APPLIED COMPUTATIONAL ELECTROMAGNETICS NEWSLETTER

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INTRODUCTION TO THE ACE NEWSLETTER

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Hello, and welcome to the first issue of the Appliied Computational Electromagnetics (ACE) Newsletter. If you've been following the somewhat lengthy path taken so far to get to this point, you know that we're a little late in getting to press. There are a variety of reasons for this, some of which will become more apparent below. For the moment, let's examine what we're trying to accomplish and why, first by reviewing some introductory material originally published in a four-page mailer describing ACE and a June 1985 IEEE AP-S Newsletter announcement.

BACKGROUND

Over the past 10 years or so, computer modeling in electromagnetics has become firmly established. Its use has become so commonplace that presentations at technical meetings and articles in technical journals, when they mention numerical issues at all, deal primarily with new developments and omit the applications' details. As a results, those areas where modeling is used on a regular basis may give the impression that the computational aspects are well understood and require no special attention. While there undoubtedly are cases where this is true, modeling still remains more art than science. It is in the area of modeling "art" where information exchange among its practicioners, both developers and users, is most needed.

Because of the perceived need to encourage and promote discussion within the applied computational electromagnetics community, a preliminary meeting having the title "1st Annual Review of Numerical Electromagnetics Code (NEC) Applications" was held at Lawrence Livermore National Laboratory March 19-21 1985. Response to and discussions at the NEC Review showed that there was a need for an applications forum, with the consensus being that both a regular meeting and a publication were appropriate. The Applied Computational Electromagnetics (ACE) Newsletter is the initial phase of a publication that we expect to become a journal at some future time. This is most likely to occur in conjunction with the formation of the ACE Society, but other mechanisms by which the Newsletter might make the transition to a journal can be visualized.

NEWSLETTER GOALS

The overall goal of the Newsletter is to foster information exchange among computer modelers in electromagnetics. In order to accomplish this, we will focus on computer-applications issues as opposed to the research and development emphasis typical of other media. In order to promote this goal and focus, the ACE Newsletter will be organized along the following lines:

1) There will be several regular departments or features to appear in the Newsletter including--

--A modeling column that will contain announcements of general interest such as problems and limitations discovered in codes and fixes developed thereto, enhancements and improvements added, innovative applications discovered, and new codes available.

--Applications' notes (see further criteria below), to consist nominally of an introductory summary which describes in 1-2 pages the problem, the approach used, the results obtained, and conclusions and recommendations, followed by a maximum of 8-10 finished Newsletter pages which provide more detail, including results and validation thereof. Shorter notes than this nominal length are welcome of course.

--Code descriptions which briefly describe new codes that have become available.

--Tutorial articles that give an introduction/ overview of various modeling topics from an applications' viewpoint, covering material likely to help the modeler do more accurate, efficient and reliable modeling.

--An ACES column to provide news of the ACE Society activites, assuming ACES eventually becomes a reality.

2) Material puslished in the Newsletter will be so organized that as subsequent Newsletters are purlished, the various sections can be removed to be kept in a common binder, thus allowing various subjects (applications' notes sorted according to antenna type, for example) to be accumulated as a separate reference document. This topic-oriented organization of the Newsletter material should provide a more useful way to collect information than the chronological format intrinsic for most journals.

3) For the ease and convenience of both readers and contributers to the Newsletter, standard formats will be adopted wherever possible for providing material for inclusion. By requiring an application-note summary to contain specific information in a specified order, the reader can always be sure of what the note's minimum content will be, and the authors will not have to design their own format.

APPLICATIONS' NOTES CRITERIA

It is our intention to promote this focus by seeking contributions that are primarily code oriented in that they describe work that satisfies at least one of the following criteria:

1) Code validation. This may be done using experimental, analytical or computational separately or in combination.

2) Code applications to design. This could include studies of parameter-sensitivity tradeoffs for the purpose of tabulating handbook-type data or optimization studies for achieving some desired performance.

3) Code studies of basic physics. This might involve using a code to simulate reality in such a way that better or new physical insight or understanding is achieved.

4) Code enhancements and fixes. This category is self explanatory, but could cover significant changes to existing codes, including extending their applicability or improving their performance, and correcting problems or removing limitations.

The specifics of format, content and style for applications notes are as follows:

FORMAT During the start-up phase of the ACES Newsletter, and for the forseeable future thereafter, we plan to employ camera-ready final copy because of the cost savings and convenience it provides. Please follow the guidelines used by the IEEE Antenna and Propagation Society Newsletter which are paraphrased below--

". . . Materials intended for publication in the Newsletter should be typed single-space in photo-ready condition--all typing must be kept within a 4 1/2 inch column width. Any plain white paper can be used but special paper with light blue column guidelines such as can be obtained from the IEEE Headquarters for its Newsletters is convenient. An alternative method is to type on plain bond paper using heavy black guidelines on a back-up sheet. . ." The length is flexible, but would nominally be a maximum of 10 Newsletter pages, which translates to about 22 pages of single-column typed copy, including figures.

CONTENT In order to ensure an appropriate level of quality control and review, contributions for the Applications' Notes will be refereed. We intend to have the Asssociate Editors perform this function whenever possible in order to reduce time delays in publication. Contributions will be reviewed both for technical correctness and for adherence to the guidelines listed above with respect to information content. Contributors might therefore want to submit the initial manuscript in draft form so that any suggested changes can be made before the photo-ready copy is prepared.

STYLE Although it is our intention to be reasonably flexible with respect to the style of the applications' notes which appear in the ACE Newsletter, it is advisable to provide some guidelines. We suggest that article preparation with respect to equations, referencing and layout follow that of Radio Science. In particular, equation numbers are put in parentheses at the right column margin, and references are given by author(s) name and year in the body of the article and listed in alphabetical order at the end.

GENERAL COMMENTS

The guidelines and goals given above seemed reasonable to those of us who were involved in initial discussions concerning the Newsletter. But they may represent more the ideal than the practical, at least for the first few issues, and a good deal of flexibility is necessary. For example, the suggested note format is unlikely to be uniformly adhered to by individual contributors, thus requiring substantial revision by the author(s) or re-typing by the ACE staff (which is painfully small right now). Since we have not so far been inundated with contributions, it's not realistic to insist that those we do receive follow an untested format, which seems likely to be revised in time in any case.

Concerning the specific issue of contributions, we welcome any which meet one of the criteria listed above. Note content is of more concernthan format, especially if the ACE Newsletter is to meet a need not now being filled by the other literature. Those of you who attended the 1st NEC Review may remember that we hoped to issue an expanded proceedings of that meeting by publishing written versions of most of the presentations given there as the first issue of the Newsletter. In retrospect, reconstructing those written versions from copies of vu-graphs and transcriptions of the talks, as we originally intended, turns out not to be very easy to implement, and it was probably a mistake not to require note-length written versions from the presenters at the Review. Therefore, the only written proceedings of that 1st Review will remain the abstracts handed out to attendees of the Review, and sent later to Newsletter subscribers. However, we do expect that quite a few of the presentations made then will appear eventually in the first first editions of the Newsletter.

This seems an appropriate place to mention that a follow-on meeting to the 1st Review is being planned. It will be held at the Naval Postgraduate School in Monterey, CA, during the week of March 17, 1986. Information concerning this meeting can be obtained by calling or writing: Dr. R. W. Adler, Naval Postgraduate School, Code 62AB, Monterey, CA 93940, (408) 646-2352; or Dr. M. J. Barth, L-153, Lawrence Livermore National Laboratory, PO Box 5504, Livermore, CA, 94550, (415) 423-1291; mentioning the meeting title "2nd Annual Review of Progress in Applied Compuational Electromagnetics". Abstract deadline will be January 30, 1986, and written notes of up to 10 pages in finished length (approximately 22 pages of 4 1/2 inch columns including figures) will be required from presenters at the time of the meeting, prepared following the guidelines given above. These notes will be published in regular issues of the Newsletter following the Reveiw. We expect to have more definitive information concerning ACES at this review. Assuming that ACES does get underway during the next year, anyone attending the 2nd Review or subscribing to the Newsletter will automatically become an ACES member.

In conclusion, I want to thank all of you who encouraged our efforts in getting the Newsletter underway. There are quite a number of subscribers beyond those who attended the 1st Review, whose subscriptions were submitted in response to the various announcements that were made concerning the ACE Newsletter or ACES. We hope that your expectations about the Newsletter and whatever might follow will be met at least in part. As is the case of any publication activity, there can be no "product" without the active partcipation of the membership/readership. Any contributions that you wish to submit will be most welcome. We have decided to "go to press" with this first issue in spite of not having the amount of material that we would have preferred, or we might never have gotten started. In the future, we would hope to include 10 or so applications articles per issue, plus the other features mentioned earlier. Contributions of any nature should be sent to Dr. Marvin Barth, L-156, Lawrence Livermore National Laboratory, PO Box 5504, Livermore, CA 94550. Until a legally chartered organization is set up, please continue to make any subscription checks payable to "Regents, University of California". If on the other hand, you are unhappy with this first issue of the Newsletter for any reason, feel free to request a refund by contacting Dr. Barth. All of those associated with this project are convinced of its value and need, but also recognize that this first issue of the Newsletter might not be what we would hope eventually to produce in terms of the number of articles, the various features intended for eventual inclusion, etc.

As for myself, I expect to continue writing a column for the Newsletter, and to serve in an editorial capacity of some sort. But by the time you receive this, I will have changed employers, to become a Professor of Electrical and Computer Engineering at the University of Kansas, Lawrence, KA 66045-2228, telephone (913) 864-4620.

EM MODELING NOTES

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One of the functions seen for the ACES Newsletter is the dissemination of news about modeling codes, including problems and limitations discovered and fixes, when known, as well as code enhancements and new codes available. Some of this information will be contained in Applications' Notes and separate code descriptions. This column, however, is seen as an initial dissemination point for such information that comes to the attention of the ACES staff. We would like to cover any commonly used EM modeling codes, but topics concerning NEC-Method of Moments will probably predominate due to the large amount of feedback that we receive as the originators of this code.

For this first issue, it seems appropriate to summarize some known errors, limitations, and potential traps in using NEC-Method of Moments. We hope that in future newsletters we will be able to report fixes to the problems inherent in the code. While this summary is not complete, it will cover some of the most frequently encountered problems.

There is a variety of reasons for which a code may not produce a satisfactory result. These can be categorized as follows:

1. Computer system errors (compiler bugs).
2. Code errors.
3. Numerical accuracy limitations due to finite precision and range for numbers.
4. Numerical modeling errors which arise from obtaining only an approximate solution to the mathematical model that is used.
5. Physical modeling errors which arise from replacing the physical problem of interest with an approximate mathematical model.
6. Model description errors (input errors).
7. Incorrect interpretation of results by the user.

With a complex code such as NEC, each of these possibilities should be considered when the results are suspect. Items falling into Categories 1 through 4 will be considered here. Categories 5 through 7 are more under the control of the user, although ways of detecting or assessing such problems may be discussed in a future column.

In the first category, compiler bugs should be rare but recently have been encountered with disturbing frequency on VAX systems, apparently due to over zealous optimization in the V4.X compilers. In each case where the error has been traced, it occurred in a statement involving complex variables. In some cases, the program crashed, while in others, it ran with wrong results. The V4.1 compiler compiled NEC3 incorrectly (subroutines SBF and TBF), producing wrong results. V4.3 ran amok on subroutine GFLD in NECGS (NEC3 specialized for a monopole on a radial-wire ground screen), causing the code to crash when the ground wave calculation was attempted. It also produced wrong results for a version of NEC-3 modified to model insulated wires. A problem was also encountered with the NEC Basic Scattering Code, developed at OSU, with wrong results for singly reflected rays. In each case, the problems went away when the code was compiled with the /NOOPTIMIZE option.

We have received a new compiler from DEC, Version V4.3-173, which seems to compile NEC correctly. Some of our VAXes still have the V4.3-145 compiler which has problems, however. If in doubt, try /NOOPTIMIZE.

Problems have also been reported on the CDC CYBER systems. In this case, NEC3 must be compiled with the Fortran Version 4 rather than Version 5 and with optimization level zero. We do not have access to a CYBER system at LLNL so cannot determine whether the problem is with the compiler or part of the code that is incompatible with Version 5.

We would like to say that code errors of Category 2 do not occur, but unfortunately, some do slip by. In September, 1985, we set up new versions of NEC2 and NEC3 in single and double precision for VAX computers using the CDC 7600 version at LLNL as a starting point. Some errors, mostly minor, were introduced at this time. The most serious was a misspelling of IND1 as IIND1 in common block /DATAJ/ in subroutine PCINT in the double precision NEC2 (NEC2DALL.FOR). This would effect results for a wire connected to a surface patch. This and other errors have been corrected, but some codes containing them may remain in circulation. We will be sending a list of corrections to recipients of these codes.

Also, there appears to be an uninitialized variable in SOMNTX, the code that generates Sommerfeld integral tables for NEC3. This code runs properly if the memory is initially set to zero, which is standard on many computers. It requires a special command, SET(0), when loading on CDC CYBER systems, however.

Category 3 errors, numerical accuracy and range limitations, have caused problems when running NEC on computers such as IBM and VAX. The code was developed on CDC computers with fourteen place accuracy and an exponent range of -293 to 322, so some problems remained hidden until it was run on other computers.

Exponent range problems have been encountered during evaluation of complex exponential functions in subroutine GASY1 in NEC3 and subroutine SAOB in SOMNTX. The problem is actually an underflow but causes the code to halt with an overflow message. This problem can be corrected by writing a new complex exponential function, for example, in double precision

```
      COMPLEX*16 FUNCTION ZEXP(Z)
      COMPLEX*16 Z
      IF(REAL(Z).GT.-85.)GO TO 1
      ZEXP=(0.,0.)
      RETURN
    1 ZEXP=EXP(Z)
      RETURN
      END
```

Function ZEXP can be substituted for CEXP or the generic form EXP where necessary. Remember to declare ZEXP as COMPLEX*16 in any routines where it is used. We plan to make this change in the codes that we release in the future.

Numerical accuracy problems occur most often when NEC is used to model electrically small antennas. Some limits for accurate results on a VAX computer are:

	Single Precision	Double Precision
Dipole of length ℓ with 5 segments	$\ell/\lambda > 10^{-3}$	10^{-7}
Loop with circumference C with 9 segments	$C/\lambda > .05$	$3(10^{-4})$
$\lambda/2$ dipole with N segments	$N < 60$	?

We hope to be able to improve this accuracy situation during the next year.

Category 4 errors differ from Category 3 in that they would remain if the code was run on an ideal computer with infinite precision. Numerical modeling errors are unavoidable in the numerical solution of the integral equation but, hopefully, can be kept small. Use of segments or patches that are too large relative to the wavelength is a common source of error. NEC is also subject to errors in modeling wires with an abrupt change in radius or in using the "charge discontinuity" source model on thick wires.

Some recent problems involving the ground include modeling a small loop antenna over groun small monopole antenna on a ground screen abov ground. The small-loop problem results from the resistance, which may be several orders of magn smaller than the reactance, being lost in the li accuracy (10^{-3} to 10^{-4} relative error) in evaluation of field due to the interface. The pr with a ground screen above ground appears to r from the condition applied to charge density a junction of the monopole and radials. Reaso results can be obtained if the segment lengths a junction are on the order of the height of the s above ground. This is not necessary, and perhap desirable, for a buried screen. More validating is needed on many such ground problems to dete how well or poorly the code is doing.

I hope that this list does not leave impression that NEC is all errors. It handles problems quite well. The purpose here was to r some problems that have been encountered in past. Hopefully, in future columns or new releas the code we will have fixes for these problems. in the NEEDS project, we plan to implement optio the code to check for errors in each of the categories listed above.

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WIRE-GRID MODELING OF SLOT ANTENNAS

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May 16, 1984

Synopsis: The ability of a computer code to model slot antennas is investigated. A comparison of experimental and numerically calculated antenna data is presented.

ABSTRACT

The wire-grid modeling option of the Numerical Electromagnetics Code has been used to simulate the behavior of rectangular slot antennas in finite ground planes. The sensitivity of calculated antenna impedance to changes in model wire-segment length and radius has been investigated. It is shown that very thin slots can be effectively modeled using the wire-grid approach. A technique for including the effects of a thin dielectric slot backing in the model is presented. A comparison of calculated and experimentally measured data shows that the wire-grid model gives precise radiation pattern predictions and calculates impedance values with sufficient accuracy for use in preliminary antenna design.

I. INTRODUCTION

During the past decade, a great deal of research has been directed toward developing numerical techniques which can be applied to complex electromagnetic boundary value problems [1,2,3]*. Many problems which previously could not be handled using classical electromagnetic theory, have been successfully solved using these new techniques in conjunction with digital computers. Several general purpose computer codes have evolved to aid users in applying these techniques to their particular problems.

One of the more comprehensive software packages is the Numerical Electromagnetic Code (NEC) developed at the Lawrence Livermore Laboratory in Livermore, California [4]. Both the Electric Field Integral Equation (EFIE) and Magnetic Field Integral Equation (MFIE) are used to model the electromagnetic response of structures. Using a technique known as the Method of Moments, the integral equations can be solved numerically to determine the currents induced on metallic structures by an exciting source or an incident electromagnetic wave. The EFIE works well in modeling thin-wire structures and conducting sheets having small or vanishing volumes. The MFIE is applicable to closed voluminous structures which have smooth surfaces. The NEC program allows hybrid structures to be modeled using both the EFIE and MFIE simultaneously.

The models which result from the application of numerical approximations to the EFIE are known as thin-wire or wire-grid models. This is due to the fact that solid conducting surfaces, when modeled, are replaced by a wire mesh or grid. Of course, as the wire mesh become more dense, the solid surface is more closely approximated. Wire-grid modeling has been used successfully for the analysis of very complex structures such as wire and loop antennas mounted on ships and aircraft [2,5].

A survey of the literature has shown a lack of information regarding the application of wire-grid modeling to slot type antennas. It is the purpose of this paper to investigate and evaluate the potential of wire-grid modeling, via the Numerical Electromagnetic Code, for analyzing slot antennas. A point-fed rectangular slot antenna mounted in a finite ground plane will be modeled using the wire-grid approach. Measured and calculated values for the antenna's terminal impedance and radiation patterns will be used to study the sensitivity of the model to changes in parameter such as wire radius and mesh density. In addition, a technique will be presented whereby a dielectric sheet covering the slot can be included in the wire-grid model.

II. GENERAL FORMULATION

A simple rectangular slot cut into a square conducting sheet was chosen as the antenna structure to evaluate the wire-grid modeling approach. This particular antenna was selected since it is representative of general slot antennas and its geometry simplifies the modeling procedure. Also, there is some qualitative information available regarding the characteristics of rectangular slot antennas and this was used as a starting point in evaluating the modeling technique [6,7].

As shown in Fig. 1, the slot antenna is excited in a point-fed fashion by the voltage source across the center of the slot. The length of the slot is given by L while the slot width is denoted by W. The ground plane, which contains the slot, is square and has sides of length S. A spherical co-ordinate system is also shown in Fig. 1 to define the ϕ- and θ-polarized components of an electric field vector with respect to the slot antenna. These components will be used later in describing the fields radiated by the antenna.

Two characteristics are important when describing antenna behavior. These are the antenna's terminal impedance and its radiation patterns. For the slot antenna, the terminal impedance is the impedance seen by the voltage source used to excite the antenna This impedance is important with regard to matching the antenna to a transmission line and a source or receiver to assure efficiency in the transfer of power. The radiation patterns of an antenna define its directionality and gain characteristics. These patterns define the spatial distribution of the ϕ- and θ-polarized electric field components radiated in the

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*Numbers in brackets designate references listed at the end of this report.

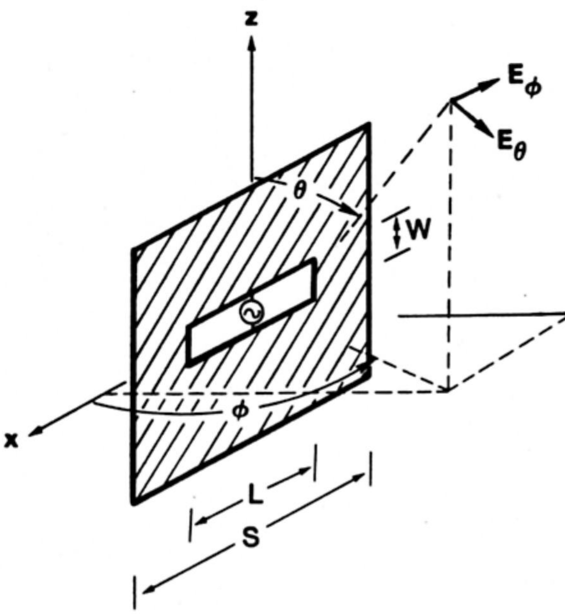


Figure 1. Slot Antenna Geometry.

far-zone of the antenna. The ability of wire-grid modeling to predict both the termination impedance and radiation patterns for the rectangular slot antenna will be evaluated.

One additional physical characteristic of slot antennas must be considered for practical modeling. It is usually desirable to place some type of dielectric backing in the slot region both for structural support and to seal against the environment. Thus, it will be necessary to include a way for the wire-grid model to account for the presence of a slot dielectric.

III. WIRE-GRID MODELING

Figure 2 shows the basic wire-grid model for a rectangular slot antenna located in a square finite ground plane. All of the dimensions on the model are given in terms of L, the slot length. Each side of the ground plane is S = 2.667 L with the slot width, W = 0.114 L. Unless otherwise stated, the radius the wire segments making up the model will be ass to be R = 0.017 L.

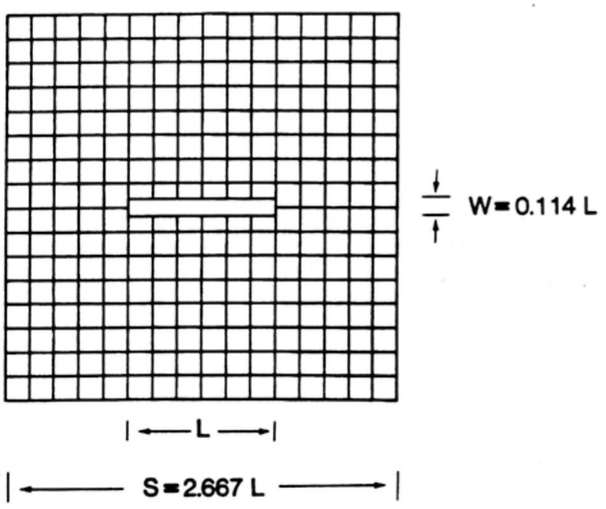


Figure 2. Basic Wire-Grid Model for the Slot Antenna.

The above model, with its particular wire density and wire segment radius, will be the basis studying the sensitivity of the wire-grid model parameter changes. The reason for choosing t particular values for mesh density and wire ra will now be discussed.

