EQUATION

D. D. Reuster, G. A. Thiele, and P. W. Eloe

In an effort to mathematically validate the convergence properties of various surface current-iterative methods, the magnetic field integral equation is analyzed for its contraction mapping properties. The analysis is performed first on the general integral operators and then on the matrices representing the discrete forms of the integral operators associated with the different iterative methods. The contraction mapping properties are determined by investigating the spectral radius of each linear operator. Conditions for the verification and validation of these iterative methods are provided, along with mathematical checks for the existence of spurious modes and the existence of internal resonance. **[Vol. 14, No. 3 (1999), pp. 76-83]**

MODELING OF LOW-GAIN ANTENNAS ON AIRCRAFT USING APATCH

J. Calusdian and D. Jenn

Radiation patterns for low-gain antennas such as those used for telemetry and collision avoidance systems were computed using the APATCH program. APATCH uses shooting and bouncing rays (SBR) to compute the radiation pattern for the antenna installed on a scattering geometry. A built-in antenna model was used to represent a telemetry antenna and its accuracy verified by comparison with measurements and results from a method of moments patch code. Antenna patterns were computed for various locations on a Cessna 172 and an F-18 Hornet. **[Vol. 14, No. 3 (1999), pp. 84-90]**

A NOVEL SPATIAL IMAGES TECHNIQUE FOR THE ANALYSIS OF CAVITY BACKED ANTENNAS

A. A. Melcón and J. R. Mosig

This paper describes a new contribution to the analysis of arbitrary shielded circuits and antennas of complex shapes, in the frame of the integral equation (IE) and Method of Moments formulation (MoM). The technique is based on the spatial image approach and a new specially truncated set of images is developed to enhance the convergence behavior of the series involved. Results show that, with the new specially truncated series of images, convergence is achieved very fast. In this paper simulated results obtained with the new approach are compared with measurements. **[Vol. 14, No. 3 (1999), pp. 91-99]**

EFFICIENT SOLUTION OF LINEAR SYSTEMS IN MICROWAVE NUMERICAL METHODS

L. Tarricone, F. Malucelli, and A. Esposito

A common bottle-neck, limiting the performance of many electromagnetic numerical methods, is the solution of sparse linear systems. Until now, this task has been typically solved by using iterative sparse solvers, whose require heavy computational efforts, especially when the problem is not well-conditioned.

An alternative strategy is based on the use of banded solvers, which numerical complexity is quadratical with respect to the matrix bandwidth. Of course, these methods are efficient provided that the matrix bandwidth is sufficiently small. In this paper, a method (called WBRA) for the bandwidth reduction of a sparse matrix is presented: it is here specifically customized to typical electromagnetic matrices. The approach is superior to all the previous algorithms, also with respect to commercial well-known packages, and is suitable also for non-symmetric problems.

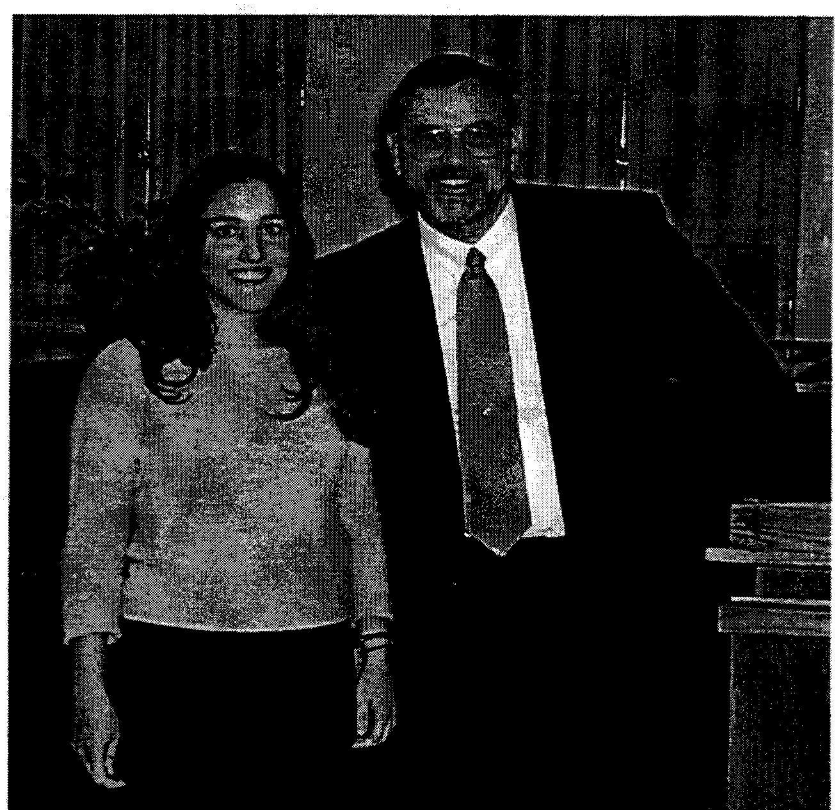
As demonstrated by results, the use of WBRA, in conjunction with common banded solvers, substantially improves (up to one order of magnitude) the solution times in several electromagnetic approaches, such as Mode-matching, FEM, and MoM analysis of microwave circuits. In conclusion, it is proved that the high efficiency and effectiveness of WBRA turns the strategy of bandwidth reduction combined with a banded solver into the most profitable way of solving linear systems in electromagnetic numerical methods. **[Vol. 14, No. 3 (1999), pp. 100-107]**

