

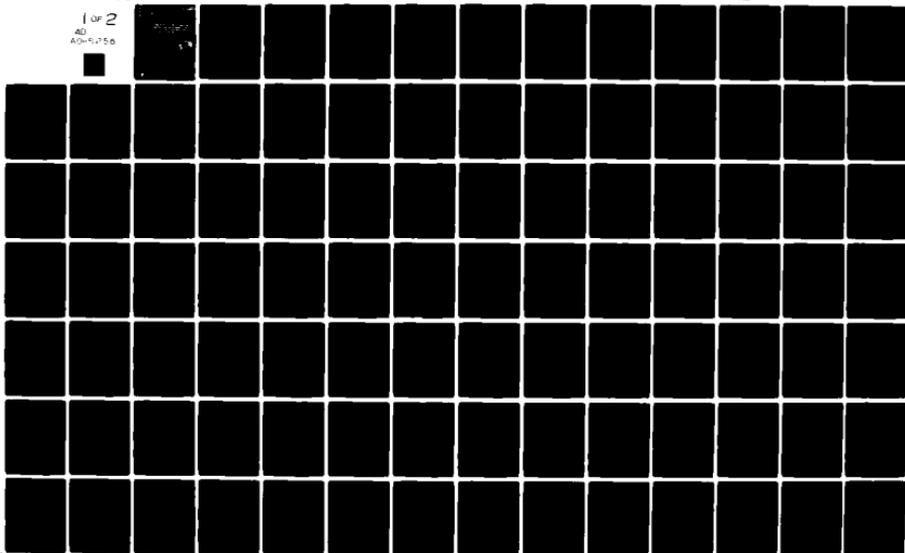
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REPORT R-856 AUGUST, 1979

UIIU-ENG 78-2249

CSL COORDINATED SCIENCE LABORATORY

**SOME MODIFICATIONS TO NEWTON'S
METHOD FOR THE DETERMINATION
OF THE STEADY STATE RESPONSE
OF NONLINEAR CIRCUITS**

FRANCIS BENEDICT BROSE, JR.

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REPORT DOCUMENTATION PAGE		READ INSTRUCTIONS BEFORE COMPLETING FORM
1. REPORT NUMBER	2. GOVT ACCESSION NO. AD-A085256	3. RECIPIENT'S CATALOG NUMBER
4. TITLE (and Subtitle) SOME MODIFICATIONS TO NEWTON'S METHOD FOR THE DETERMINATION OF THE STEADY STATE RESPONSE OF NONLINEAR CIRCUITS		5. TYPE OF REPORT & PERIOD COVERED Technical Report
7. AUTHOR(s) Francis Benedict Grosz, Jr		6. PERFORMING ORG. REPORT NUMBER R-856, UIIU-ENG-78-2249
9. PERFORMING ORGANIZATION NAME AND ADDRESS Coordinated Science Laboratory University of Illinois at Urbana-Champaign Urbana, Illinois 61801		8. CONTRACT OR GRANT NUMBER(s) DAAB 07-72-C-0259, DAAB 07-72-C-0010 N00014-79-C-0424
11. CONTROLLING OFFICE NAME AND ADDRESS Joint Services Electronics Program		10. PROGRAM ELEMENT, PROJECT, TASK AREA & WORK UNIT NUMBERS
14. MONITORING AGENCY NAME & ADDRESS (if different from Controlling Office)		12. REPORT DATE August, 1979
12. 139		13. NUMBER OF PAGES 132
15. SECURITY CLASS. (of this report) UNCLASSIFIED		15a. DECLASSIFICATION/DOWNGRADING SCHEDULE
16. DISTRIBUTION STATEMENT (of this Report) Approved for public release; distribution unlimited.		
17. DISTRIBUTION STATEMENT (of the abstract entered in Block 20, if different from Report)		
18. SUPPLEMENTARY NOTES		
19. KEY WORDS (Continue on reverse side if necessary and identify by block number) Nonlinear Circuits Computer-Aided Analysis Steady State Oscillatory Response		
20. ABSTRACT (Continue on reverse side if necessary and identify by block number) Many methods have become available recently for the determination of the steady state response of nonlinear circuits. The Newton method of Trick, Aprille and Colon is one of the more widely available methods. However, in certain cases, the method fails to converge to a steady state solution. Three modifications to the Newton method to increase convergence reliability or speed are proposed. The implementation of these modifications is discussed. Several example circuits are given and the results compared. Some guidelines for the use of these modified methods are also given.		