A. Wire Mesh Density

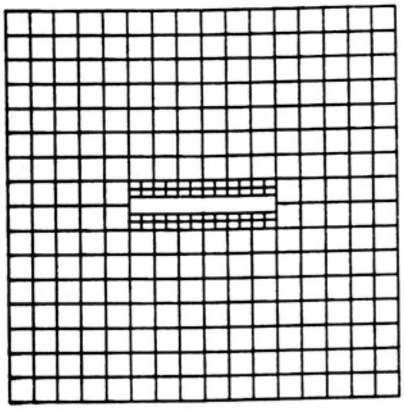
The term mesh density, as used in t report, actually deals with the length of the segments which form the grid for the mode surface. Thus, decreasing the length of wire segme has the effect of increasing the density of the or grid. A basic question is, how tight must the be in order to obtain reasonable results from model. The penalty for a tight mesh with very s wire segments is large computer storage area and l execution time.

Other studies, where conducting surf have been replaced with a wire-grid model, sug that the wire segment length should not be gre than one-tenth of the free space wavelength at frequency being considered [4,8]. It is reasonable to assume that the conducting ground p for the slot antenna should be modeled with a where the largest wire segment is less than one-t wavelength. We are interested in modeling rectangular slot antenna near its first reso frequency. This frequency occurs when the length, L, is approximately one-half wavelength. meet the above conditions as well as preserve symm in the model, the wire segment lengths were chose be one-twelfth of a wavelength. This translates the six wire segments across the length of the slo shown in Fig. 2. In all, the basic wire-grid m for the rectangular slot antenna has 555 wire segm (one segment is required to represent the vol source which feeds the antenna).

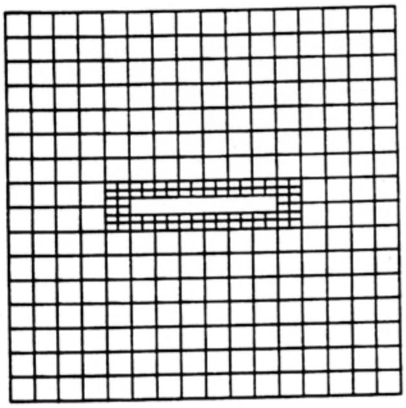
Another important question regarding wire mesh density is, how sensitive is the mode increasing and decreasing the tightness of the m To answer this question, it was decided that the density of the model would be varied and the term impedance, as calculated by NEC, would be comp with that of the basic model (Fig. 2). Term impedance was chosen for this comparison since i very sensitive to variations in the near-f behavior of an antenna.

Starting with the basic 555 segment mo the mesh density near the slot opening was incre as shown by the two models in Fig. 3. If a tig mesh is required in modeling the slot antenna, one would expect to see large differences between values of terminal impedance predicted by different models. Figure 4 shows the calcul resistive and reactive components of the term impedance for the 555, 641, and 683 segment wire- models. The impedance terms are plotted as a func of the ratio of slot length, L, to free s wavelength, λ. Very little change in the calcul impedance is observed when the mesh density in basic model is increased near the slot. indicates that decreasing wire segment length be the one-tenth wavelength criterion provides neglig benefits.

The basic 555 segment model can also be to study the effect of decreasing wire mesh dens



(a) 641 Segments



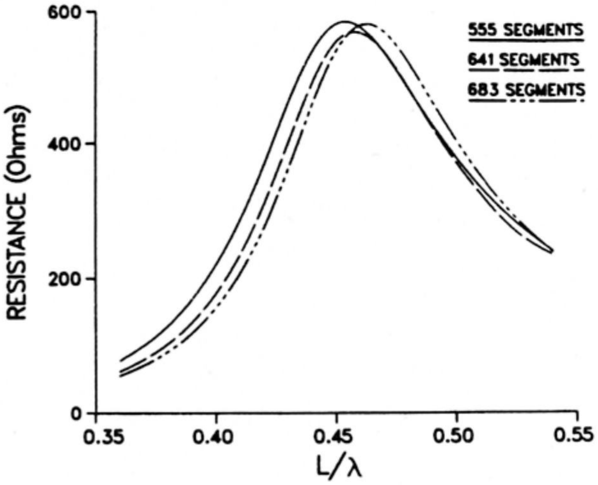
(b) 683 Segments

Figure 3. Wire-Grid Models with an Increased Number of Segments.

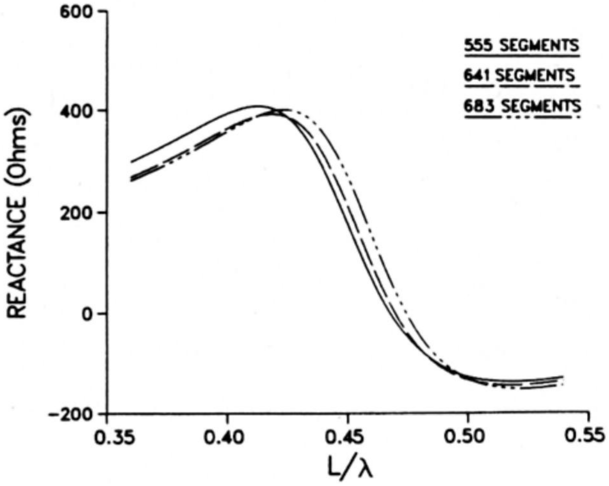
Figure 5 shows two additional models for the slot antenna where the density of the mesh has been decreased in areas away from the slot. The calculated impedance components for these 383 and 209 segment models are compared with the basic 555 segment model in Fig. 6. Very little difference in impedance is detected between the 383 and 555 segment models. There is, however, a substantial change in impedance between the 555 segment and the 209 segment models. Thus, the impedance values calculated by the model are sensitive to increases in wire segment length, beyond the one-tenth wavelength criterion, in the vicinity of the slot. If it is absolutely necessary that the number of wire segments composing the model be decreased, this should only be done in regions that are far removed from the slot.

B. Wire Segment Radius

When solid surfaces are modeled using the wire-grid approach, wire segment radius is an important parameter. In modeling targets for radar cross-section studies, it has been found that the best results are obtained when the wire radius is chosen to be between 0.005 λ and 0.01 λ [8]. Since the length of wire segments in wire-grid modeling is approximately one-tenth wavelength, the wire radius of the segments should then be one-tenth of the wire length for accurate results.



(a) Resistance

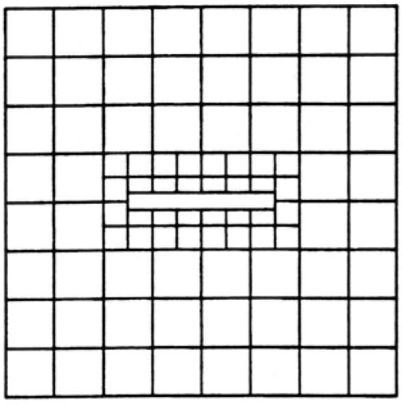


(b) **Reactance**

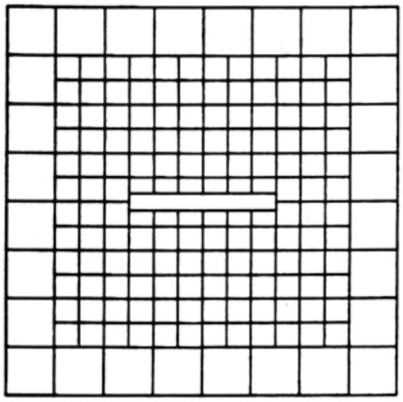
Figure 4. Effect of Increased Number of Modeling Wire Segments on Calculated Antenna Impedance.

Additional work which supports this choice for wire radius has been done by Moullin [9]. He found that solid reflective screens for antennas could be replaced by a series of parallel wires, if the spacing and radius of the wires were properly chosen. This choice is based on the theory that when the self-inductance of the wires is equal and opposite to the mutual inductance between wires, the wires will behave as a solid sheet. Again, for the one-tenth wavelength spacing in the wire-grid, Moullin's work indicates that the wire radius should be approximately one-tenth of the wire segment length.

The basic 555 segment wire-grid model for the slot antenna was used to calculate the terminal impedance for different values of wire segment radius. The results are shown in Fig. 7, where wire radius is given in terms of L, the length of the

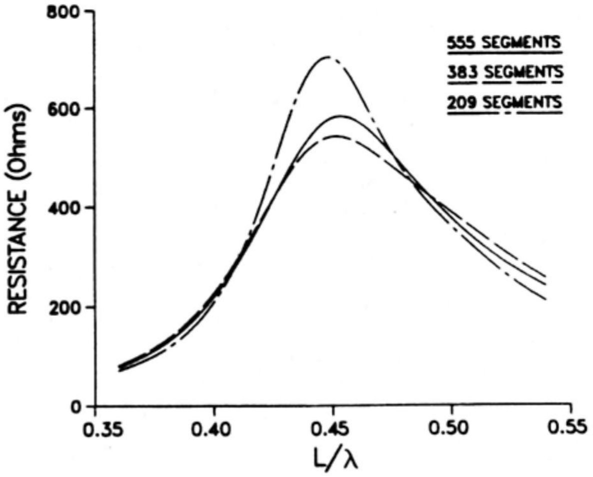


(a) **209 Segments**

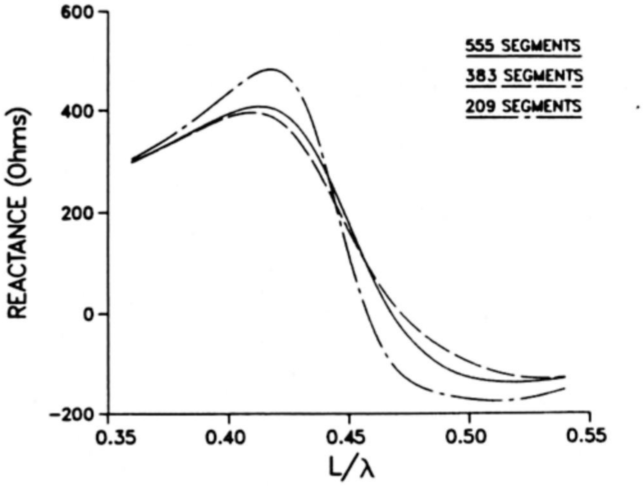


(b) **383 Segments**

Figure 5. Wire-Grid Models with Decreased Number of Segments.



(a) **Resistance**



(b) **Reactance**

Figure 6. Effect of Decreased Number of Modeling Wire Segments on Calculated Antenna Impedance.

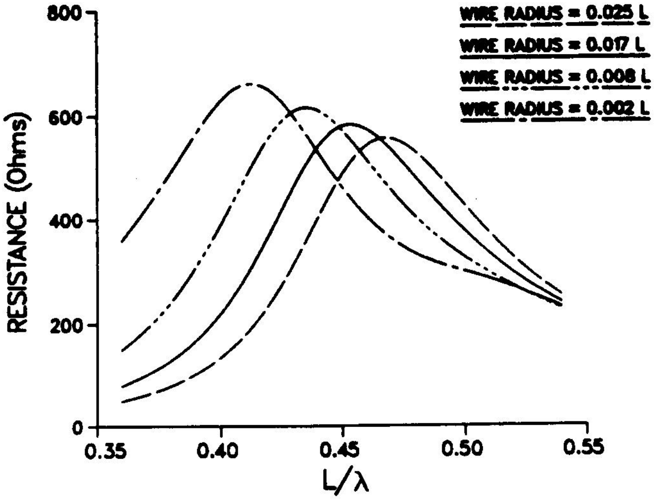
slot. This data shows that the model is very sensitive to changes in wire radius. If one uses the criterion that the wire radius should be one-tenth of the wire segment length, then the most accurate impedance curves, in Fig. 7, will be those with wire radius equal to 0.017 L. A later comparison of calculated and measured impedance values will show that this is indeed the case.

C. Modeling Thin Slots

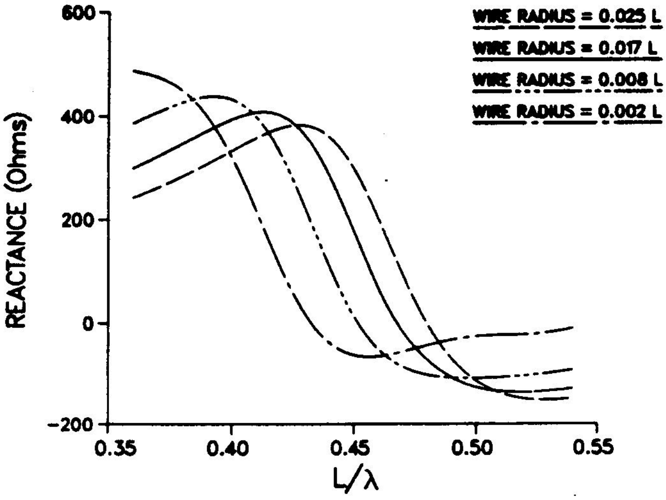
Originally, one of the major concerns associated with the wire-grid approach was that of modeling very thin slot antennas. The NEC user's guide cautions that placing parallel wires close together, within a few radii, may cause numerical instability and inaccurate results.

The 555 segment slot antenna model was again used to determine the capability of the wire-grid technique to model thin slots. Figure 8 gives the calculated terminal impedance for different values of slot width to slot length ratio. The modeled slot width is the distance between the centers of the wire segments bounding the sides of the slot. Even though the centers of the wires are brought to within 2.33 radii, the calculated impedance values show no instability. These results also show that very little effect on the calculated impedance can be expected if the model slot width is decreased beyond the W/L = 0.039 value.

One way to qualitatively test the results from the thin slot model is to compare them with the impedance values predicted theoretically for thin slot antennas located in infinite ground planes. Figure 9 shows the behavior of very thin slots for the infinite ground plane case. These impedance values were obtained by using Schelkunoff's method for determining the impedance of thin dipoles and then applying Babinet's principle to obtain the impedance of complementary slot antennas [6,10]. The infinite ground plane slots show very little change in terminal impedance as the slot width is decreased, just as suggested by the wire-grid model. In fact, the impedance calculated from the wire-grid model, with W/L = 0.039, closely approximate the theoretical

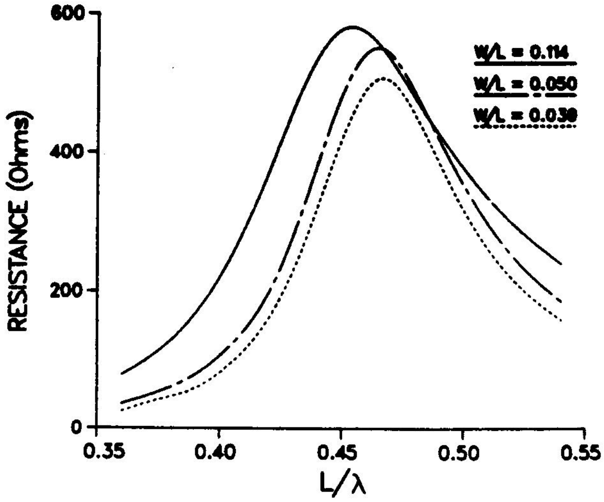


(a) Resistance

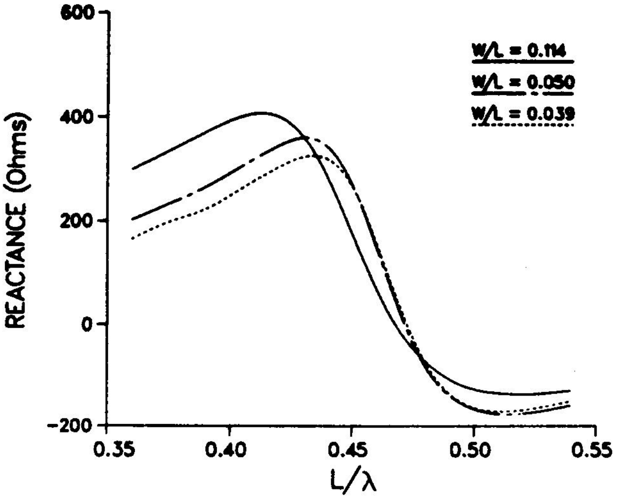


(b) Reactance

Figure 7. Calculated Slot Antenna Impedance for Different Values of Wire Segment Radius in the Wire-Grid Model.



(a) Resistance



(b) Reactance

Figure 8. Calculated Impedance for Narrow Slot Antennas Using the Wire-Grid Model.

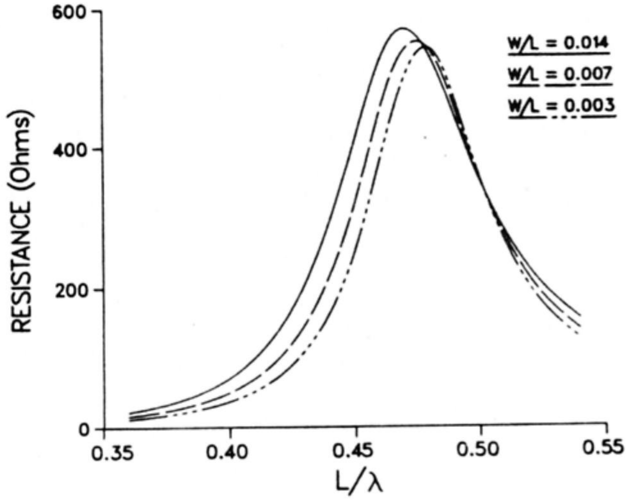
results for the very thin slots of Fig. 9. It then appears that the wire-grid model can be used to obtain reasonable calculations for terminal impedance even when the slot has a very small width.

D. Wire Segment Impedance

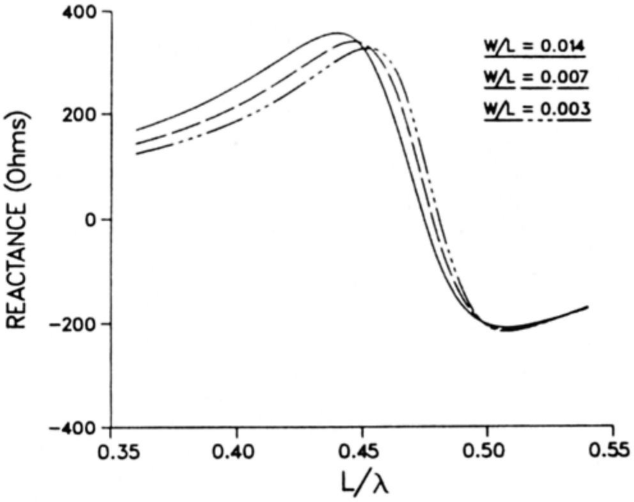
Up to this point in the discussion, it has been assumed that the slot antenna being modeled has an air-filled slot. As stated earlier, for the wire-grid model to be practical, it must include a means for incorporating the effect of a thin dielectric material in the slot region. Fortunately, the NEC program has the capability of impedance loading wire segments composing the model.

If one considers the slot antenna to behave as a slot transmission line, then the effect of a slot dielectric is to increase the distributed capacitance of the slot line. The effect of the dielectric can then be included in the wire-grid model by placing discrete, equal valued capacitors across the slot as shown in Fig. 10. The only requirement is that the capacitors should be spaced no greater than one-tenth wavelength along the slot. The capacitance will then appear to be approximately distributed in nature, since an artificial transmission line has been created [11].

The value of capacitance used in the model will depend upon the slot width as well as the type and thickness of the dielectric material. At this time, a theoretical technique has not been developed to establish an exact value for the capacitors. The capacitance is determined by measuring the resonant frequency of a slot antenna with and without the dielectric backing. The value of the capacitors in the model is then chosen to match this behavior. It should be stated that longer slot antennas, having the same width and dielectric, can be modeled by merely distributing more of the capacitors along their length.



(a) **Resistance**



(b) **Reactance**

Figure 9. Theoretical Impedance for Narrow Slot Antennas in Infinite Ground Planes.

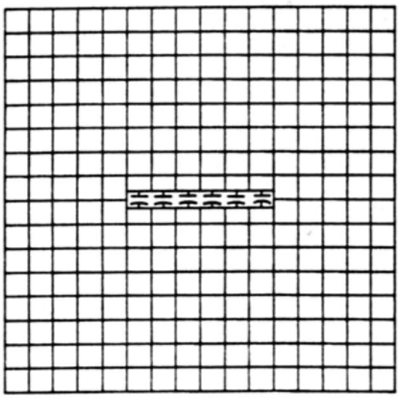


Figure 10. Wire-Grid Model for a Dielectric-Backed Slot Antenna.

IV. NUMERICAL AND EXPERIMENTAL RESULTS

Several experimental measurements were made verify the performance of the wire-grid slot ante model. A slot antenna was constructed and terminal impedance and radiation patterns have b measured. In this section of the report, experimental data is compared with the numeri results predicted by the wire-grid model.

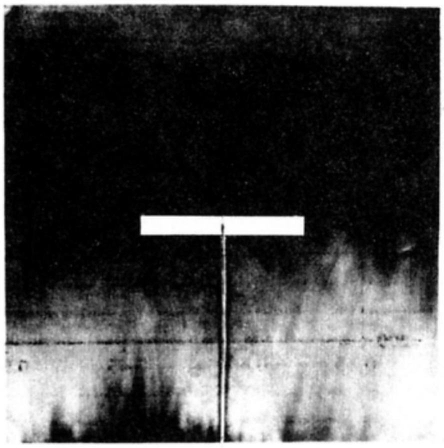


Figure 11. Photograph of the Experimental S Antenna.

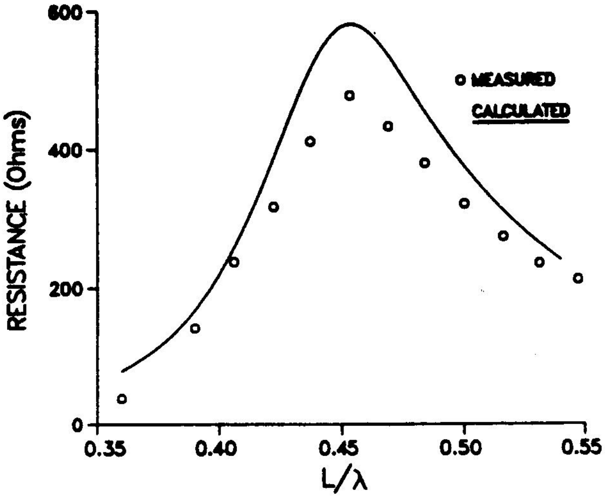
A. Slot Antenna Fabrication

A simple slot antenna was constructed correspond to the dimensions of the 555 segment wi grid model given in Fig. 2. Figure 11 shows photograph of the experimental slot antenna and coaxial feeding structure. The antenna was made f a 0.5 m square sheet of copper having a thickness 1.4 mm. The length of the slot cut into the sheet 0.188 m and the width was 0.021 m. To simulat dielectric in the slot, a sheet of SMC (sheet mold compound) plastic was used. This sheet of plastic 2.8 mm thick and was tack glued to one side of copper sheet when a slot dielectric was needed.