PARALLEL IMPLEMENTATION OF GALERKIN TECHNIQUE IN LARGE SCALE ELECTROMAGNETIC PROBLEMS

D. I. Kaklamani, K. S. Nikita, and A. Marsh

An integral equation formulation in conjunction with a palallelised Galerkin technique is employed to solve large-scale electromagnetic (EM) problems. The proposed technique is applicable to EM structures consisting of similar conducting or dielectric parts, defined as "elements." Coupled integral equations are derived in the frequency domain, written in terms of the conductivity currents or the electric fields developed on the conducting or dielectric "elements" surfaces, respectively. The system of integral equations is numerically solved via the parallel computed Galerkin technique, with convenient entire domain basis functions. Even for electrically large structures, the use of entire domain basis functions leads to relatively small order linear systems and the main computational cost refers to the matrix fill rather than the matrix solution. The parallelisation introduced to the computation of the matrix elements overcomes the limitation of using Method of Moments at lower and resonant frequencies. The inherent parallelism of the introduced technique allows for the results to be obtained with minimal addition to sequential code programming effort. Two indicative electromagnetic compatibility applications are presented, concerning the coupling of incident waves with multiple conducting rectangular plates and the coupling phenomena occurring in a multi-element waveguide array looking into a layered lossy cylinder. Numerical results are presented, while the applicability/suitability of diverse High Performance Computing platforms is judged, based on both performance obtained and ease of code portation. **[Vol. 14, No. 3 (1999), pp. 108-116]**

President Perry Wheless presenting award to Daniel Reuster



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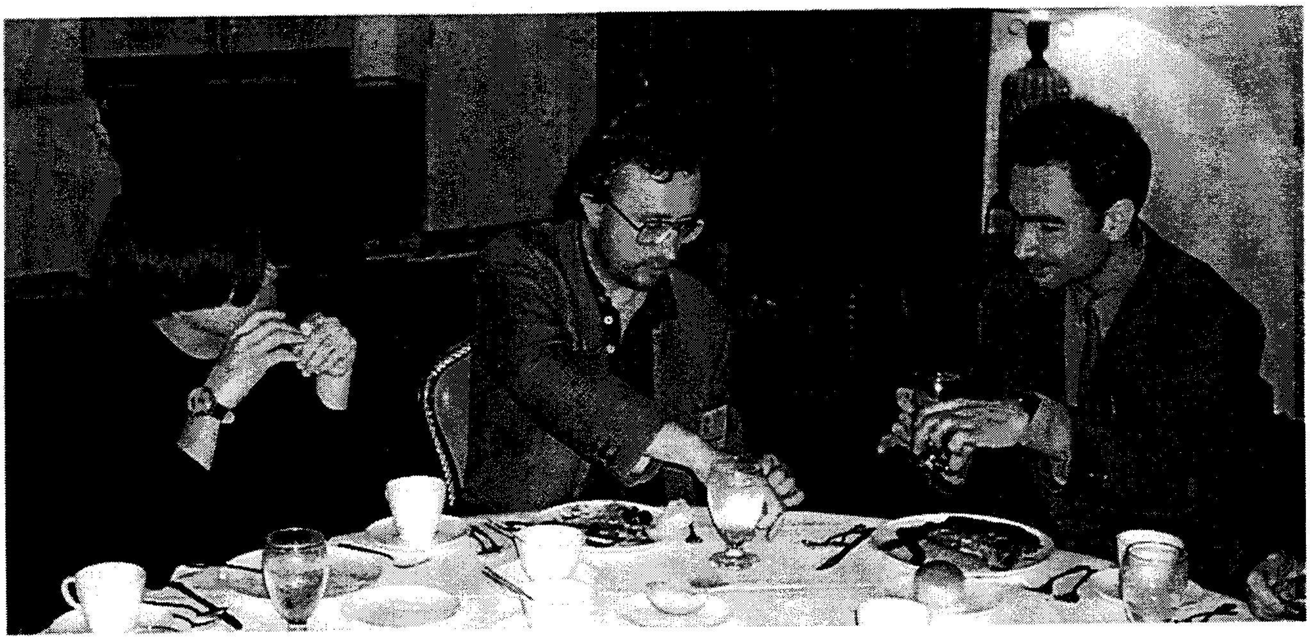

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