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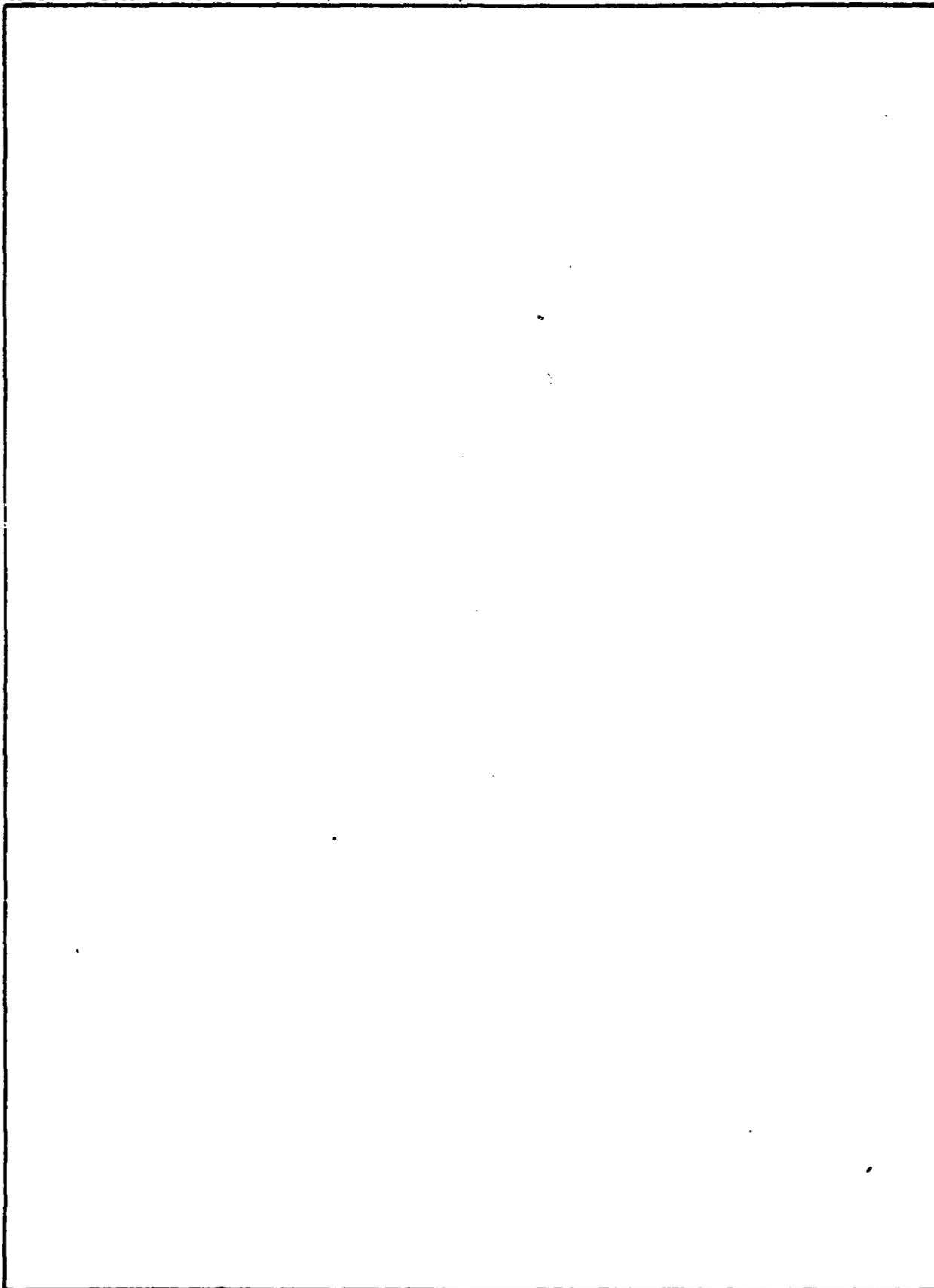
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This work was supported in part by the Joint Services Electronics Program (U.S. Army, U.S. Navy, and U.S. Air Force) under Contract DAAB-07-72-C-0259, DAAG-29-78-G-0016, and N00014-79-C-0424.