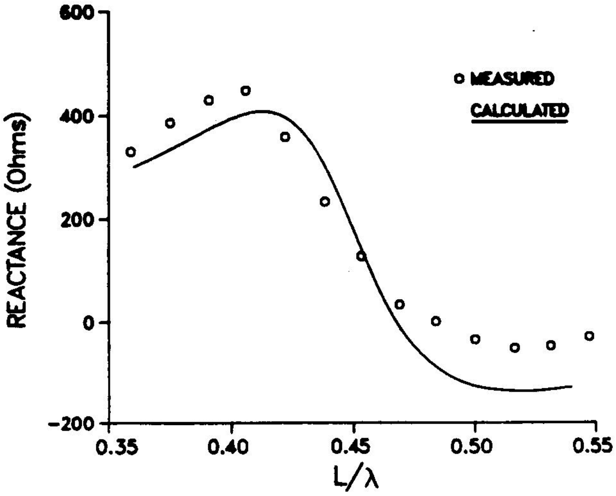
B. Terminal Impedance

The procedure used to measure the termi impedance of the experimental slot antenna relatively simple. The antenna was placed 1.5 m ab the ground and several meters from any structure wh might interfere with the measurements. A netw analyzer was then used to measure the impedance of antenna at several frequencies in the range of 550 800 MHz. The network analyzer was calibrated indicate an open circuit for the coaxial feed bef it was attached to the slot antenna. This reduced parasitic effects of the feed structure on impedance measurements. The terminal impedance of slot antenna, with and without the dielectric backi was measured in this fashion.

Figure 12 compares the measured impeda values with the calculated impedance obtained from 555 segment wire-grid model. In this case the s was not backed with the dielectric material. It generally very difficult to obtain an accurate ma between calculated and measured antenna impedance here, there is reasonably close agreement.



(a) Resistance

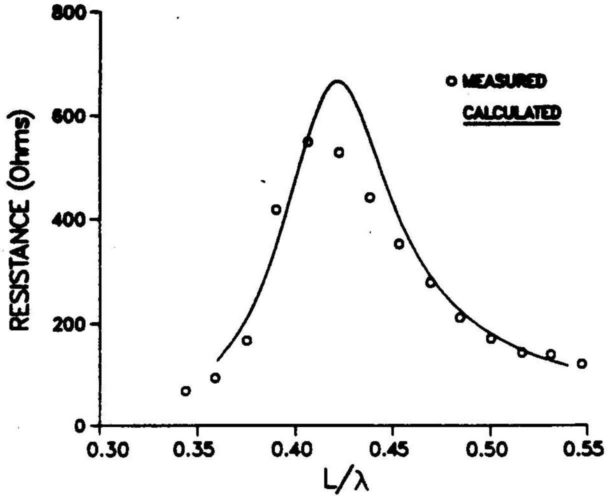


(b) Reactance

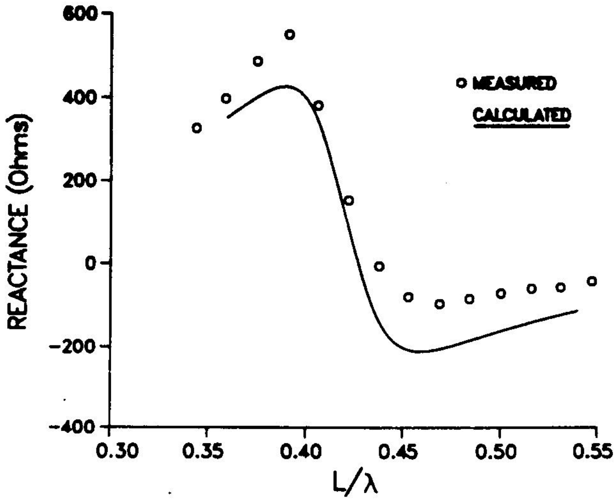
Figure 12. Measured and Calculated Impedance for an Air-Filled Slot Antenna.

Measured and calculated terminal impedance for the dielectric backed slot antenna are presented in Fig. 13. It was found that placing six 0.8 pF capacitors across the basic slot model, as shown in Fig. 10, provided a reasonable match between the measured and calculated values. Comparing the data in Figs. 12 and 13, one sees that the effect of the dielectric is to shift the resonant frequency of the slot antenna to a smaller value of L/λ. The capacitors which have been placed across the modeled slot account for this effect.

It then appears that the wire-grid modeling approach can be successfully used to predict the terminal impedance of a slot antenna with and without a dielectric backing. The degree of agreement between measured and calculated values shows that this modeling technique can be used to provide preliminary design data for matching the antenna and estimating such quantities as antenna bandwidth.



(a) Resistance



(b) Reactance

Figure 13. Measured and Calculated Impedance for Dielectric-Backed Slot Antenna.

C. Radiation Patterns

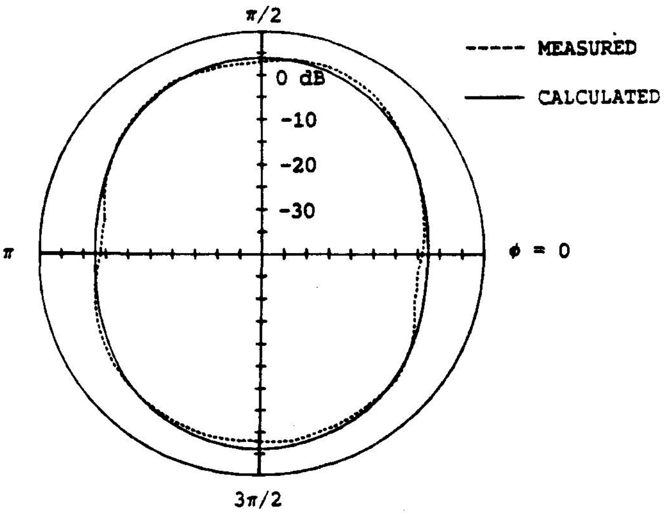
Two sets of radiation patterns were als measured for the experimental slot antenna with an without its dielectric backing. These patterns wer measured at an outdoor test range which had a inground turntable available for rotating a tes antenna [12]. The slot antenna was placed over tr center of the turntable and elevated 1.5 m with styrofoam support. A tuned dipole was place approximately 8 m from the slot and orientated t receive the desired polarized signal radiated from tr slot.

Considering the co-ordinate system c Fig. 1, the electric field radiated from the slc antenna is principally polarized in the θ-directior The patterns measured on the experimental antenna the consisted of two cuts of the θ-polarized radiatior One pattern was obtained by varying ϕ thru 2π radiar (table rotation) while holding θ fixed at π/2. Fc the other pattern, ϕ was fixed at $\pi/2$ and θ was allowed to vary thru 2π radians. Physically, this was accomplished by placing the plane of the sheet containing the slot antenna perpendicular to the plane of the turntable and then orienting the slot either vertically or horizontally for each pattern measurement.

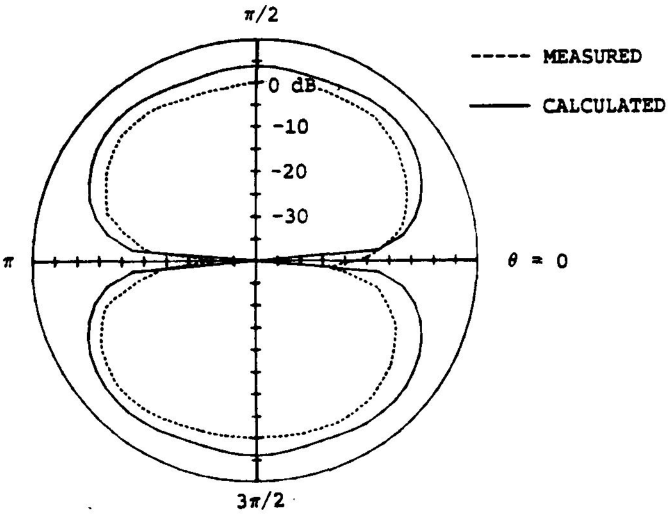
The two θ-polarized patterns for the slot antenna with and without the dielectric were measured at their approximate resonant frequencies of 700 and 765 MHz, respectively. A comparison of the measured radiation patterns with those predicted from the wire-grid models are given in Figs. 14 and 15. The calculated and measured θ-polarized patterns as a function of ϕ are in excellent agreement for both the air and dielectric backed antennas. A comparison of the patterns, which vary with θ, show the same shape, but the measured patterns are approximately 4 dB below the calculated data in both cases.

With the angles ϕ and θ both fixed at $\pi/2$, the level of both of the measured θ-polarized patterns should be identical since the same polarization and position are being considered. The 4 dB difference must then be attributed to measurement error and not to a lack of accuracy in the model. As one can see from the shape of the radiation patterns, the ground and metal turntable top will be illuminated differently when measuring the two θ-polarized patterns. The discrepancy in the measured data was most likely produced by a change in reflections from these surfaces. With this in mind, the above data shows that the wire-grid model gives very accurate predictions of the radiation patterns for the slot antenna.

It should be noted that in the above comparisons of numerical and experimental data, the calculated radiation patterns were not merely normalized to fit the experimental patterns. A

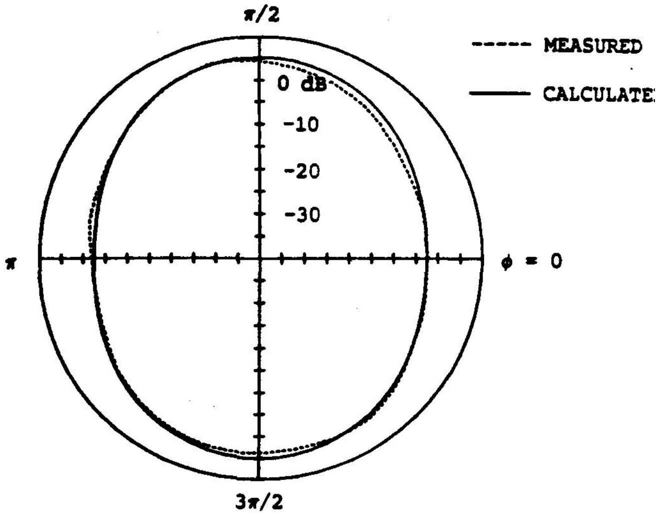


(a) $\theta=\pi/2$.

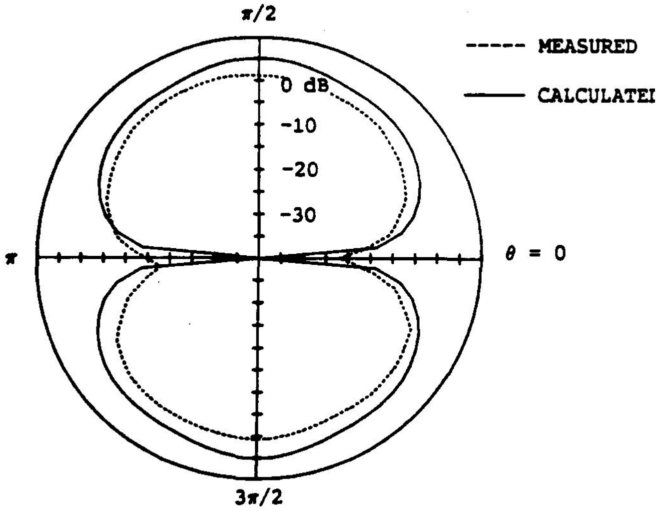


(b) $\phi=\pi/2$.

Figure 14. Measured and Calculated θ-Polarized Radiation Patterns for an Air-Filled Slot Antenna.



(a) $\theta=\pi/2$.



(b) $\phi=\pi/2$.

Figure 15. Measured and Calculated θ-Polarized Radiation Patterns for a Dielectric-Backed Slot Antenna.

calibration factor (in dB) was determined and added to the calculated patterns. At each of the measurement frequencies, the calibration factor was obtained by first measuring the horizontally polarized radiation of a tuned dipole antenna and then using the NEC program to calculate the pattern. The calibration factor was determined by subtracting the maximum value on the predicted pattern from the corresponding point on the measured pattern. This difference then became the calibration factor to be added to all patterns computed by the models at that frequency.

V. SUMMARY

The wire-grid modeling option of the Numerical Electromagnetic Code has been used to simulate a rectangular slot antenna in a finite ground plane. The effects of varying the key modeling parameters, wire radius and wire segment length, were presented. The importance of selecting the proper values for these parameters, when modeling slot antennas, was discussed. A technique was developed for including, in the wire-grid model, the effects due to backing the antenna with a dielectric material.

A comparison of numerical and experimentally measured data showed that wire-grid modeling can be used to simulate slot type antennas. The radiation patterns predicted by the model are very accurate. Calculated values for antenna impedance were shown to agree well enough with experimental results to provide preliminary design information regarding antenna bandwidth, resonant frequency, and resistance at resonance.

ACKNOWLEDGMENTS

The author wishes to thank Dr. R. Adler of the Naval Postgraduate School in Monterey, California for providing a Fortran version of the Numerical Electromagnetic Code. The author also wishes to gratefully acknowledge all of the persons contributing to development of the NEC program at the Lawrence Livermore Laboratory in Livermore, California. Without this program, none of the work reported here could have been accomplished. Special thanks is also due Mr. M. O'Rourke at the General Motors Research Laboratories for his assistance in constructing and measuring antennas.

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FOUR-ELEMENT BEVERAGE ARRAY

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An experimental four-element high frequency Beverage antenna array has been devised for tests on medium range and long range ionospheric circuits. The radiation pattern of the experimental array was calculated using Numerical Electromagnetics Code (NEC) and found to be skewed in azimuth. The antenna installation was subsequently modified in accordance with NEC predictions and the desired beam direction was obtained. Measurements confirmed that the NEC predictions were valid.

I. INTRODUCTION

The Beverage Antenna (H. H. Beverage, C. W. Rice, and E. W. Kellogg, The Wave Antenna, A New Type of Highly Directive Antenna, Trans. AIEE, 1923) can provide a marked unidirectional pattern. This property and its inherent simplicity and broadband behavior make the wave antenna an attractive option for use on medium and long range HF ionospheric radio circuits.

An experimental four-element Beverage array has recently been investigated and tested on HF long range circuits between Fort Monmouth, New Jersey and Los Banos, California and several intermediate sites. Some results are discussed below.

II. SINGLE WIRE WAVE ANTENNA

The simplest Beverage wave antenna consists of a single long horizontal wire terminated in its characteristic impedance to ground at each end. See Fig. 1. It is known that the wave antenna depends for its operation on the finite conductivity of the earth and, in fact, it operates better over poorly conducting ground. For an incident plane wave over imperfect ground there is a horizontal component of the electric field intensity due to the power absorbed by the earth. This horizontal component induces the voltage in the wire and gives the Beverage antenna its directivity.

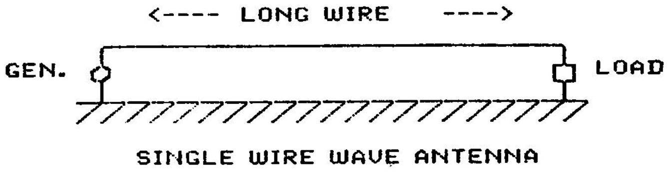


Figure 1.

The efficiency of a single element Beverage wave antenna is usually low. The power available from the wave antenna over real ground can be compared with that available from a lossless short vertical antenna over perfect ground (S. A. Schelkunoff and H. T. Friis, Antennas-Theory and Practice, J. Wiley and Sons, 1952, pp. 492-497). This power ratio can be roughly approximated as follows.

$$\frac{\text{P WAVE ANTENNA}}{\text{P VERTICAL ANTENNA}} = \frac{8\pi^2}{3\lambda\sigma Z_0}\left(\frac{\ell}{\lambda}\right)^2$$

Where σ is the earth conductivity, λ t wavelength and ℓ and Z_0 are respectively the leng and "characteristic impedance" of the wire.

For example, with $\sigma = .01$ (good ground), $\lambda = 15$ $Z_0 = 400$ we find that

$$\text{PW.A./PVERT.} = 0.0439 \left(\frac{\ell}{\lambda}\right)^2$$

and in this case a wave antenna about 4.8 wavelengt long should give as much power output as a lossle short vertical antenna over perfect ground. If t conductivity is low, for example $\sigma = .001$, $\lambda = 15$ and $Z_0 = 400$ as before, we find that

$$\text{PW.A./PVERT.} = 0.439 \left(\frac{\ell}{\lambda}\right)^2$$

and a shorter wave antenna 1.5 wavelengths long shou give the same power output as a lossless sho vertical antenna over perfect ground. The abo approximation is optimistic but is useful for rou estimates of wave antenna performance over actu ground.

III. ARRAYS OF WAVE ANTENNAS

An array of wave antennas can be used to offs the low efficiency of the single element antenna. T array technique is certainly not new and has in fa often been used to overcome the low efficiency various electrically small antennas.

Recently a four-element wave antenna w investigated at Fort Monmouth as part of an HF te series which also included a terminated sloping antenna. This wave antenna array consists of fo parallel horizontal wires 143 meters in length stru 2.4 meters above ground and 6.7 meters apart. S Fig. 2. The wires are individually excited throu step-up transformers and a power divider with equ amplitude in-phase signals and terminated at the en nearest the distant station in grounded 400-o resistors. Because of an access road at the te site, the antenna wires had to be installed in staggered arrangement so that the maximum wire leng (470 feet) could be accommodated without crossing t road. This unconventional antenna installation turn out to be unacceptable, however, because it result in skewed radiation patterns.

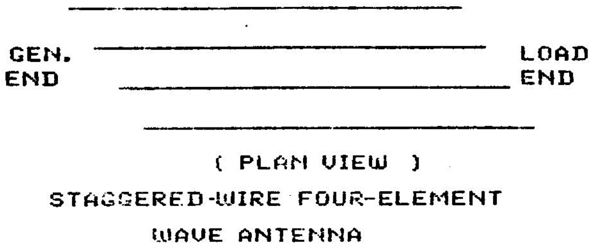


Figure 2.

IV. NEC MODELS OF WAVE ANTENNA ARRAYS

First performance tests of the staggered wire wave antenna array operated in the transmitting mode were not up to expectations. Faulty antenna operation was brought to light by an analysis of the signal strength received at the Los Banos test site where it was noticed that a single-wire Beverage antenna at Fort Monmouth was outperforming the Beverage array. The single-wire Beverage antenna was obtained by simply disconnecting three of the wires comprising the experimental array.

Subsequent analysis using NEC (Numerical Electromagnetics Code) showed that the Beverage array radiation patterns were shifted nearly 20-degrees off bearing as a result of the misaligned (staggered) wire arrangement.

The aligned (correct) and skewed-wire (incorrect) experimental arrays were modeled with NEC-2. The ground connections were approximated in the NEC computer model by open-ended quarter wavelength wire extensions as shown in Fig. 3 (Laport, p. 309). An open quarter wavelength wire gives an impedance at its inner end that is low over a narrow frequency band near its resonant frequency. These quarter wavelength wires thus provide artificial "ground" connections at the wire ends of the excitation and for the resistive terminations.

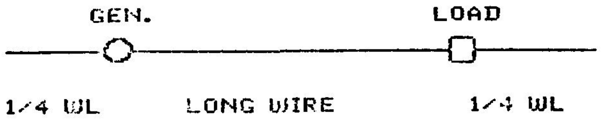


Figure 3.

The NEC "Sommerfeld ground" option was used i the computer models and "good" ground was assume (σ = .01 mhos/m and $\varepsilon/\varepsilon_o$ = 15). The actual groun constants at the antenna test site are as yet unknow but will eventually be determined from in sit measurement by the wave-tilt method and/or th transmission line method.

A. Calculated Radiation Patterns of Misaligne Four-Element Beverage Array

The computed azimuthal and elevatio radiation patterns of the staggered wire four-elemen wave antenna are shown in Figs. 4 and 5 (8 MHz) and i Figs. 6 and 7 (15 MHz). These patterns show the mai beam to be skewed so that it does <u>not</u> lie along th bearing to the distant station.

B. Calculated Radiation Patterns of Aligne Four-Element Beverage Array

The computed azimuthal and elevatio radiation patterns of the aligned four-element wav antenna are shown in Figs. 8 and 9 (8 MHz) and i Figs. 10 and 11 (15 MHz). These patterns show tha the main beam lies along the array axis and on th correct bearing to the distant station as intended.

Comparison of these computed patterns show that the array gain on the main bearing was reduced b 7 db (at 8 MHz) and by 15 db (at 15 MHz) as a resul of the misalignment of the antenna wires.

V. EXPERIMENTAL RESULTS

As mentioned earlier it was noticed durin preliminary operational tests that a single-wir

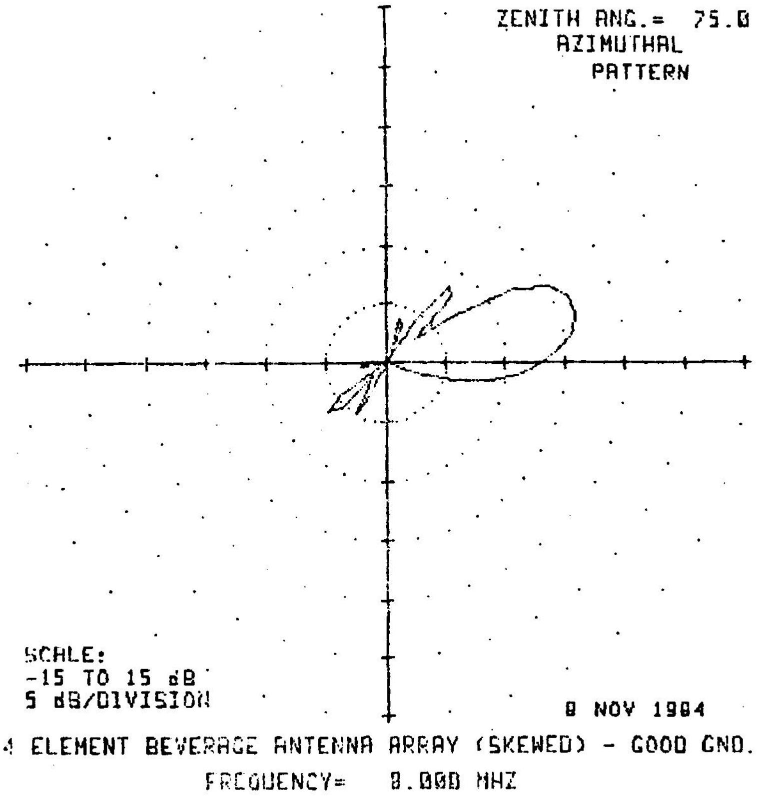


Figure 4.

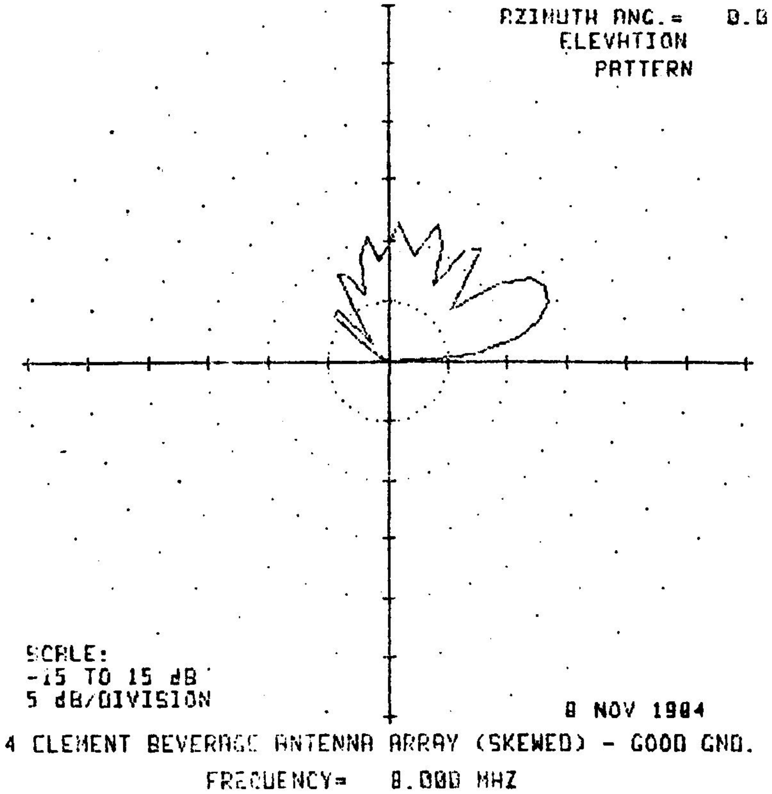
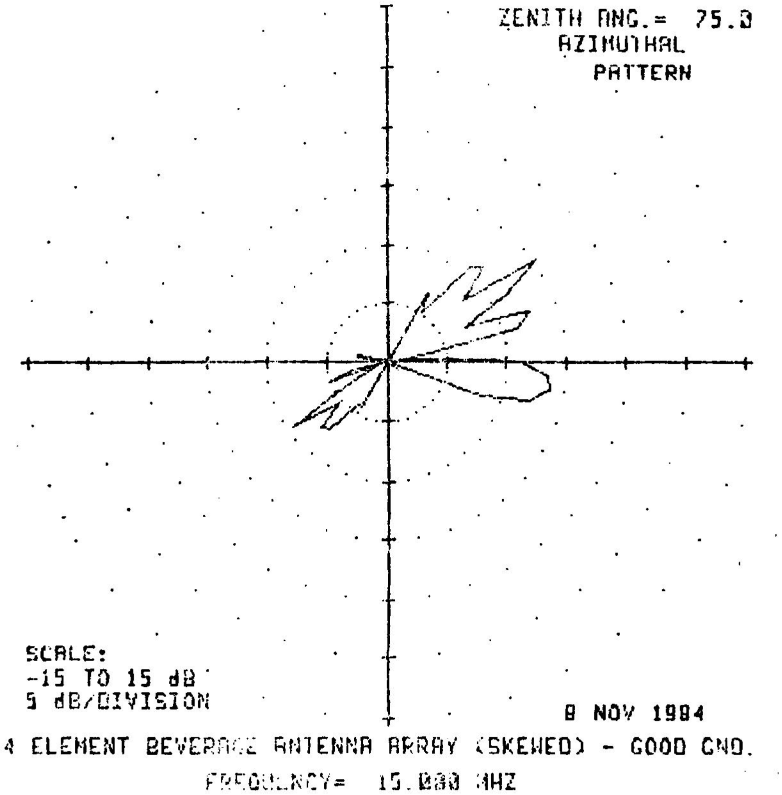


Figure 5.



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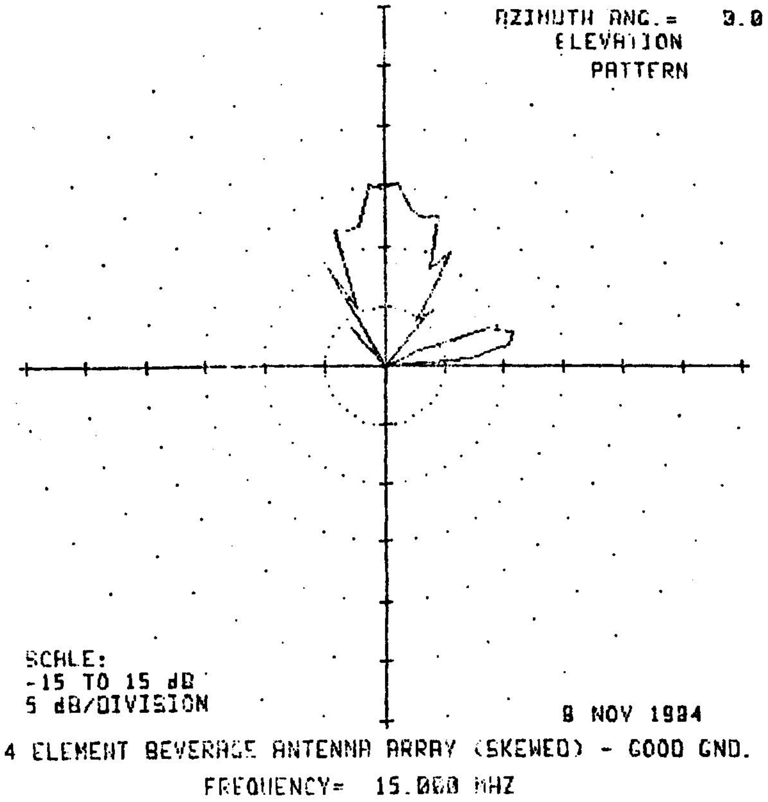


Figure 7.

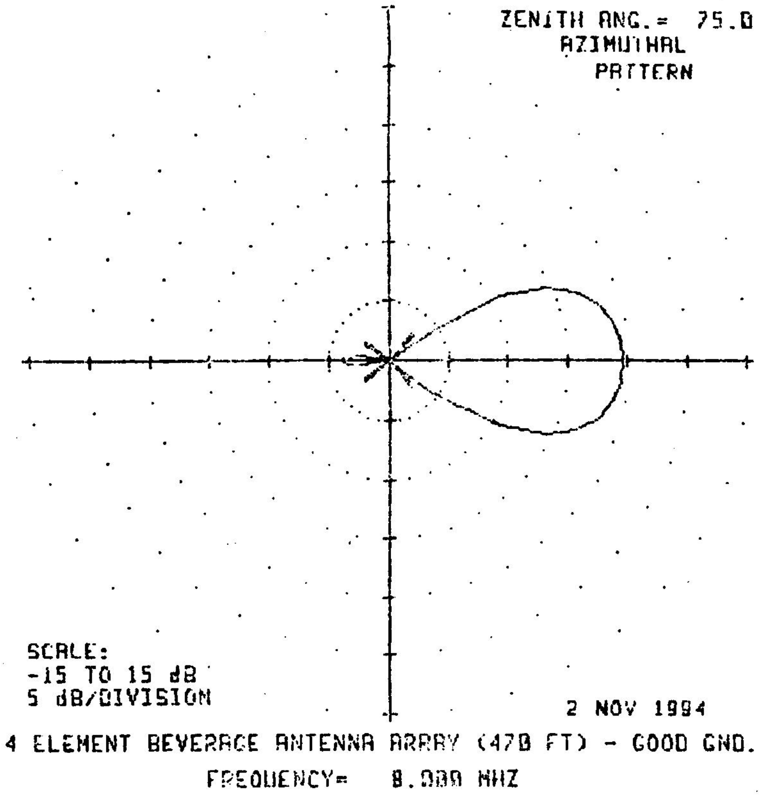


Figure 8.

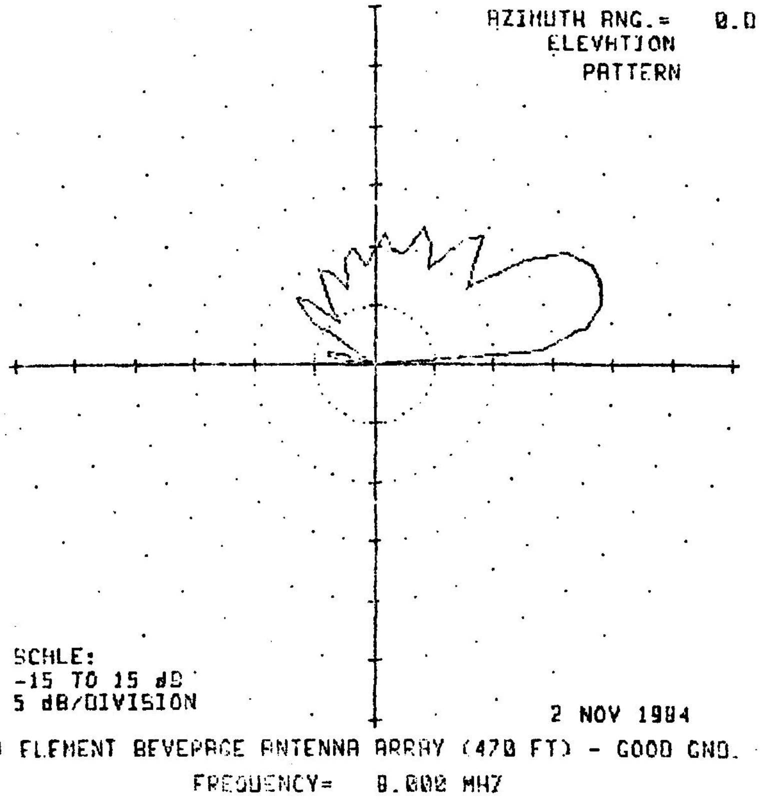


Figure 9.

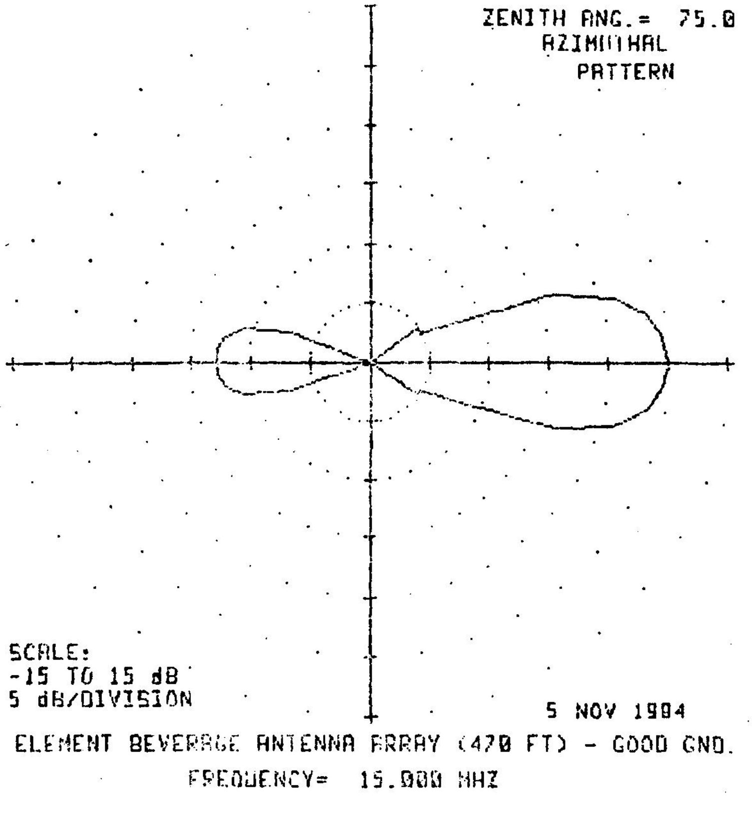
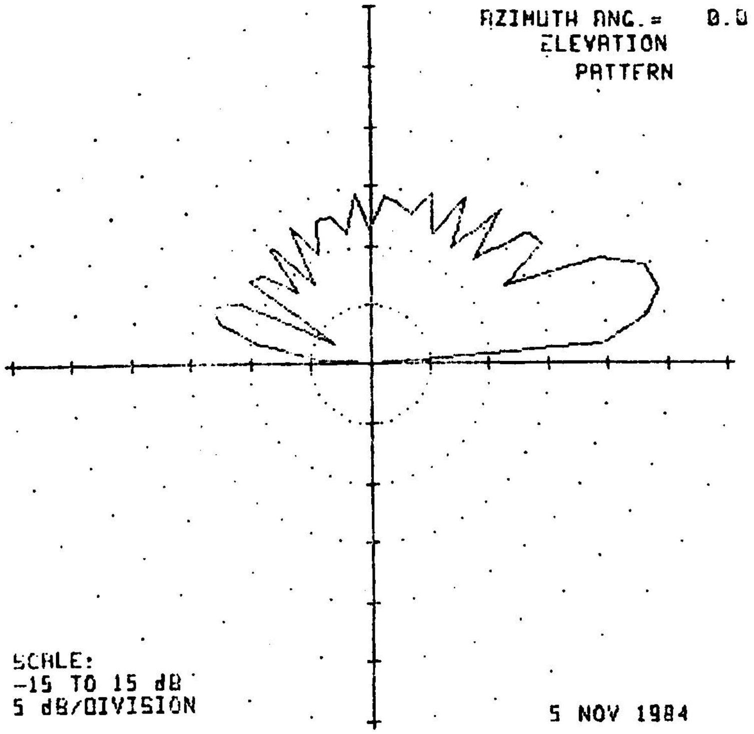


Figure 10.



4 ELEMENT BEVERAGE ANTENNA ARRAY (470 FT) - GOOD GND.

FREQUENCY= 15.000 MHZ

Figure 11.

Beverage antenna outperformed the misaligned Beverage array. This discovery and the subsequent NEC analysis confirmed that the poor performance of the first array was due to the staggered wire arrangement. A 15 db increase in received signal strength at the Los Banos site resulted when the array wires were properly aligned. The wires also had to be somewhat shortened to a length of 125 meters because of the access road.

VI. CONCLUSION

The NEC program was successfully used to determine the radiation characteristics of an experimental four-element array of Beverage antennas and correctly predicted pattern distortion and gain reduction caused by misalignment of the antenna wires.

MININEC APPLICATIONS GUIDE

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The Mini-Numerical Electromagnetics Code - MININEC has become very popular among amateur and professional antenna practitioners. Its widespread acceptance is due in no small part to its great versatility, user friendliness and extreme compactness which permits installation on PC's. Prior to MININEC, serious numerical modeling of antennas had to be accomplished primarily on mainframe computers. With MININEC, it is now possible to solve complex wire antenna problems in a conversational mode on a modest desk top computer, for example a PC having 64k of RAM. The present paper is concerned with application guidelines for MININEC. The input data set is discussed and numerical results are interpreted. A variety of illustrative linear antenna problems are worked out.

I. INTRODUCTION

The Mini-Numerical Electromagnetics Code-MININEC (Julian, Logan, Rockway, NOSC TD-516, 1982) has become very popular among amateur and professional antenna practitioners. With MININEC one can solve complex wire antenna problems in a conversational mode on a desk top computer. The present paper provides several supplemental worked out examples and information which elaborates on some of the original material for the benefit of the new user. The original MININEC document is out of print. However, a more recent version of MININEC is now available for those interested in details (Li, Rockway, Logan, Tam, "Microcomputer Tools for Communications Engineering," Artech House, Inc., 1983).

II. INPUT DATA

A. Geometry

Linear antennas are modeled by MININEC as one or more straight wires which may be located in free space or above perfectly conducting ground. The antenna geometry is set up by specifying the number of wires comprising the structure, the end point coordinates and radius of each wire, the number of segments in each wire, and connections between the various wires and to ground. Antenna excitation is specified after the geometry data is entered. Additional information is requested by the program concerning the frequency, impedance loading of wires, whether or not patterns are required and whether or not to print the currents.

In MININEC only straight wires are allowed. Bent wires are modeled by a sequence of straight wires. A wire is described by its radius and the x, y, z coordinates of its ends. Dimensions are in meters. The ground plane, if used, is the X-Y principal plane.

B. Segmentation of Wires

The number of segments in each wire is user specified when the antenna geometry data is entered. Each wire is then automatically divided by the code into equal length segments. The numerical solution for the currents is based on pulses (unknowns) centered at adjacent segment junctions.

A large number of segments (unknowns) will, in principle, lead to a more accurate solution. A curve can be prepared showing convergence of the current distribution or input impedance to a final value as the number of unknowns is increased within limits.

Numerical convergence of solutions is discussed in the original MININEC document. It is recommended that segmentation be varied to study the solution behavior. In some antenna problems coarse segmentation will suffice, for example in determining the qualitative behavior of antenna radiation patterns. To obtain accurate input impedance data, on the other hand, a large number of segments (unknowns, pulses) may be required.

C. Wire Connection Data

Wire connections are NOT automated in MININEC - connection data must be supplied by the user. Instruction for connection of wires is provided in the original documentation. However, the inexperienced MININEC user may become confused when entering wire connection data, especially when treating a complicated radiating system.

The MININEC wire connection rules are summarized in Fig. 1. An example of wire connection data is shown in Fig. 2. In Fig. 2 the wire ends are numbered. Connection data is shown inside the rectangular boxes.

Figure 1. MININEC WIRE CONNECTION RULES

Connection Data is required for each wire end after the wire end coordinates have been specified.

1. Proper connection requires <u>identical</u> coordinates for the end points of wires to be connected.

2. Zero elevation (Z=0) is required for wire connection to ground. The ground plane is in the X-Y principal plane.

3. Connection can only be made to a <u>previously specified</u> wire.

4. A zero indicates no connection (a free wire end).

5. A negative integer with magnitude equal to the wire number indicates a connection to ground.

6. Either end of a wire, but not both, may be connected to ground.

7. A negative integer with magnitude less than the wire number indicates that end one of the wire in question is connected to end one of an already defined wire ... or end two of the wire in question is connected to end two of an already defined wire.

8. A positive integer with magnitude less than the wire number indicates that end one of the wire in question is connected to end two of an already defined wire ... or end two of the wire in question is connected to end one of an already defined wire.

Example. If the wire is the first wire to be specified and end one is to be connected to ground, then -1 is used for the end one connection but 0 is used for the end two connection (even though end two may subsequently connect to another wire) because that wire has not yet been specified.

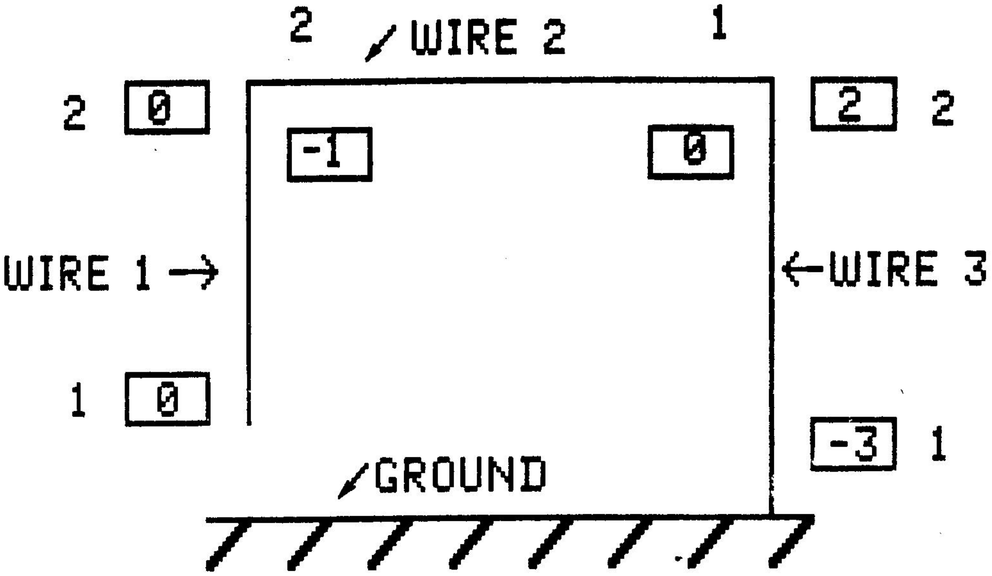


Figure 2.

III. SOME APPLICATIONS OF MININEC

A. Solution Convergence

In practice, accurate method-of-moments (MOM) solutions for antenna problems can be effected by allowing the number of unknowns (pulses, segments) to be sufficiently large within limits. The MOM solution procedure in MININEC employs pulse expansion functions and pulse testing functions. As such it is called a Galerkin method since the expansion functions and testing functions are the same.

A converged numerical solution can usually be realized with MININEC with only a moderate number of unknowns, depending on the problem. As an example, convergence of the input impedance to a final value is shown in Fig. 3 for a half-wavelength dipole antenna. For practical purposes, seven or nine pulses (unknowns) may suffice in this case.

In applying MININEC to a more complicated problem, some difficulty was experienced in obtaining a converged (stable) solution for the impedance of a "Y" antenna fed against ground. See Fig. 4. In this case fine segmentation was required to obtain convergence. Also, as the wire radius was decreased, even more segments were required.

The "Y" antenna was modified so that the wires intersect at right angles, as shown in Fig. 5. Convergence was then obtained with fewer unknowns.

The difficulty in securing convergence of MININEC in problems involving closely-spaced connected wires is under investigation and is mentioned here as a caution for users. It is expected that coding modifications will alleviate the need for excessive segmentation in problems of this sort.

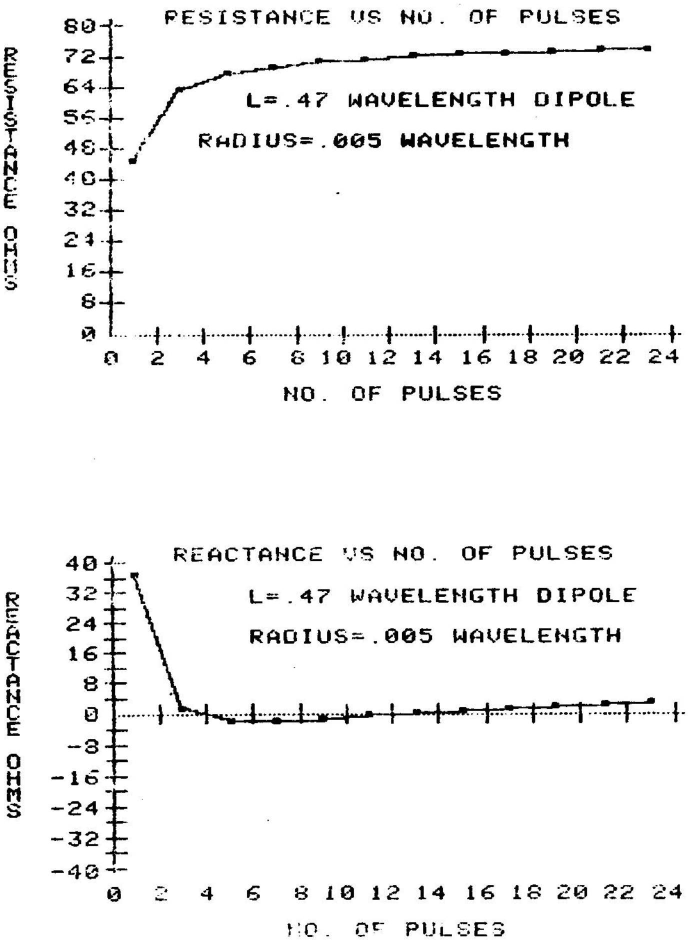


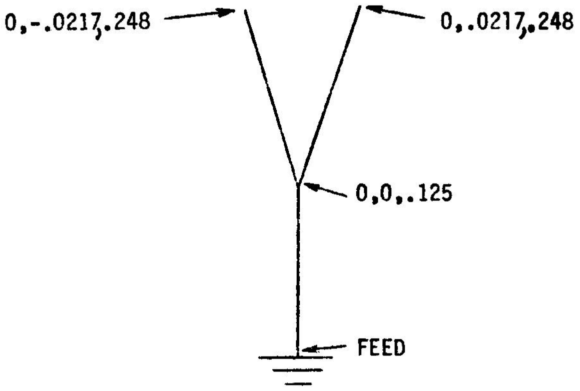
Figure 3.