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BY

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B. S., Louisiana State University at New Orleans, 1970
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THESIS

Submitted in partial fulfillment of the requirements
for the degree of Doctor of Philosophy in Electrical Engineering
in the Graduate College of the
University of Illinois at Urbana-Champaign, 1979

Thesis Adviser: Professor T. N. Trick

Urbana, Illinois

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Department of Electrical Engineering
University of Illinois at Urbana-Champaign, 1979

Many methods have become available recently for the determination of the steady state response of nonlinear circuits. The Newton method of Trick, Aprille and Colon is one of the more widely available methods. However, in certain cases, the method fails to converge to a steady state solution. Three modifications to the Newton method to increase convergence reliability or speed are proposed. The implementation of these modifications is discussed. Several example circuits are given and the results compared. Some guidelines for the use of these modified methods are also given.

ACKNOWLEDGMENT

I wish to express my deep appreciation to my advisor, Professor Timothy N. Trick for his help and encouragement throughout the course of this research. I also wish to thank the members of my committee Professors M. E. VanValkenburg and W. R. Perkins for their support.

My special thanks also to my family and friends for their encouragement and support.

TABLE OF CONTENTS

Chapter	Page
1. INTRODUCTION	1
2. STEADY STATE ANALYSIS METHODS	3
2.1 FREQUENCY DOMAIN METHODS	3
2.2 TIME DOMAIN METHODS	10
3. SOME PROBLEMS ENCOUNTERED IN THE CLASSIC NEWTON METHOD AND POSSIBLE SOLUTIONS	23
3.1 THE CLASSIC NEWTON METHOD	23
3.2 HIGH-SENSITIVITY PROBLEMS	23
3.3 COLON'S DAMPED NEWTON METHOD	26
3.4 VECTOR DAMPED NEWTON METHOD	26
3.5 PROOF OF LOCAL CONVERGENCE	28
3.6 A MODIFIED NEWTON METHOD FOR AUTONOMOUS SYSTEMS	32
3.7 SUMMARY OF THE DAMPED NEWTON METHOD	33
3.8 AN EXTERNALLY-SUPPLIED DAMPING VECTOR AND STATE ELIMINATION	35
3.9 PROVISION OF CONTRACTION MAPPING PERIODS FOR THE DECAY OF FAST TRANSIENTS	35
4. THE SINC PROGRAM	37
4.1 INTRODUCTION TO SINC (SIMULATOR FOR INTEGRATED NONLINEAR CIRCUITS)	37
4.2 CHANGES AND CORRECTIONS TO SINC	38
4.3 NEW IMPLEMENTATIONS	43
4.4 CONTL-TIME DOMAIN ANALYSIS CONTROL ROUTINE	58
5. EXAMPLES	79
5.1 INTRODUCTION	79
5.2 LINEAR CIRCUIT WITH PARASITICS	80
5.3 POWER SUPPLY	83
5.4 COLON WIEN BRIDGE	86
5.5 COLPITTS OSCILLATOR	89
5.6 CLASS-C RF AMPLIFIER	92
5.7 CLASS-C 1GHz RF AMPLIFIER	92
5.8 TRANSISTOR SWITCHING POWER SUPPLY	96
5.9 ANALYSIS OF RESULTS	101
6. CONCLUSION	103
6.1 GUIDELINES FOR THE SELECTION OF MODIFIED NEWTON METHODS	103
6.2 FUTURE WORK	106
APPENDIX. SINC USER'S MANUAL	107
REFERENCES	128
VITA	132

Chapter I

INTRODUCTION

A major concern in the computer-aided design of electronic circuits is the determination of the steady state response of nonlinear oscillatory systems. Improvements in DC, AC, and transient analysis techniques have resulted in fast, reliable methods for these problems. Here we shall be concerned with improved methods for the determination of the steady state response of nonlinear oscillatory circuits.

One of the most useful of the many suggested steady state analysis methods is the Newton technique of Trick, Aprille, and Colon. This method appears to be becoming more widely accepted, but is not universally used. One possible reason is that the method does not always converge to a solution. It is difficult to gain acceptance for a solution that is not almost universally reliable. The problem was to study the application of the Newton method to the determination of the steady state response of a variety of nonlinear oscillatory circuits and to identify the causes of convergence failure in order to design a modification of the Newton method which eliminated the convergence problem.

In order to analyze a variety of nonlinear circuits the program "SINC" was chosen. SINC is a program for the DC and transient analysis of nonlinear circuits. It was modified by Fan to perform the Newton iterations for the determination of the steady state. The program was studied, certain errors were corrected, new devices and commands implemented, and eventually the modifications of the Newton method were installed and tested.