B. Comparison of Measured and Computed Admittance

As part of the validation process, the admittance of a monopole antenna measured on a large ground plane was compared with the computed admittance. See Fig. 6. In trying to reconcile the results, it was found that a shift in susceptance occurred due to 10 picofarads capacitance added by the connector in the measuring set up. The equivalent circuit of the test set up is shown in Fig. 7. When this effect was accounted for, excellent agreement was obtained as seen in Fig. 8. Other validation data is discussed in TD-516.

f=299.8 MHz

WIRE RADIUS (m) a	30 SEGMENTS (10 per wire) Z ohms	45 SEGMENTS (15 per wire) Z ohms	60 SEGMENTS (20 per wire) Z ohms
.000001	2.83-j2906	4772-j22334	292.7+j2933.2
.00001	158.7+j868.5	79.9+j377.9	67.48+j281.23
.0001	63.25+j177.6	60.2+j157.8	
.001	62.9+j99.7	63.3+j99.1	



CONVERGENCE TEST - MININEC

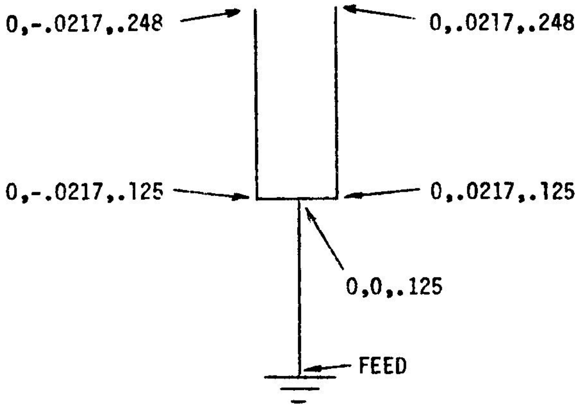
15 Dec 83
USACECOM

Y - junction of wires

Figure 4.

f=299.8 MHz

WIRE RADIUS (m) a	17 SEGMENTS (5,5,5,1,1) z ohms	40 SEGMENTS (12,12,12,2,2) z ohms	60 SEGMENTS (18,18,18,3,3) z ohms
.000001	48.59+j878.4	52.16+j526.4	53.28+j434.63
.00001	51.05+j214	54.42+j213.03	55.24+j216.83
.0001	52.96+j142.12	56.04+j161.02	56.91+j165.57



CONVERGENCE TEST - MININEC

15 Dec 83
USACECOM

90-degree wire junction

Figure 5.

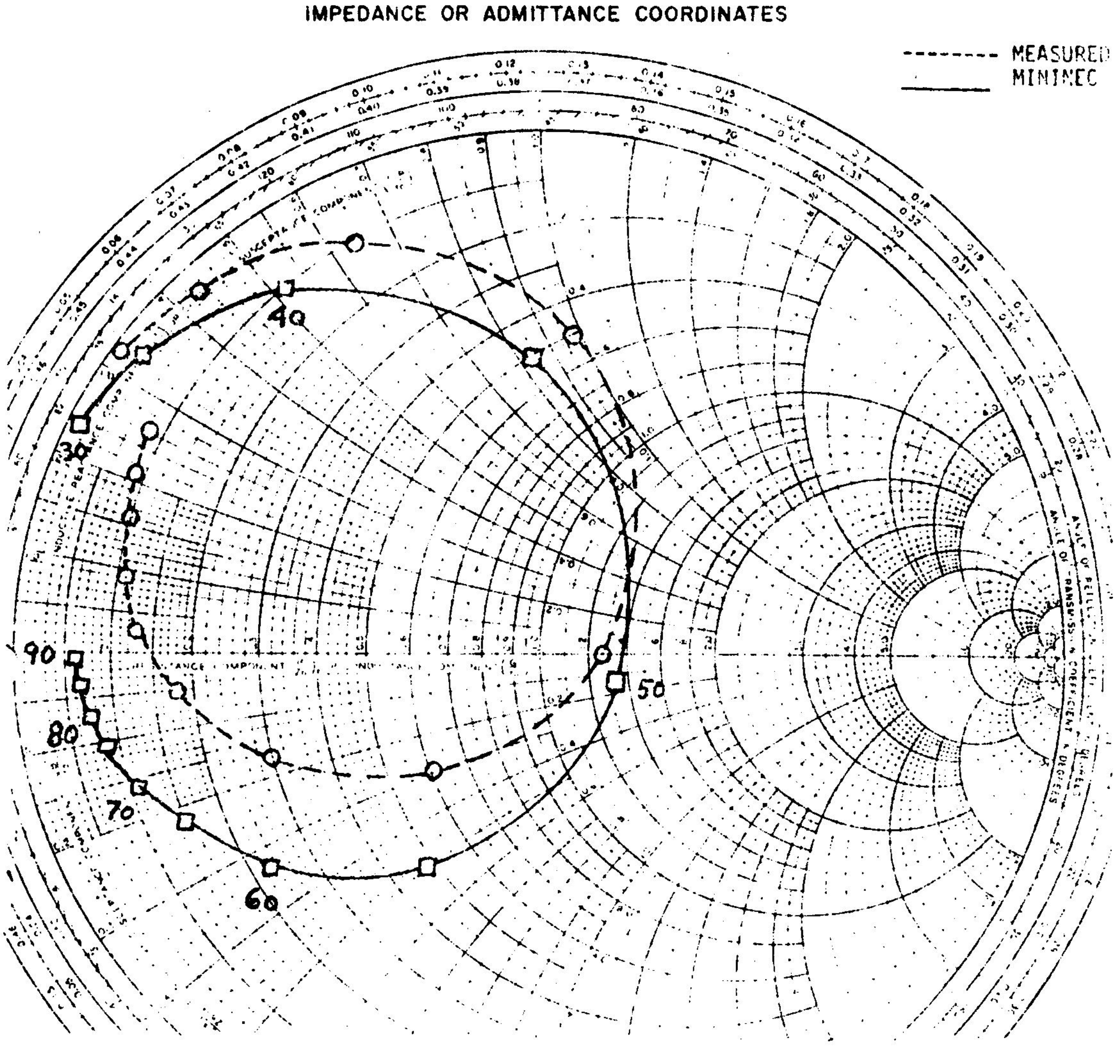


Figure 6.

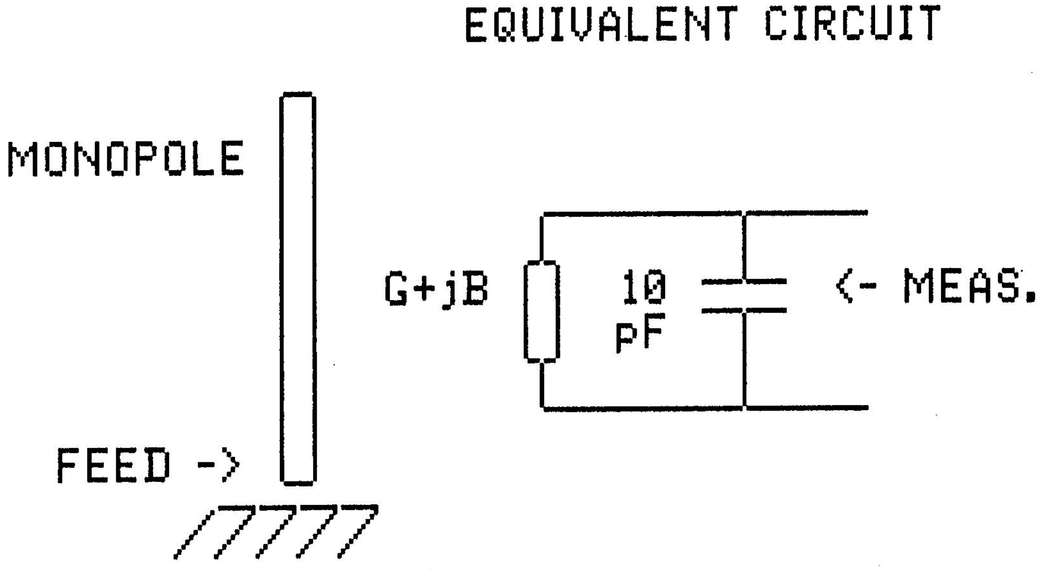


Figure 7.

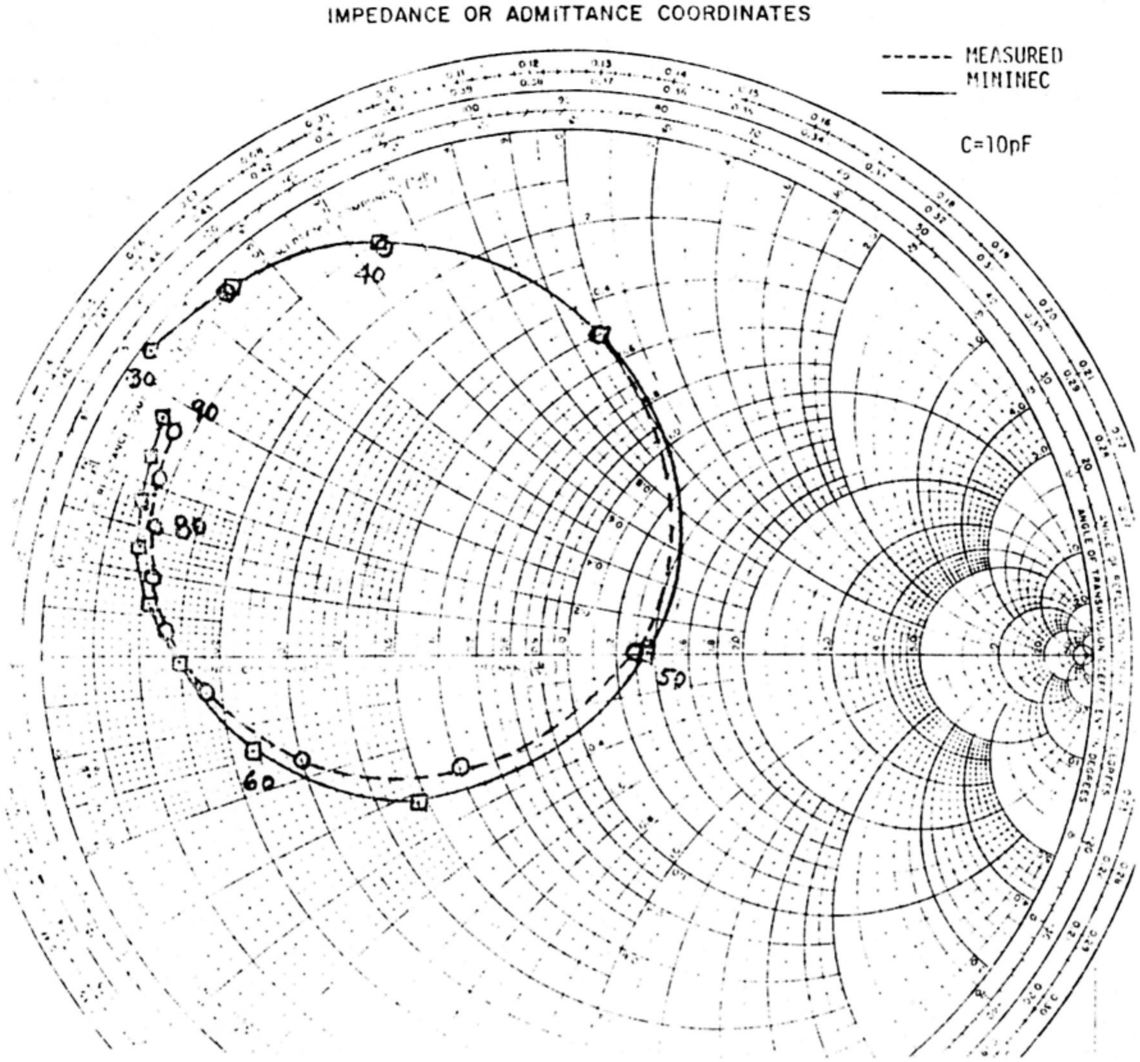


Figure 8.

C. Suppression of Parasitic Currents on Antenna Feed Lines

Parasitic rf currents may be induced on the outer surface of coaxial cable transmission lines and metal supports of an antenna system. These extraneous currents can degrade antenna performance and cause it to behave unpredictably with changes in the operating frequency.

MININEC was used to investigate the parasitic rf currents on a feed line located near a vertical half wave dipole configured as shown in Fig. 9. The calculated current distribution is shown in Fig. 10.

The amplitude of the parasitic currents can be significantly reduced by inserting high impedance reactive loads in series with the feed line. Suppression can be accomplished in a narrow frequency band by connecting a quarter-wave detuning stub to the feed line. Current can also be suppressed over an octave bandwidth or more by inserting a sequence of cable chokes in the feed line at quarter-wave intervals as shown in Fig. 11. A cable choke may consist of coaxial cable wound on a helical or toroidal core to form a high impedance reactor. A VHF cable choke, for example, can provide 500 ohms or more reactance between 30 to 90 MHz.

The current distribution on the feed line with a sequence of cable chokes inserted is shown in Fig. 12. The parasitic currents are essentially eliminated by this technique. In effect, the chokes act as a band-elimination filter for the rf currents allowing the antenna to operate as if the feed line were not present. In the MININEC model, the chokes are treated as point load reactors connected in series with the "wire" representing the feed line. The wire diameter equals that of the coaxial cable transmission line outer conductor. This example illustrates the loaded wire option available in MININEC.

IV. REVISIONS TO MININEC

Since its publication in 1982, MININEC has been revised several times to incorporate enhancements generated at NOSC and by other users. The known revisions and enhancements are listed in Fig. 13. The version of MININEC contained in the book (Artech House, Inc.) incorporates all of these revisions except for the modification allowing wires with a single pulse (unknown).

V. CONCLUSION

MININEC has proven to be a very versatile and widely accepted MOM code. The supplemental information and several examples provided here should be helpful to new users of MININEC.

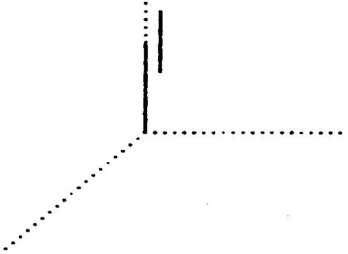


Figure 9.

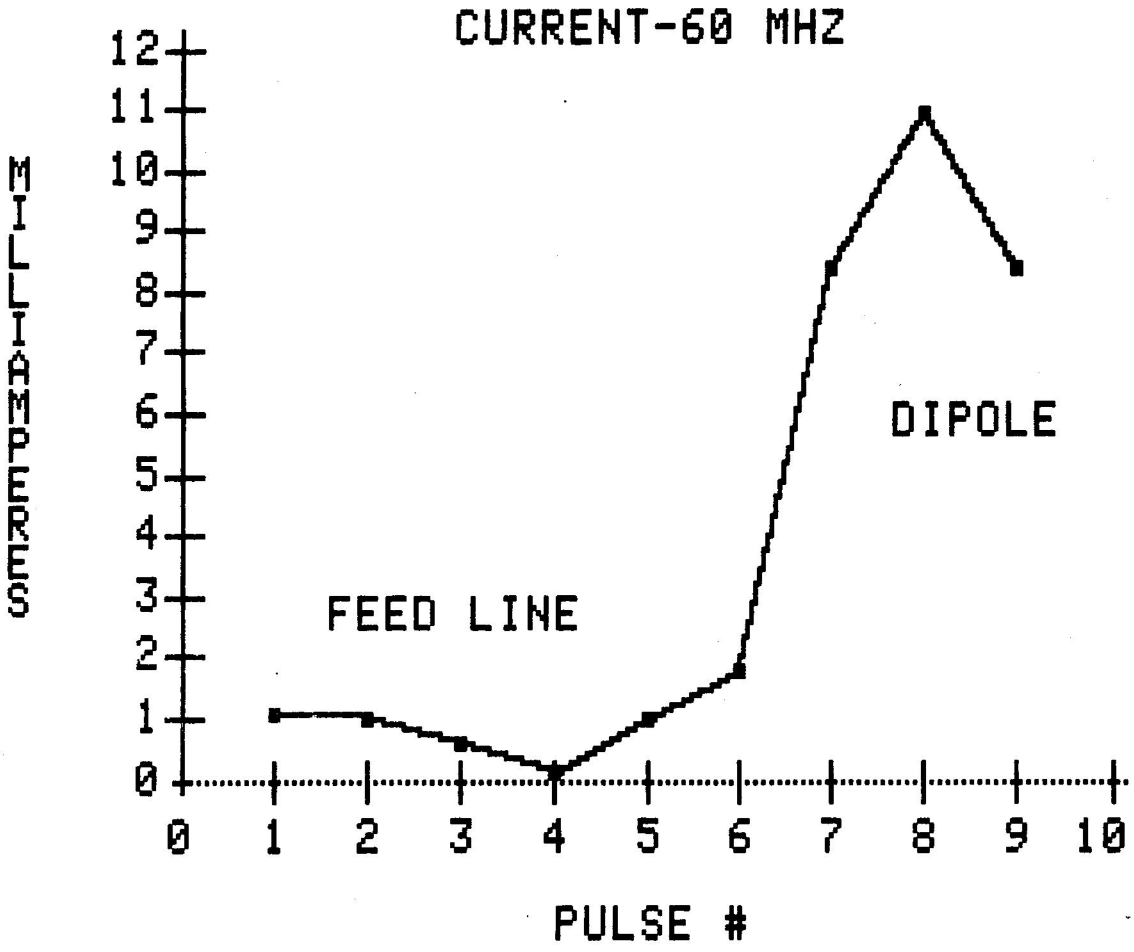


Figure 10.

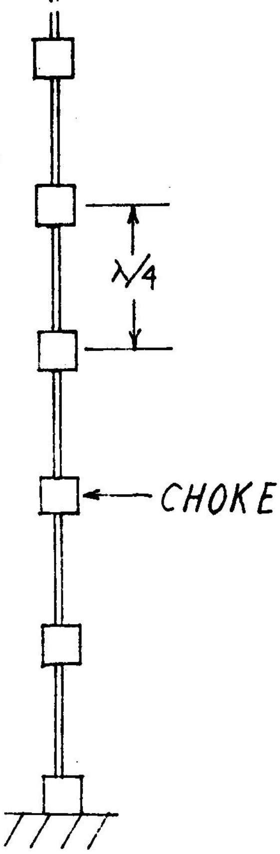


Figure 11.

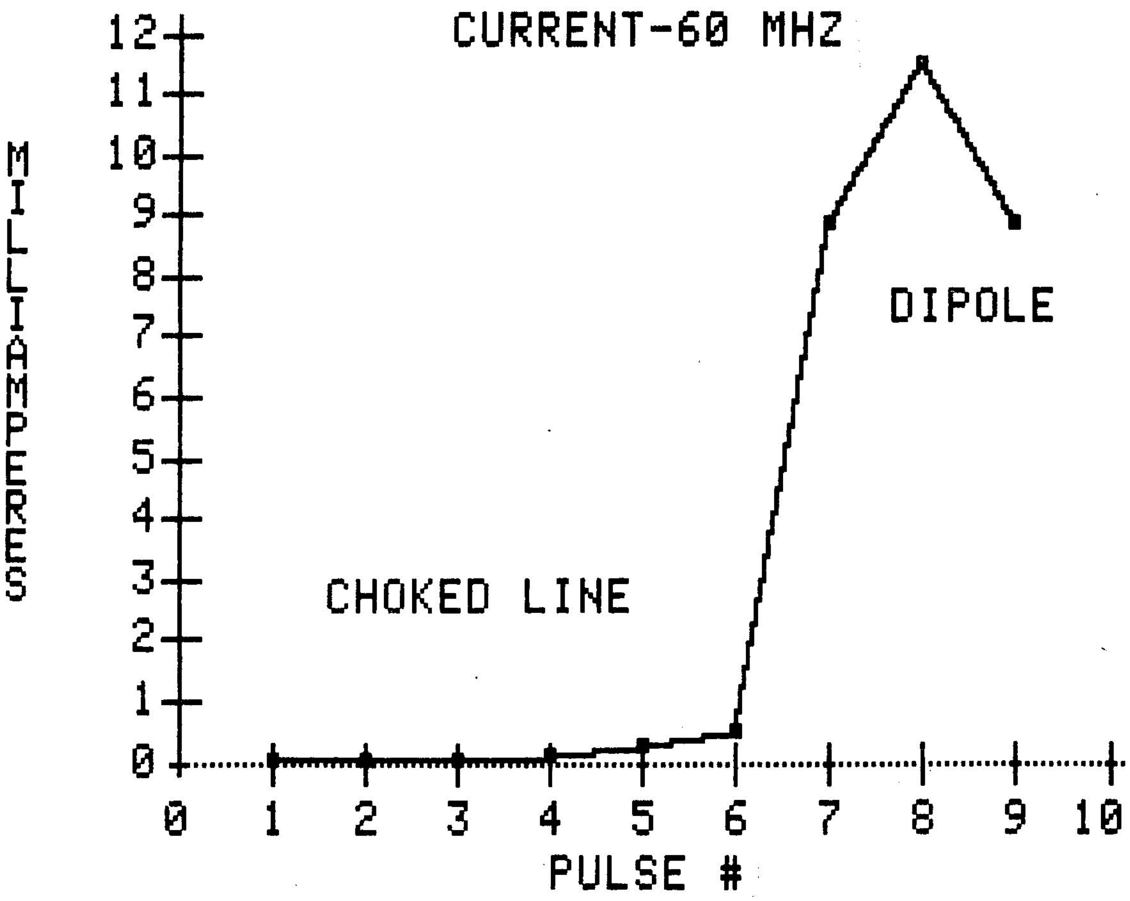


Figure 12.

MININEC

DOCUMENT	DATE	TITLE	DESCRIPTION
TD516	9/6/82	MININEC: A Mini-Numerical Electromagnetics Code	Original Tech Report & Code
MEMO	3/14/83	Enhancement of MININEC Capabilities	Wires inserting ground plane any angle
MEMO	3/17/83	Lumped-Parameter Impedance Loading Option in MININEC	a. Load and feed collocated base of monopole b. Corrects last constant in
MEMO	4/25/83	Corrections in MININEC	Corrects pattern calculation
TN-EMC-84-02	3/12/84	"Bugs" in Geometry Section of MININEC	Modification allowing wires w single pulse

Figure 13.

MODIFICATION TO MININEC FOR THE ANALYSIS OF WIRE ANTENNAS WITH SMALL RADIUS TO WAVELENGTH RATIOS

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MININEC is a useful and compact method of moments antenna program, but MININEC does not give reasonable values for the input reactance of very thin wires at low frequencies. This problem greatly restricts the use of MININEC in the design and analysis of vlf and lf antennas. A modification to the program which eliminates this restriction is discussed. The modification consists of treating both the source and observation segments as filaments and only considering the wire radius when computing the self-impedance. A listing of the changed computer code is included.

I. INTRODUCTION

MININEC is a method of moments microcomputer program, written in BASIC, that analyzes thin-wire antennas [Julian et al., 1982]. This compact code is based on a modified Galerkin procedure that was described by D. R. Wilton to solve an integral equation for the electric field [Li et al., 1983; Wilton, 1981]. With proper modeling, MININEC can solve accurately for the current and impedance on most arbitrarily oriented wires.

However, the input reactance given by MININEC for an electrically short, thin monopole begins to diverge from the expected value for wire radii less than approximately $10^{-5}\lambda$, where λ is the wavelength. This limitation in the program is evident in Fig. 1, which displays the input reactance X_a calculated by MININEC and the expected values for a 90.5 m monopole at 150 kHz for $10^{-7} < a/\lambda < 10^{-4}$ where a is the wire radius. The expected value is the input reactance for a short vertical radiator given by the equation

$$X_a = -Z_o \cot \frac{2\pi h}{\lambda} \text{ ohms} \quad . \tag{1}$$

The characteristic impedance Z_o is given by

$$Z_o = 60 \left[\ln \left(\frac{h}{a}\right) - 1\right] \text{ ohms} \quad , \tag{2}$$

which fits experimentally measured values, where h is the height of the radiator [Jasik, 1961]. The above expression is approximately the same as the more complicated expression for the input reactance given by the induced emf method, using a sinusoidal current distribution [Jordan and Balmain, 1968]. The exact a/λ value where MININEC is no longer valid depends on the particular microcomputer, whether double precision variables are being used, and the number of segments chosen. This limitation prevents MININEC from being used to design vlf and lf antennas that use wires with small radius to wavelength ratios.

A modification to the program has been developed which replaces the code for integral psi with code that treats the current as a filament on the wire axis. This change has resulted in reactances that differ by less than 1 percent from the values given by Eq. (1) and that are in excellent agreement with the experimental values for both a monopole and a top-loaded monopole. This is a fix which allows MININEC to be used to analyze vlf and lf antennas.

II. MODIFICATION AND VERIFICATION

MININEC is based on the assumptions that the wire radius is very small in comparison to a wavelength and that the radius is small with respect to the segment length so that there will be no azimuthal component of the current [Julian et al.]. Evidently, from Fig. 1, the assumption on a/λ is overpowered by a computational error for values less than approximately 10^{-5}.

In order to determine the vector and scalar potentials from a current carrying wire, MININEC evaluates an integral psi given by

$$\psi_{m,u,v} = \int_{s_u}^{s_v} k(s_m - s')ds' \quad , \tag{3}$$

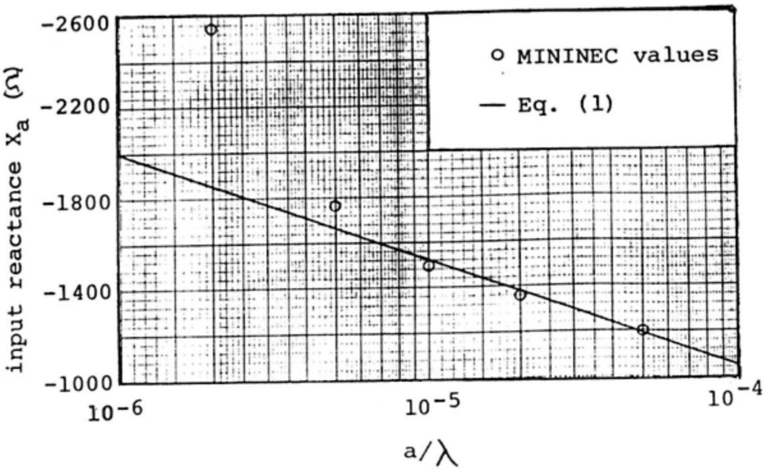

Figure 1. Input reactances from MININEC for a 90.5 m monopole divided into five segments at 150 kHz.

where

$$k(s_m - s') = \frac{1}{2\pi} \int_{-\pi}^{\pi} \frac{e^{-jkr_m}}{r_m} \, d\phi \quad , \tag{4}$$

$$r_m = [(s_m - s')^2 + 4a^2 \sin^2 \frac{\phi}{2}]^{1/2} \quad , \tag{5}$$

s_m is the observation point, s_v and s_u are the upper and lower end points, respectively, of the source segment and a is the wire radius. Equation (3) results in an elliptical integral of the first kind due to a singularity at $r_m = 0$. Since the wires are very thin in terms of wavelength when MININEC becomes unreliable, the current and charge densities are approximated by filaments of current and charge on the wire axis following Harrington. Thus, the double integral is simplified to a single integral [Harrington, 1981]

$$\psi(m,n) = \frac{1}{4\pi\Delta l_n} \int_{\Delta l_n} \frac{e^{-jkR_m}}{R_m} \, dl \quad , \tag{6}$$

where the distance between the source segment and observation point is

$$R_m = \begin{cases} [\rho_m^{\,2} + (z-z_m)^2]^{1/2} & m \neq n \\ (a^2 + z^2)^{1/2} & m = n \end{cases} \quad , \tag{7}$$

Δl_n is the length of the source segment, n is the center of source segment, m is the observation point, ρ_m is the horizontal distance between m and n, z_m is the vertical distance between z and m, and z is the vertical coordinate along the source. The integral can be approximated by expanding e^{-jkR_m} with a Maclaurin series to two terms. For m=n, this approximation yields

$$\psi(m,n) \approx \frac{1}{2\pi\Delta l_n} \ln \left(\frac{\Delta l_n}{a}\right) - j \frac{k}{4\pi} \quad , \tag{8}$$

For m≠n, use the crudest approximation with R_m constant so that

$$\psi(m,n) \approx \frac{e^{-jkR_m}}{4\pi R_m} \quad .$$

This simplifying modification was incorporated into MININEC by changes to the two subroutines in lines 20-890 [Li et al., 1983]. Primarily, most of the elliptical integration subroutine was deleted (specifically lines 70-220). The remaining integral is still performed numerically with the Gaussian quadrature subroutine. In addition, other statements had to be changed in order to adapt this modification into the program without changing the variables. These changes included the deletion of the variable I6 in lines 650 and 860, and the square of the wire radius A2 is no longer necessary in line 730. Also, lines 281-288 were inserted to treat the m=n case, including when a half segment is being calculated, and two lines (60 and 785) were added to direct the program to the proper lines based on the value of distance D. The modified subroutines of an IBP Personal Computer version of MININEC are listed in the Appendix.

Figure 2 displays the results of the modified MININEC along with the same expected values of input reactance. The agreement is within 1 percent for $10^{-7} < a/\lambda < 10^{-4}$.

This modified MININEC was also used to determine the input impedance at 50 kHz of a 192 m top-loaded monopole with six radial wires. The tower was divided into five segments, and the radial wires were each divided into three segments. The tower had a radius of 0.48 m and the six top radials each had a radius of 0.0127 m ($a/\lambda \approx 2 \times 10^{-6}$) which gave an input reactance of 0.621 - j625 ohms, or within 3 percent of the value obtained from experimental data (0.636 - j609 ohms) [Devaney et al., 1966].

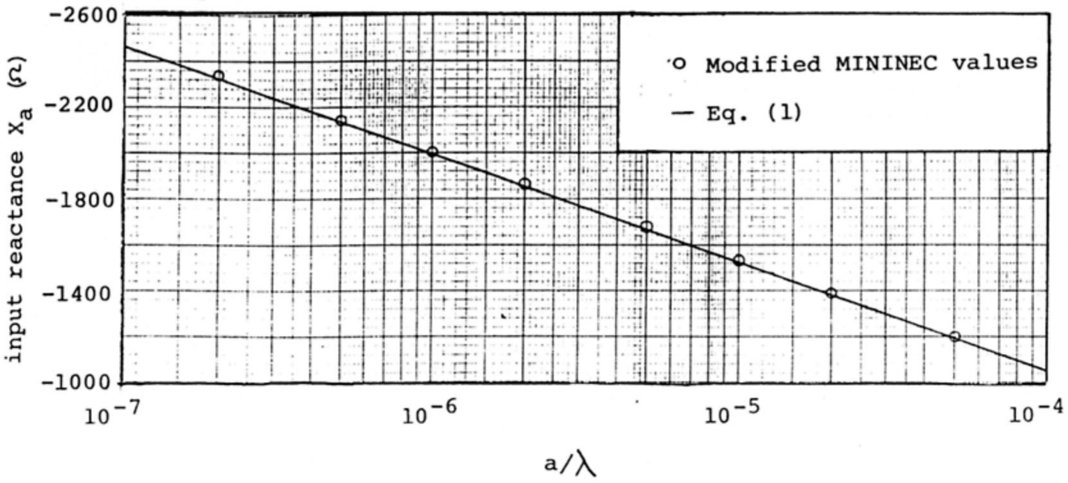

Figure 2. Input reactances from modified MININEC compared to expected values for a 90.5 m monopole divided into five segments at 150 kHz.

III. CONCLUSIONS

A quick modification to MININEC has been presented that allows the program to analyze vlf and lf antennas. The adapted program yields reasonable values for the input reactance of wires with very small radius to wavelength ratios, without introducing significant error in the resistance values. This alteration could be implemented as an option in the program instead of replacing the valid code for larger wires. Also, the modification could be included in a change that allows for antennas with both relatively large and small radius wires to be analyzed.

IV. ACKNOWLEDGMENTS

The authors would like to thank J. C. Logan of the Naval Ocean Systems Center and A. W. Glisson of the University of Mississippi for helpful conversations. We would also like to thank J. C. Logan for providing us a copy of MININEC for the IBM PC.

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APPENDIX: LISTING OF MODIFIED MININEC SUBROUTINES

```
20 X3=T*D(P4)
30 Y3=T*F(P4)
40 Z3=F*T*G(P4)*F0
50 D=SQR(D0-F3*(X3*X2+Y3*Y2+Z3*Z2)+X3*X3+Y3*Y3+Z3*Z3)
60 IF D<1 THEN RETURN
230 B1=D*W
240 T3=T3+COS(B1)/D
250 T4=T4-SIN(B1)/D
260 RETURN
270 F1=1
280 F0=1
281 IF K=1 AND P3-P2=1 AND P1=(P3+P2)/2 THEN GOTO 282 ELSE GOTO 285
282 T1=2*LOG(S(P4)/A(P4))
283 T2=-(W*S(P4))
284 GOTO 880
285 IF I=J AND K=1 AND P3-P2=.5 AND 1-P1/INT(P1)=0  THEN GOTO 286 ELSE GOTO 290
286 T1=LOG(S(P4)/A(P4))
287 T2=-(W*S(P4)/2)
288 GOTO 880
290 IF C(J,1)=-C(J,2) THEN F0=SGN(C(J,F1))
300 F3=2*SGN(C(J,F1))
310 IF P1-INT(P1)=H0 GOTO 440
320 I4=INT(P1+1)
330 I5=INT(P1)
340 X1=(X(I4)+X(I5))/2
350 Y1=(Y(I4)+Y(I5))/2
360 Z1=(Z(I4)+Z(I5))/2
370 X2=X1-X(P2)
380 Y2=Y1-Y(P2)
390 Z2=Z1-Z(P2)*F
400 X3=X1-X(P3)
410 Y3=Y1-Y(P3)
420 Z3=Z1-Z(P3)*F
430 GOTO 530
440 I4=INT(P2+1)
450 IF P2-INT(P2)=H0 THEN I4=P2
460 I5=INT(P2)
470 X2=X(P1)-(X(I4)+X(I5))/2
480 Y2=Y(P1)-(Y(I4)+Y(I5))/2
490 Z2=Z(P1)-F*(Z(I4)+Z(I5))/2
500 X3=X(P1)-X(INT(P3))
510 Y3=Y(P1)-Y(INT(P3))
520 Z3=Z(P1)-F*Z(INT(P3))
530 D0=X2*X2+Y2*Y2+Z2*Z2
540 D3=X3*X3+Y3*Y3+Z3*Z3
550 S4=(P3-P2)*S(P4)
560 F2=1
570 N3=7
580 T1=H0
590 T2=T1
600 T=S(P4)+.001*A(P4)
610 A2=A(P4)*A(P4)
620 IF SQR(D3)+SQR(D0)>T GOTO 680
630 I6=16*A(P4)/S(P4)
640 IF P1-INT(P1)>0 THEN F2=2
660 IF P1-INT(P1)>0 THEN S4=S(P4)/2
670 GOTO 720
680 T=SQR(D0)/ABS(S4)
690 IF T>3 THEN N3=3
700 IF T>5 THEN N3=1
710 I6=H0
720 I5=N3*2
730 D0=D0
740 L=N3
750 T3=H0
760 T4=T3
770 T=S4*(Q(L)+.5)
780 GOSUB 20
785 IF D<1 THEN GOTO 282
790 T=S4*(.5-Q(L))
800 GOSUB 20
810 L=L+1
820 T1=T1+Q(L)*T3
830 T2=T2+Q(L)*T4
840 L=L+1
850 IF L<I5 GOTO 750
860 T1=F2*T1*S4
870 T2=F2*T2*S4
880 RETURN
```

DREDGING-DEPOSITION, A TWO LAYER GROUND PROBLEM IN MF-RADIO WAVE PROPAGATION

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A limited area in close proximity to an MF broadcasting station was foreseen for deposition of dredging material. The influence that the lossy material with finite height might have on MF propagation was studied.

I. INTRODUCTION

The subject that is presented here is classified to the topic: Ground Interface Effects.

Figure 1 illustrates the problem: A limited area in close proximity to a medium wave broadcasting station was foreseen for deposition of dredging material.

System 1	represents a VHF concrete tower of about 300m,
System 2	is a two mast MF aerial for omnidirectional propagation by day and with a directional pattern at night,
System 3	is an active MF installation in parallel operation.

The limited area marked out with the solid line was foreseen for deposit of dredging material. The effects of this deposition on the performance of the MF antennas are studied.

II. PRELIMINARY CONSIDERATIONS

Several questions arise for the deposition effects on the MF-antenna performance:

1. How does the deposition affect the MF propagation?
2. Which permissible deviations of the MF propagation characteristics are definable?
3. What shape of the deposition should be preferred?
4. What is the minimum permissible distance to the MF antennas?
5. What is the maximum permissible height of the deposition?

The foregoing is an example where the NEC code could help. The two medium ground approximation (cliff model) was the starting option for using the code.

Figure 2 shows the well known view of this cliff model. For this model the parameters for a second

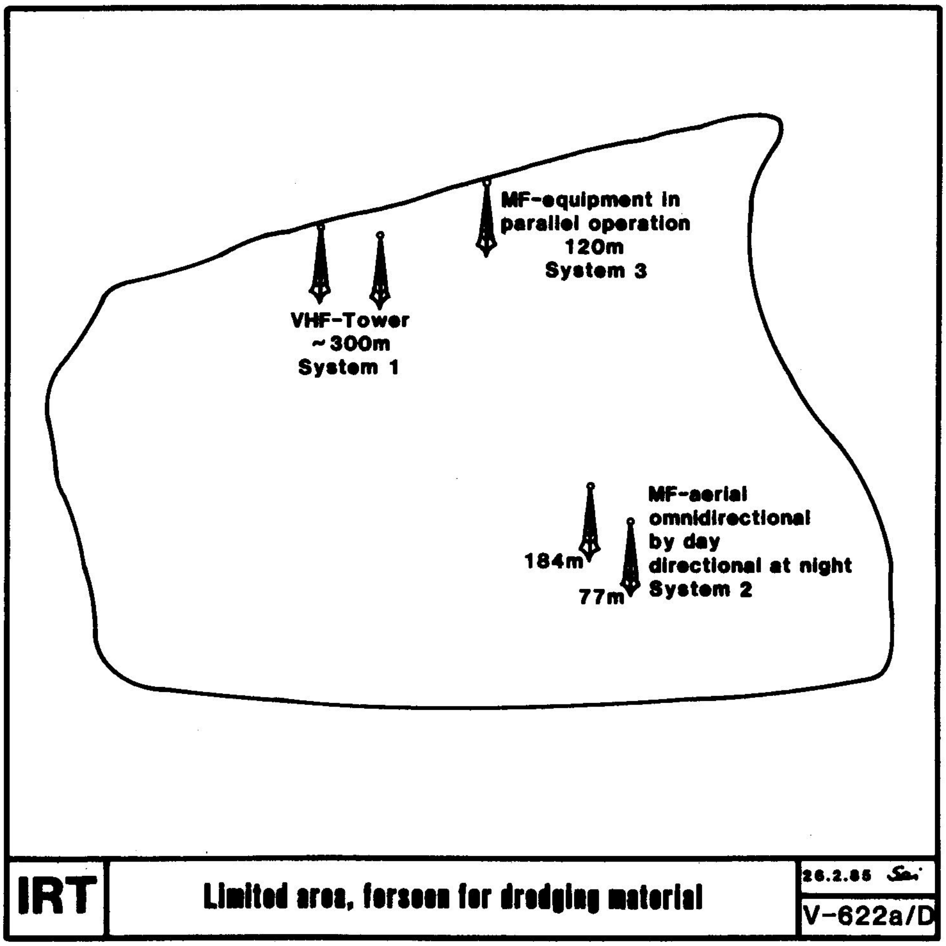


Figure 1.

ground medium are fixed by the GD card (s. NOSC, Techn. Doc. 116, Vol. 2). The restrictions to this approximate model are:

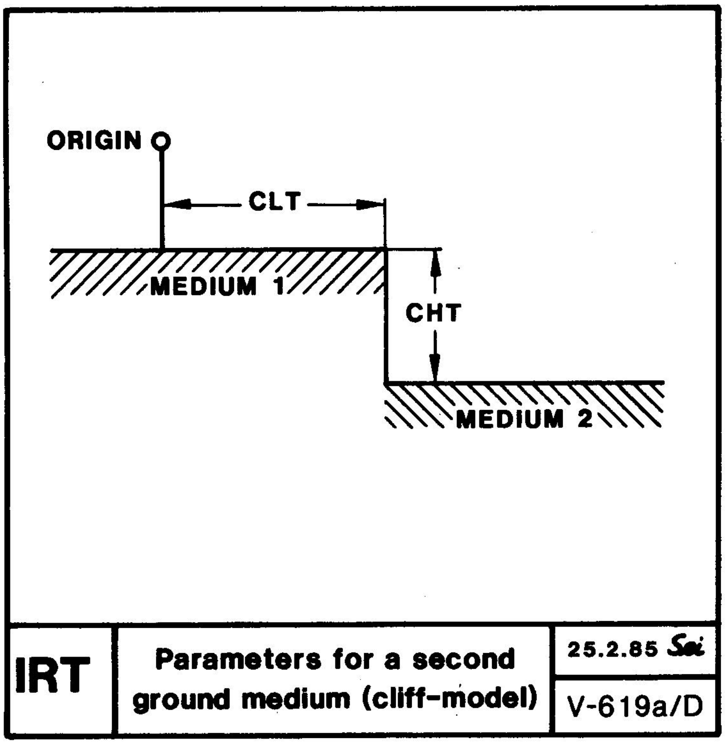


Figure 2.

1. The use of the GD card does not require recalculations of the matrix,
2. the ground parameters of a second medium are only effective beyond the immediate vicinity of the antenna,
3. the parameters for the second medium are used only in the far field calculation,
4. no edge diffraction by the cliff is taken into account.

In mind of this, the cliff model is modified for the deposition effects on the performance of the MF antennas in the following way (s. Fig. 3):

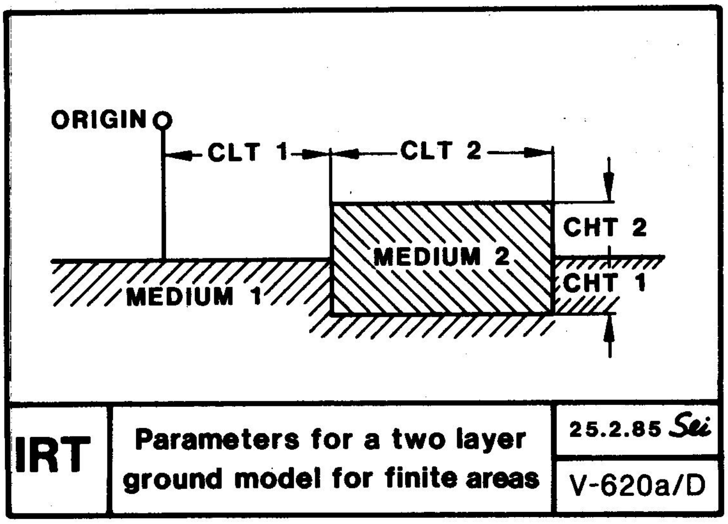


Figure 3.

For the stratified (two layer) region, fixed by CLT1, CLT2 and CHT1, CHT2 the effect of stratification on the reflection of waves is indicated in a simple way according to Wait /1/.

III. TWO LAYER CLIFF MODEL

According to Wait the surface impedance Z_1 for any angle of incidence is represented by the relation:

$$Z_1 \cong Q \cdot K_1$$

with

$$Q = \frac{(\gamma_1/\gamma_2) + \tanh \gamma_1 h_1}{1 + (\gamma_1/\gamma_2) \cdot \tanh \gamma_1 h_1} \qquad \text{for } \mu_1 = \mu_2 = \mu$$

and

$\gamma_m = (i\mu\omega(\sigma_m + i\varepsilon_m\omega))^{1/2}$ = propagation const.

Z_1 = surface impedance at the air-ground interface

K_1 = characteristic impedance of the upper layer

Q = correction to K_1 to account for the presence of the lower layer

For radio frequencies and moderate conducting ground ($\sigma \sim 10^{-2}$ S/m) a suitable formula is given by:

$$Q = \frac{(\sigma_1/\sigma_2)^{1/2} + \tanh\left(\sqrt{i} \cdot V\right)}{1 + (\sigma_1/\sigma_2)^{1/2} \tanh\left(\sqrt{i} \cdot V\right)}$$

where

$$V = (\sigma_1 \mu\omega)^{1/2} h_1 \text{ with } h_2 \text{ ---> } \infty$$

This function Q and its argument q are plotted as a function of V for various values of the ratio (σ_1/σ_2) in Fig. 4.

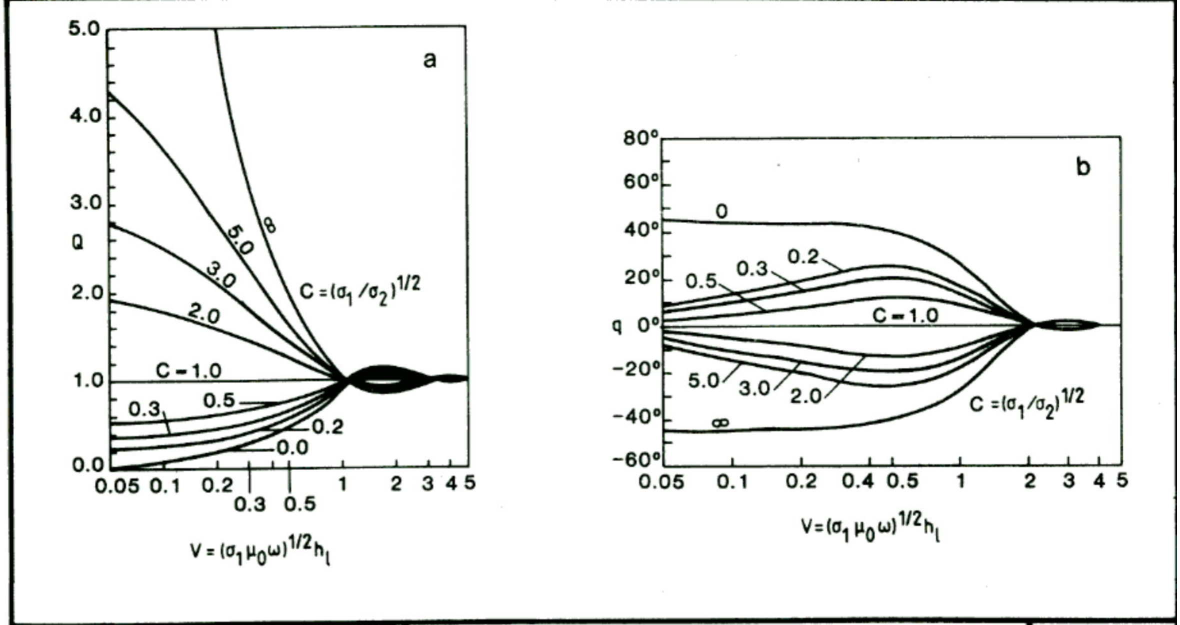
IV. NEC MODIFICATIONS

A subroutine called QZRQ was inserted into the NEC code and permits the determination of this Q factor to correct the surface impedance ZRATI2 in the common block /GND/ of the code. In Fig. 5 a part of the modified linkage chart is shown. Therefore, the contribution of each image segment as it is modified by the reflection coefficient for cliff problems will be taken into account by this correction factor. The values of Z (ZRATI2 in the code) are assigned in subroutine RDPAT, and in the subroutine FFLD the reflection coefficients are calculated for ground areas fixed by the new two layer ground model.

V. RESULTS

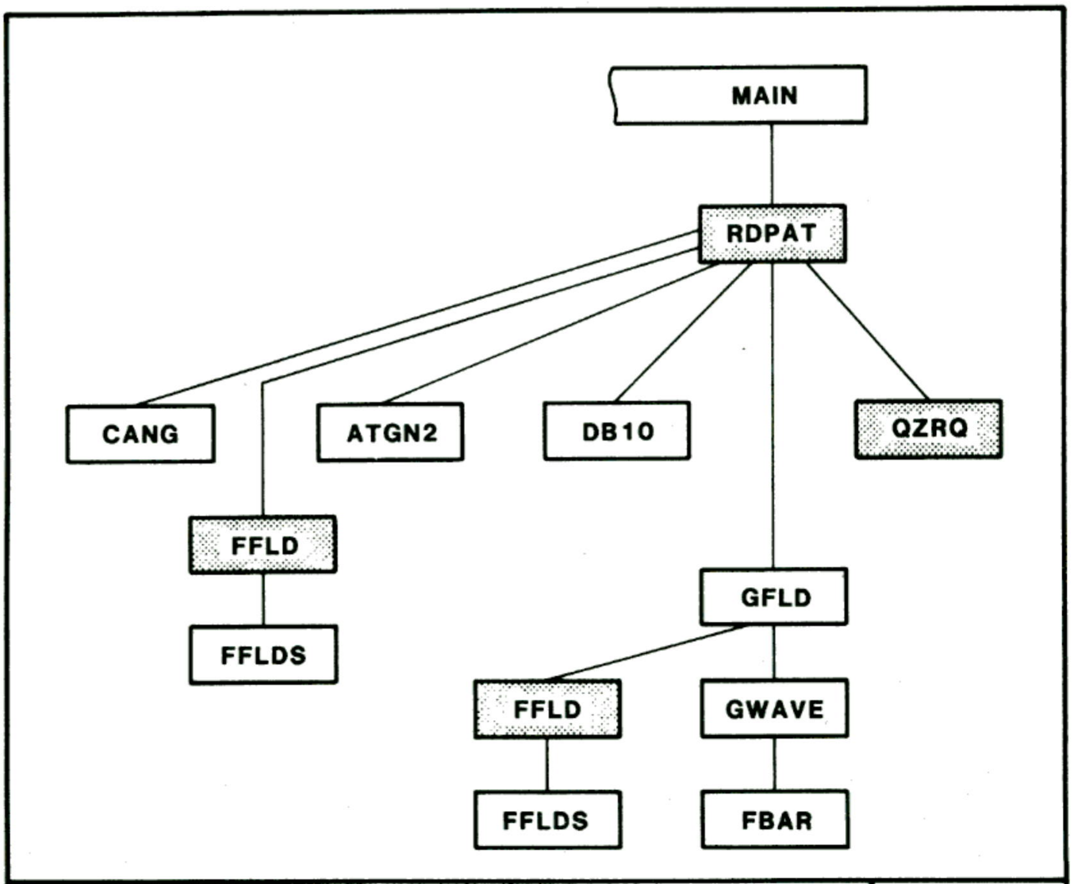
Detailed pattern calculations for the three MF systems were made. Various shapes of the deposition and different values for the conductivity of the dredging material were taken into account.

As an example the vertical and horizontal patterns of a 0.6λ monopole (system 2 by day) affected by the deposition are shown in Figures 6 and 8. The thick line represents the pattern without additional dredging material. The shape of the deposition is toroidal with abrupt walls. The conductivity of the upper layer, the muddy material, is changed. Fixed parameters are the dimensions of CH, CLT1 and CLT2. As can be seen the vertical pattern is strongly affected by interaction with the deposition.



IRT | **Correction factor for a two-layer ground after Wait.**
a-Amplitude b-Phase
27.2.85 Sei
V-624a/D

Figure 4.



IRT | **NEC-Subroutine linkage chart (continued) after NOSC TD116, Vol1**
25.2.85 Sei
V-621a/D

Figure 5.

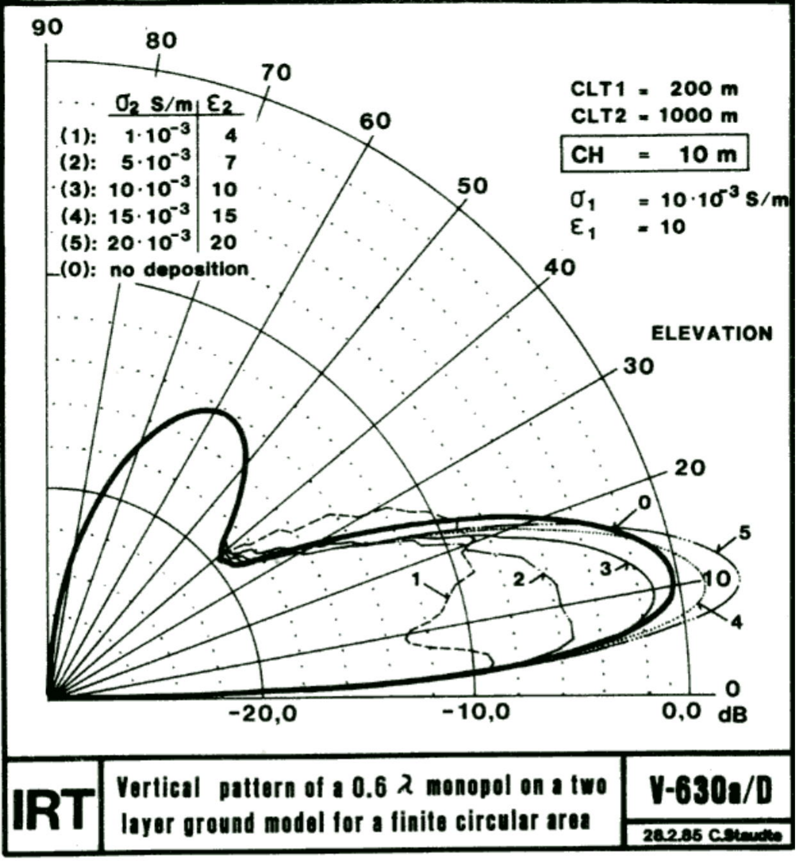


Figure 6a.

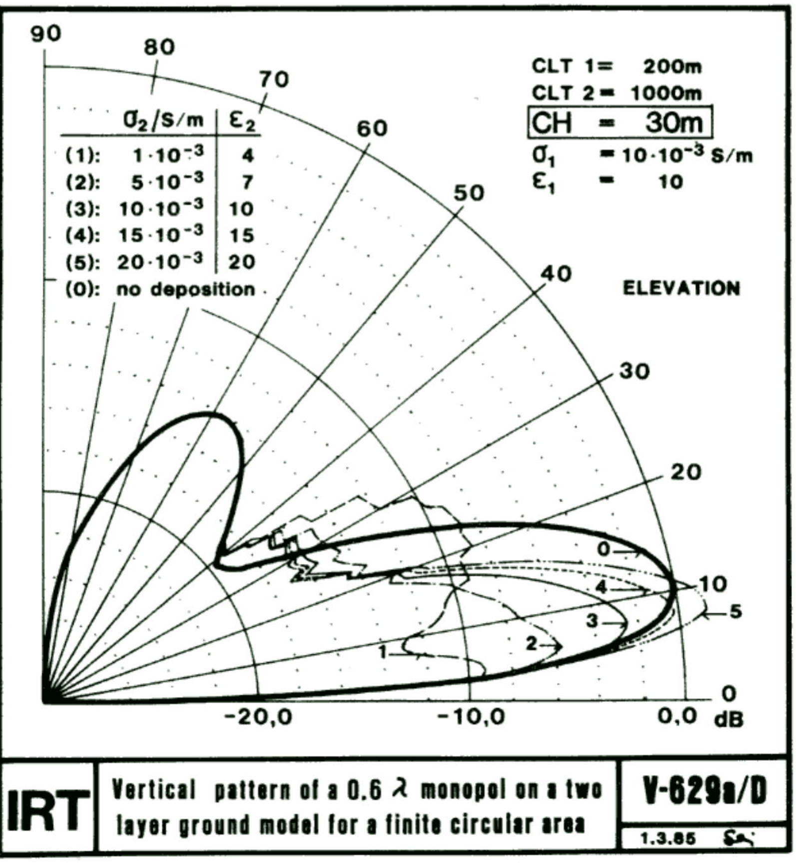


Figure 6b.

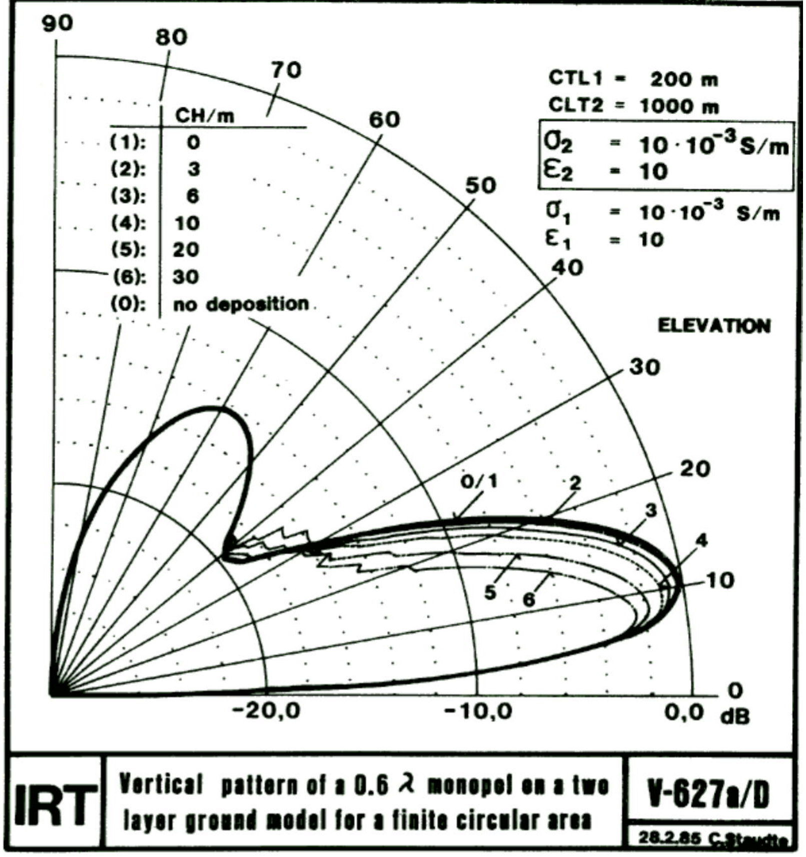


Figure 6c.

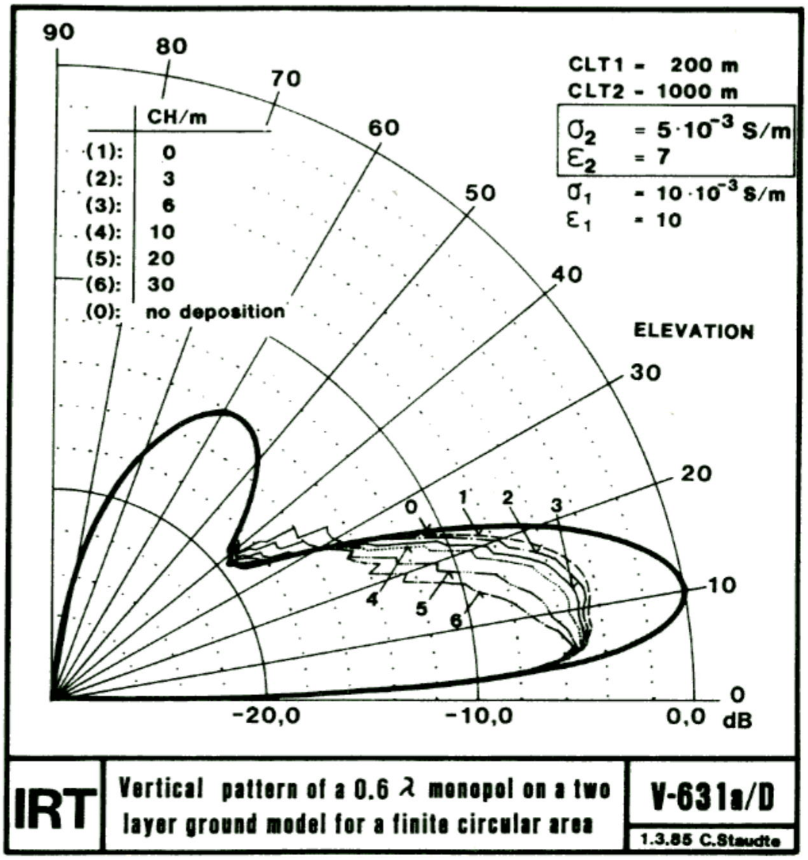


Figure 6d.

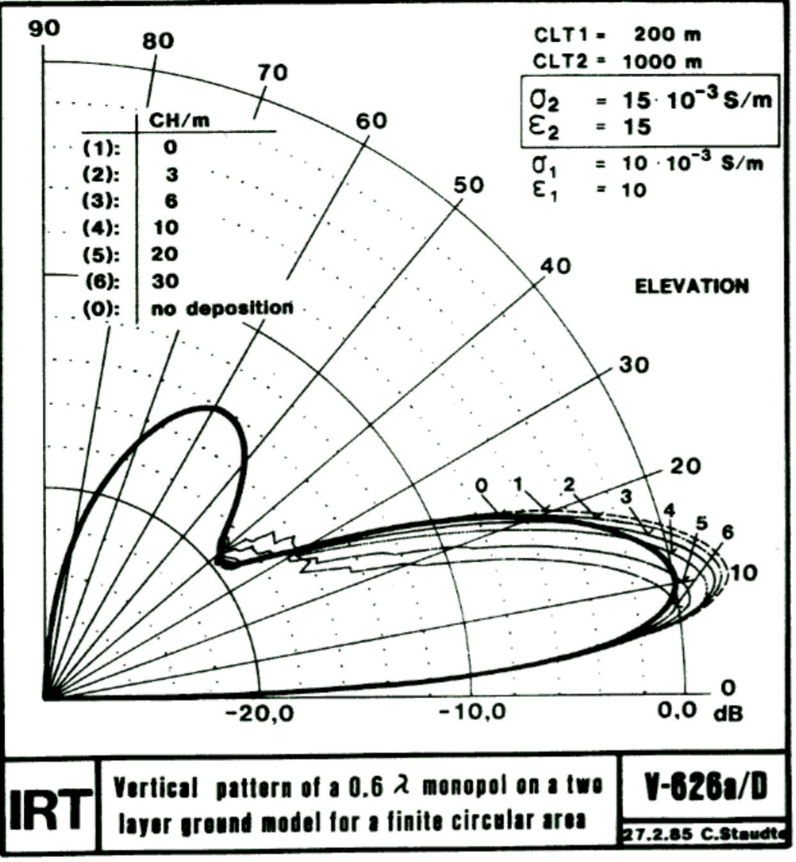


Figure 6e.

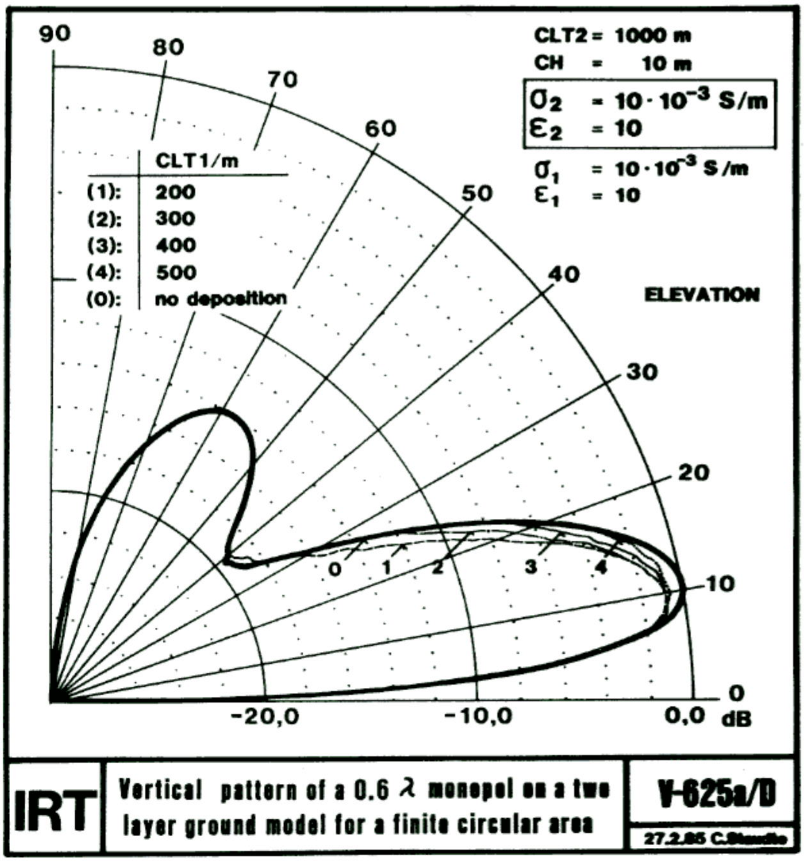


Figure 6f.

More or less, the effect caused by the deposition of dredging material changes the main beam of the vertical pattern.

To calculate the horizontal pattern (surface pattern) the limited area was sectorized. In Fig. 7 this sectorization of the limited area is shown. Since the surface wave is not included in the reflection coefficient approximation in NEC the horizontal pattern was computed at a small angle above the horizon to obtain a nonzero space wave.

In Fig. 8a and 8b the horizontal pattern as a function of the sectorized area for various conductivities of the dredging material and different values of CH are shown. The deviation of circularity should be noted.

VI. CONCLUSION

One main purpose of this study is to get permissible deviations of the MF propagation characteristics by dredging deposition effects.

From the pattern calculation pointed out, limiting values for such deviation can be derived:

1. the reduction of field strength should not yield more than 1 dB,
2. the elevation angle of the direction of maximum radiation should not change by more than 1 degree,
3. the deviation from circularity should not be more than 1 db.

With these permissible deviations in mind a deposition of dredging material on the limited area is tolerated only for conductivities of the dredging material close to or better than those of the lower ground. Furthermore, a distance of about 1λ should be maintained between the antennas and the nearest edge of the deposition. Finally the height of the deposition should not be more than 10 m.

With only small modifications of the NEC code the problem of affecting MF propagation by dredging deposition could be solved to our full satisfaction.

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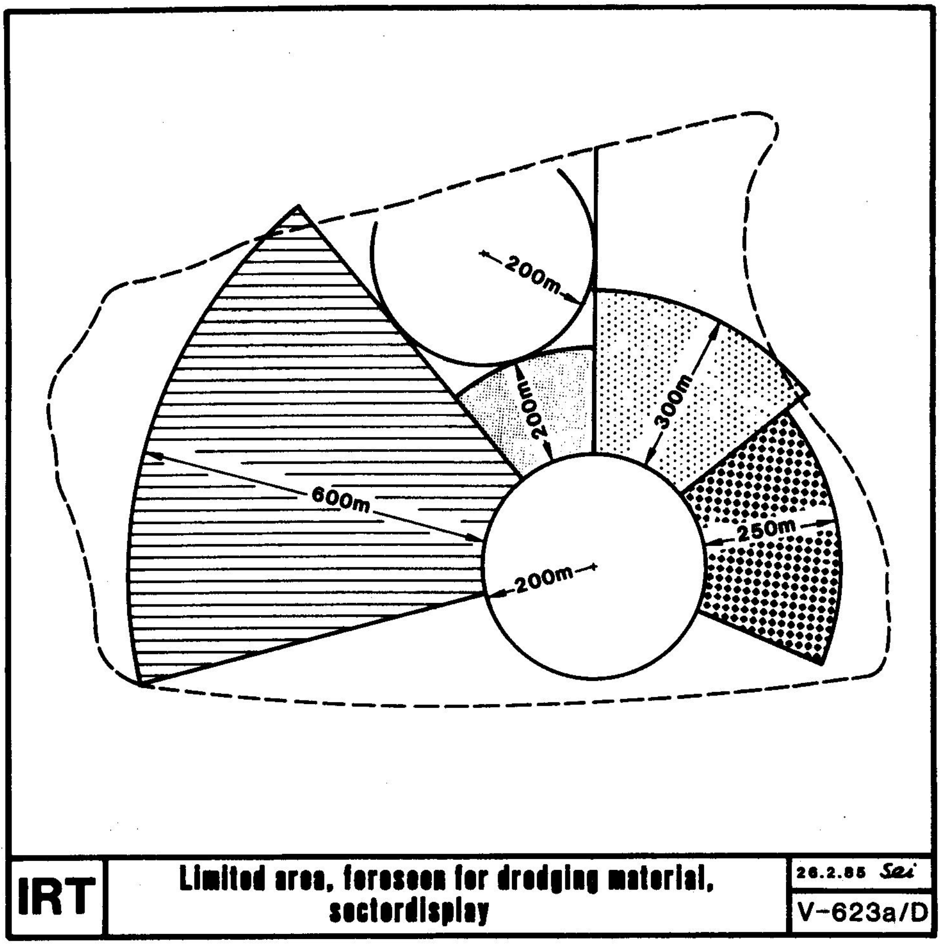
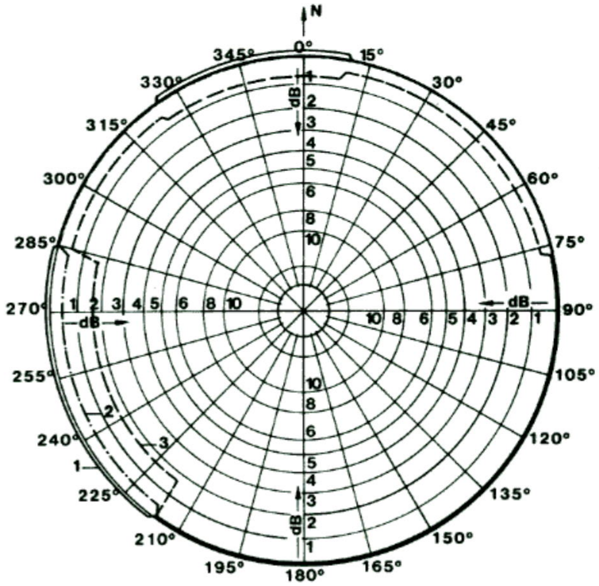


Figure 7.



CH-10m

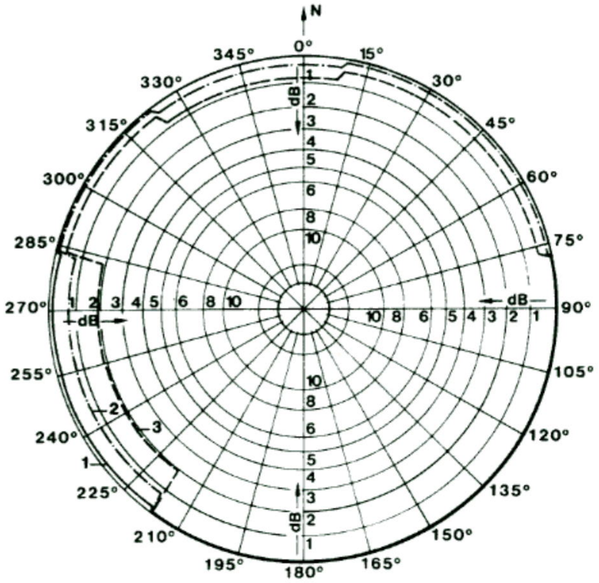
$\sigma_1 = 10 \cdot 10^{-3}$ S/m

$\varepsilon_1 = 10$

	σ_2/S/m	ε_2
(1):	$15 \cdot 10^{-3}$	15
(2):	$10 \cdot 10^{-3}$	10
(3):	$5 \cdot 10^{-3}$	7

IRT	Horizontal pattern of a 0.6λ monopol on a two layer ground model for finite sectorial areas	28.2.85 Sei V-633a/D

Figure 8a.



CH- 30m

$\sigma_1 = 10 \cdot 10^{-3}$ S/m

$\varepsilon_1 = 10$

	σ_2/S/m	ε_2
(1):	$15 \cdot 10^{-3}$	15
(2):	$10 \cdot 10^{-3}$	10
(3):	$5 \cdot 10^{-3}$	7

IRT	Horizontal pattern of a 0.6λ monopol on a two layer ground model for finite sectorial areas	28.2.85 Sei V-632a/D

Figure 8b.

BROADCAST ANTENNA OPTIMIZATION WITH MICROCOMPUTERS

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INTRODUCTION

Increasingly intense competition for the limited frequency allocation in the AM broadcast band (550-1600 kHz) has led to the use of directional array antennas to reduce interference. In this band, all available channels have been assigned, and for a new station to receive a frequency assignment they must guarantee no interference with existing stations. Thus, new stations must be located a significant distance away from any existing station on that frequency, and often additional protection must be provided by using directional array antennas. In the U.S., more than 25% of the AM broadcast antennas are directional arrays. This allows considerably more use of each channel and is perhaps the first attempt at practical "frequency reuse" [1].

Typically, the directional antenna design is based on a coverage and protection specification. Typical specifications cover the following:

(a) Radiation is maximized over the sector covering the city of license;

(b) Radiation is minimized at headings where distant locations are to be protected.

The level of protection (null depth) varies with frequency, distance, and power level of the local transmitter and transmitter to be protected. The required width of the nulls depends upon propagation conditions and geometry of the region to be protected.

For daytime coverage, only ground wave propagation is considered and the required nulls are all at 0° elevation.

For night time, the existence of sky wave propagation forces elevation pattern nulls at various azimuths. This usually results in designs combining both quarter wave and taller towers.

In the past, standard practice has been to calculate the array excitation voltages, mutual impedance, and pattern on the basis of a sinusoidal current distribution on each element, and superposition of the field from each element individually [2]. A problem results from this because the currents are not sinusoidal. Consequently, the actual mutual impedance is not as predicted, and the actual field for each element (with the others open circuited) must account for scattering from the other elements. This scattering arises from the fact that a current distribution can exist on a tower even though the input is open circuited. The sinusoidal current distribution assumption does not allow this. In general, the taller the towers the more pronounced will be the difference between actual results and those calculated on the basis of sinusoidal currents. In fact, determining the actual excitations to give a desired pattern can be extremely difficult since they may be considerably different than the calculated factors [3].

There now exists a computer program, Mini-Numerical Electromagnetics Code (MININEC) [4], which can be run on various microcomputers that gives acceptably accurate calculations of antenna characteristics, including actual current distribution, self- and mutual impedance and far fields. A procedure for the optimization of the feed excitation voltages of a broadcast array antenna to yield a given antenna radiation pattern is described. The procedure makes use of MININEC [4] and can be used on a microcomputer.

MININEC

The MINI-NUMERICAL ELECTROMAGNETICS CODE is a frequency domain Method of Moments computer code for the analysis of wire antennas. MININEC is written in the BASIC language for use on microcomputers with as little as 64k memory. The MININEC program is based on the numerical solution of an integral representation of the electric fields. A modified Galerkin procedure is used to solve the integral equation. This formulation results in a uniquely compact (i.e., requiring little core computer storage) code suitable for implementation on a microcomputer. The MININEC code solves for the impedance and currents on arbitrarily oriented wires in free space and over a flat perfectly conducting ground plane. Configurations with multiple wire junctions can also be used. Options include lumped parameter impedance loading of the wires and calculation of far-field patterns.

The solution to an antenna problem generated by a thin wire method of moments computer program is at best an approximation. Nonetheless, highly accurate answers can be obtained by careful modeling of the antenna configuration, taking into account the inherent limitations of the computer code.

POWER GAIN

The broadcast antenna specifications are given i terms of power gain or directive gain in db. In th direction (θ,ϕ)

$$G = 4\pi\frac{P(\theta,\phi)}{P_{IN}} \tag{1}$$

where $P(\theta,\phi)$ is the power radiated per unit soli angle in the given direction and P_{IN} is the tota power accepted by the antenna.

$$P_{IN} = \frac{1}{2} R_e(V \cdot I^*) \tag{2}$$

where V is the applied source voltage and I* is th conjugate of the resulting feed point current.

$$P(\theta,\phi) = \frac{R^2}{n} \bar{E} \cdot \bar{E}^* \tag{3}$$

where R is the observation sphere radius, $\bar{E}$ is the fa electric field, and n is the intrinsic impedance of free space.

The only difference between directive and power gain is that P_{IN} is replaced by P_{rad}, where

$$P_{rad} = P_{IN} - P_{loss} \tag{4}$$

and P_{loss} is calculated as the power loss in the antenna system. The gain in db is expressed

$$G_{dB} = 10 \log G \qquad (5)$$

For the purpose of feed point optimization, the multiple feeds of the antenna are considered to be an N-port network, where N is the number of feeds. The relationship between the port voltages and currents can be expressed

$$I_i = \sum_{j=1}^{N} Y_{ij} V_j \qquad (6)$$

where Y_{ij} are the traditional short-circuit admittance parameters. The Y-parameters are defined and calculated under short-circuit conditions at either the input or the output port [5].

MININEC is used to calculate the field caused by having one element excited, with the others open circuited. NEC calculates the currents flowing on all elements, and from these currents the field at a distant point. The total field pattern of the antenna is equal to the sum of the fields due to the individual elements excited. Changing the magnitude and phase of the individual element excitation voltages amounts to multiplying the fields from each individual element (with the others open circuited) by the appropriate amplitude and phase prior to summing.

If $ei(\theta,\phi)$ is the far electric field due to the implied voltage of $V_i = 1 + \theta j$, the total far electric field is

$$\bar{E}(\theta,\phi) = \sum_{i=1}^{N} V_i e_i(\theta,\phi) \qquad (7)$$

where V_i would be the actual feed point voltages impressed on the antenna. Substituting equations (2), (3), (6), and (7) into equation (1)

$$G(\theta,\phi) = \frac{4\pi R^2 \sum_{i=1}^{N} V_i e_i(\theta,\phi) \cdot \sum_{i=1}^{N} V_i e_i(\theta,\phi)^*}{n \sum_{i=1}^{N} R_E \; V_i \sum_{j=1}^{N} Y_{ij} V_j^*} \qquad (8)$$

Thus power gain can be calculated for any set of feed point voltages if $ei(\theta,\phi)$ and the short-circuit admittance parameters, Y_{ij}, are known. These values can be calculated using MININEC.

OPTIMIZATION

For the purpose of optimizing the excitation voltages of an array antenna system to provide for a desired radiation pattern, the following error function is defined:

$$\text{Error} = \sum_{j=1}^{M} W_j(\theta,\phi)[R_j(\theta,\phi) - G_j(\theta,\phi)]^2 \qquad (9)$$

where M is the number of pattern points used to define the required radiation pattern, $Rj(\theta,\phi)$ is the required radiation pattern at point j, and $Gj(\theta,\phi)$ is the achieved antenna pattern given by equation (8). The error is defined over the entire synthesis range. The reason for the squares is to ensure that, when the error is a minimum, each component of the sum is also a minimum. Therefore, the best approximation to $Rj(\theta,\phi)$ by $Gj(\theta,\phi)$ is achieved in the mean. $Wj(\theta,\phi)$ is a weighting function that allows the synthesis precision to be changed over the synthesis range.

The objective of optimization is to minimize the error function of equation (9). A BASIC program, called OPTIMAL, was written to perform this optimization using the NEWTON-RAPHSON method [6]. The NEWTON-RAPHSON method is the most widely used of all iteration formulas. Note that the error function of equation (9) is a function of 2N parameters. There is both a real and an imaginary voltage for each of the feed points. Using v to denote each of these 2N parameters, each successive iteration of all 2N parameters is given by

$$V_{n+1} = V_n - \frac{\text{Error}(V_n)}{\frac{d[\text{Error}(V_n)]}{dV_n}} \qquad (10)$$

Analytical expressions have been obtained for each of the required derivatives and are used in OPTIMAL.

SUMMARY

Using the programs MININEC and OPTIMAL, a microcomputer can be used to optimize the antenna system feed points for a desired radiation pattern. Example calculations are given. The procedure demonstrates the growing importance of the microcomputer in antenna modeling.

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MODELING MONOPOLES ON RADIAL-WIRE GROUND SCREENS

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October 2, 1985

A special version of NEC-3 (NEC-GS) has been developed to model a monopole on a uniform radial-wire ground screen. By taking full advantage of the physical symmetry and excitation symmetry, NEC-GS can model large radial-wire screens in much less cpu time and storage than would be required by NEC-3. Little experimental data is available for comparison, however, so validation has relied on self consistency checks such as convergence and the balance of input power with radiated power. Also, a comparison with an approximation developed by Wait and Pope (1954) has helped to validate both the NEC results and the approximation.

One question in the NEC model was the treatment of the junction of the monopole and the radial wires, where many wires meet with narrow angles of separation. The current at the junction is forced to satisfy Kirchhoff's current law while the derivative of the current at the junction on the monopole above ground (I'_+) and the radials below ground (I'_-) is made to satisfy the condition

$$\frac{I'_+}{I'_-} = \frac{1}{\epsilon_g}$$

where

$$\epsilon_g = \epsilon_- - j\,\frac{\sigma_-}{\omega\epsilon_0} .$$

This condition on I' was derived for a vertical wire penetrating the interface and its application to the ground screen with a junction at the interface could be questioned.

The most difficulty, however, has been encountered in modeling ground screens floating above the ground. In this case, since all wires are in air, the derivatives of current are forced to be equal for equal wire radii. Apparently, this condition needs refinement when the radials are closely coupled to the ground while the monopole is effectively shielded from the ground. To obtain converged results for ground screens at small height ($\sim 10^{-5}\lambda_0$) above ground, it was necessary to use segment lengths at the junction on the order of the height of the screen above ground, tapering to larger segments away from the junction. This was particularly important for electrically small monopoles and screens (Logan, 1984).

Results for buried ground screens are generally found to be stable with varying segment lengths. For dense, buried ground screens the segments can often be considerably longer than recommended for a wire in the ground since, due to the shielding effect of the screen, the current behaves more like that on a wire in air.

As a check on the solution for a buried ground screen, the computed input impedance of a monopole was compared with that predicted by an approximation developed by Wait and Pope (1954). For this approximation, a current $I_0 \cos(kz)$ is assumed on the monopole. The difference in input impedance (ΔZ) between that of the monopole on the radial-wire screen on real earth and that of the monopole on an perfectly conducting ground is obtained, application of the compensation theorem, integral over the surface. The magnetic fi the ground screen is taken to be that on the ground and the electric field is related to surface impedance for the screen and grou derivation of this approximation and computed are included in Wait (1969).

This approximation should be accurate screen radius in wavelengths (b/λ) somewhat than δ where

$$\delta = |\epsilon_g|^{-1/2} .$$

Also, the number of wires in the groun must be sufficiently large for the surface i approximation to be valid.

The ΔZ computed by the NEC-GS is compa that from the approximation in Figs. 1 through input impedance computed by NEC-GS for the mon an infinite perfectly conducting ground wa j22.2 ohms. Agreement in ΔZ is generally b/λ greater than δ. The larger discrepancie radials may be due to inaccuracy of the impedance approximation for a sparse screen. low conductivity ground of Fig. 3, the discr are due to standing waves on the screen wir are not taken into account by the approximatic comparison seems to support the accuracy of th results since the agreement is best for a dens and highly conducting ground for whi approximation should be most accurate and the of-moments solution more difficult.

On screens above ground, large standing exist for any range of earth conductivity. surface impedance approximation is not accu screens above ground.

Results obtained with NEC-GS show that i a useful tool for modeling radial wire screens. Caution is called for in appl however, until it is more fully validated. S small radius are a particular problem s surface impedance approximation fails as a ch

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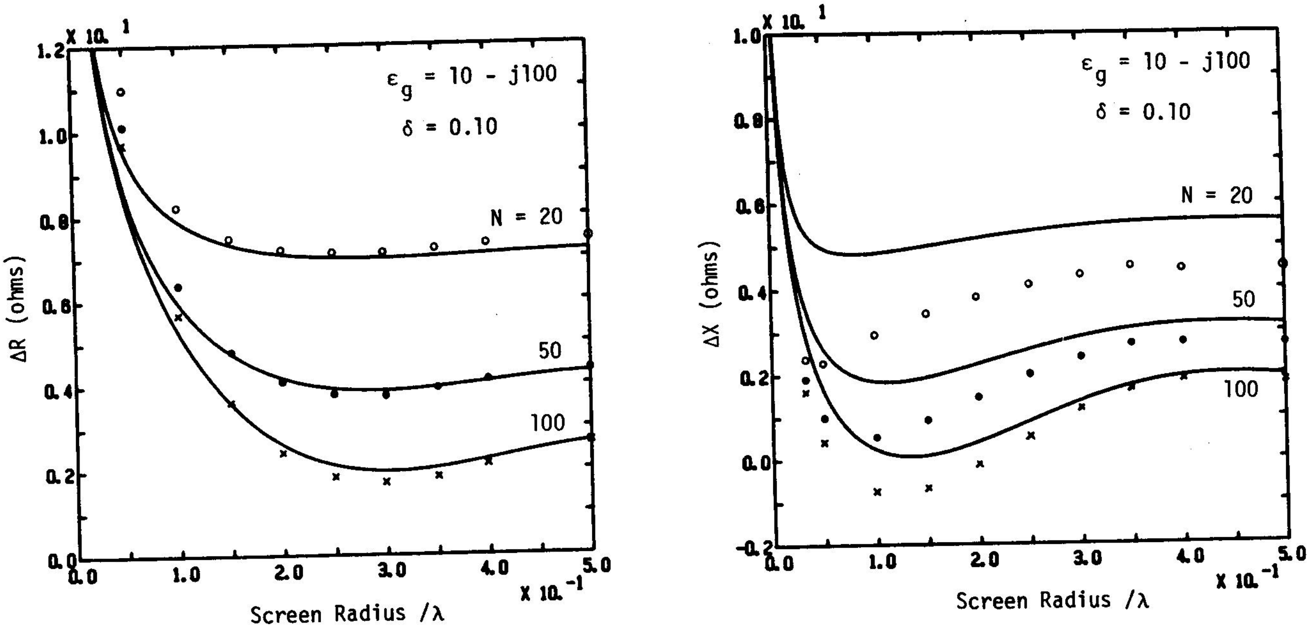

Figure 1. Impedance change (ΔR + jΔX) for a 0.25 λ monopole on a ground screen of N radial wires with ε_g = 10 - j100. Symbols (o • x) represent the NEC solution for 20, 50, and 100 radials, respectively. Solid lines are from the approximation of Wait and Pope (1954).

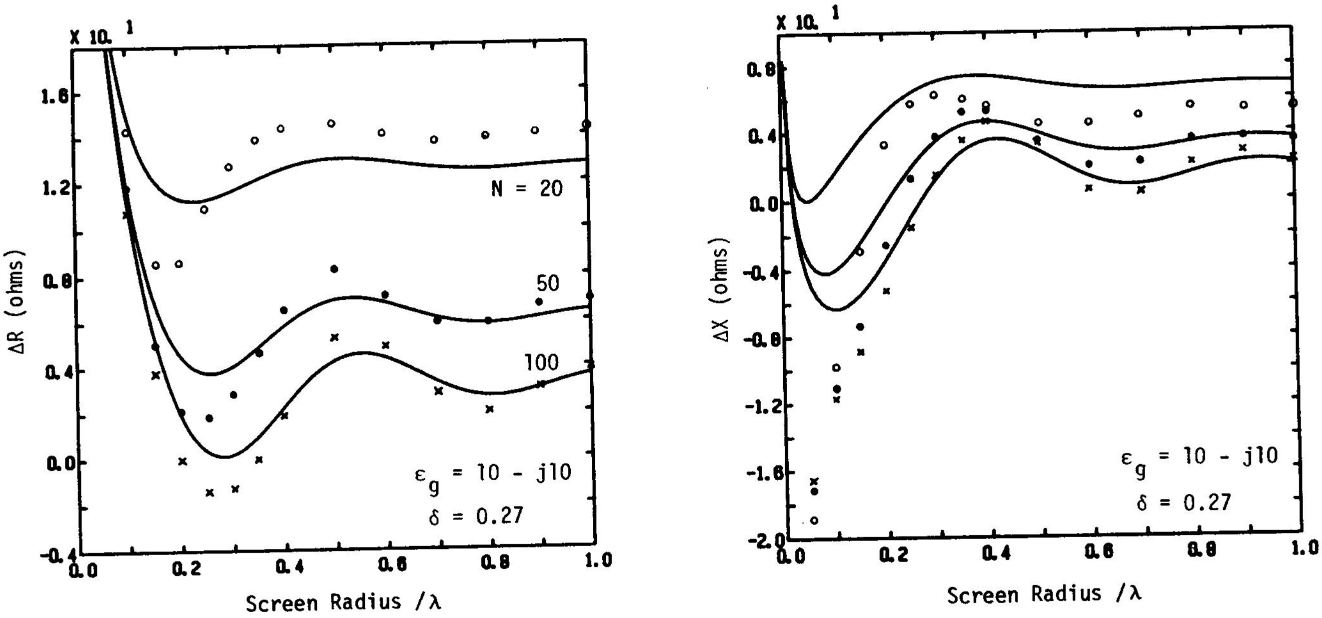

Figure 2. Impedance change (ΔR + jΔX) for a 0.25 λ monopole on a ground screen of N radial wires with ε_g = 10 - j10.

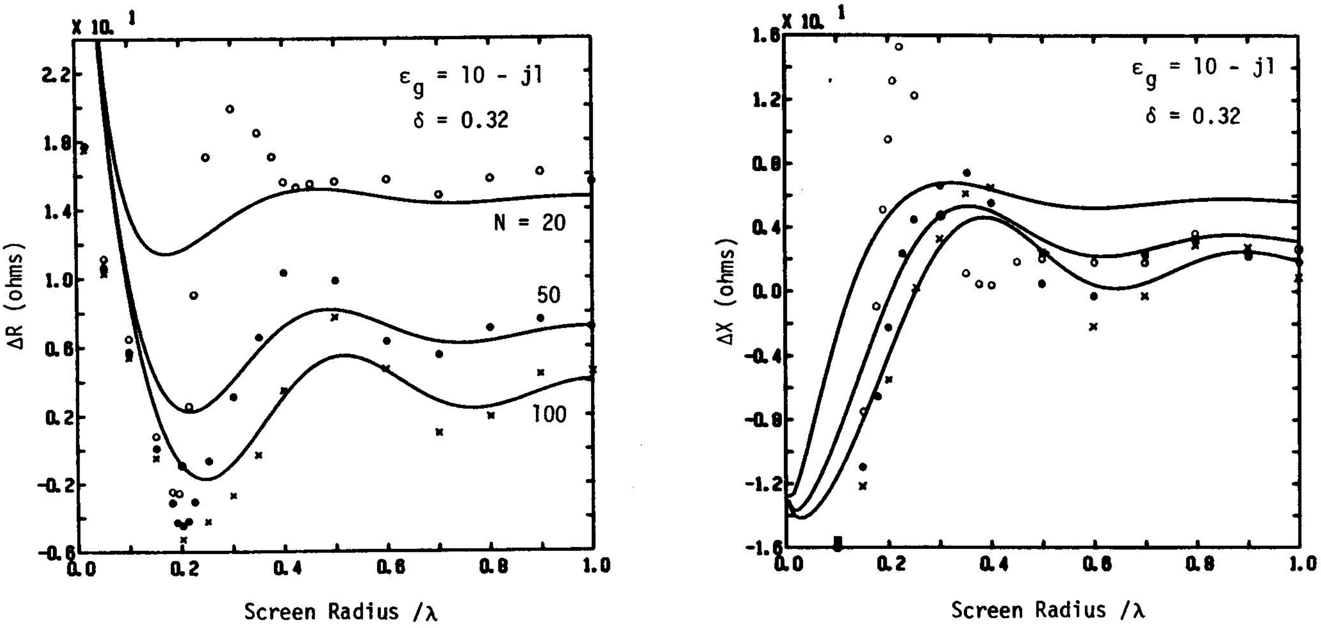

Figure 3. Impedance change ($\Delta R + j\Delta X$) for a 0.25 λ monopole on a ground screen of N radial wires with $\varepsilon_g = 10 - j1$. With low ground conductivity,[8] a large standing wave can exist on the radials that is included in the NEC solution but not in the surface impedance approximation.

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