A variety of circuit examples were run using the proposed modifications to determine the usefulness of the new techniques. Comparisons were also made of the speed of operation for the different methods, and conclusions were drawn concerning their usefulness.

Chapter 2 is a review of the techniques currently available for the steady state analysis of nonlinear systems. Some discussion is also made of the advantages and disadvantages of each method.

Chapter 3 contains a detailed discussion of the Newton method as applied to circuits by Trick and Aprille, and of the modifications of Trick, Colon, and Fan. Some possible causes of convergence failure are identified and discussed. Modifications to the Newton method are proposed and the convergence behavior of the modified method discussed. The use of the modified methods in the case of autonomous circuits is also discussed.

Chapter 4 is an introduction to the SINC program and an explanation of the changes made to the current version. Some errors and changes are pointed out. New implementations are discussed, and the changes necessary to implement the modified Newton techniques developed in Chapter 3 are reviewed.

Chapter 5 contains several example circuits and compares the results obtained using these techniques on each example. A number of different types of circuits are included to provide a reasonable basis from which to determine the usefulness of these techniques.

Chapter 6 offers some general guidelines for the use of these techniques and some suggestions for future development.

An Appendix contains a users manual for those interested in using the SINC program to analyze other circuits.

Chapter 2

STEADY STATE ANALYSIS METHODS

In recent years, several techniques for the determination of the steady state response of a system have been proposed. In this chapter we present a review of the more prominent schemes. They are naturally divided into frequency domain and time domain methods.

2.1 FREQUENCY DOMAIN METHODS

2.1.1 The Harmonic Balance Method

The Harmonic Balance method of Bailey [1] and Lindenlaub [2] requires that the user choose a fundamental frequency and specify a set of harmonics which determine the response of the system. The method attempts to find a periodic solution based on these components.

We assume that the system equations can be written as

$$\dot{\underline{x}} = \underline{f}(\underline{x}, t) \quad (2.1)$$

which is periodic in t with period T , and where \underline{x} and \underline{f} are N -vectors. We also assume that the system (2.1) has a solution of period $T = \frac{2\pi}{\omega}$ which can be represented as

$$\underline{x}(t) = \underline{x}_0 + \sum_{k=1}^{\infty} \underline{x}_{1k} \cos k\omega t + \underline{x}_{2k} \sin k\omega t \quad (2.2)$$

If we truncate this series to M components,

$$\hat{\underline{x}}(t) = \hat{\underline{x}}_0 + \hat{\underline{x}}_{11} \cos \omega t + \hat{\underline{x}}_{21} \sin \omega t + \dots + \hat{\underline{x}}_{1M} \cos M\omega t + \hat{\underline{x}}_{2M} \sin M\omega t \quad (2.3)$$

If we then substitute this set into (2.1), we obtain a response $\hat{f}(t)$, periodic in t with period T , having $N(2M+1)$ unknowns. These are determined either by

1) Integration: We form the set of integral equations

$$\frac{1}{T} \int_0^T \hat{f}(t) dt \triangleq \underline{f}_0(\underline{x}_0, \underline{x}_{1k}, \underline{x}_{2k}) \quad (2.4)$$

$$\frac{2}{T} \int_0^T \hat{f}(t) \cos kwt dt \triangleq \underline{f}_{1k}(\underline{x}_0, \underline{x}_{1k}, \underline{x}_{2k}) \quad (2.5)$$

$$\frac{2}{T} \int_0^T \hat{f}(t) \sin kwt dt \triangleq \underline{f}_{2k}(\underline{x}_0, \underline{x}_{1k}, \underline{x}_{2k}) \quad (2.6)$$

We then solve the above equations for $\underline{x}_0, \underline{x}_{1k}, \underline{x}_{2k}$.

2) Vector Optimization Techniques: We assume some solution for $\underline{x}_0, \underline{x}_{1k}, \underline{x}_{2k}$, $k=1, \dots, M$, and substitute into (2.3) and (2.1). An optimization technique is used to change the coefficient vectors until some error function is minimized.

This technique in general suffers from several problems. For example,

- 1) The choice of meaningful frequency components can be difficult and significant harmonics can be ignored.
- 2) The method suffers from convergence problems. Convergence is obtained only if the initial guess for the coefficients is "close" to a solution.
- 3) The number of variables can be quite large. For example, consider a system with 5 states, and for which we wish to consider a fundamental and 3 harmonics; then we have $5(6+1) = 35$ unknowns, the magnitude and phase for each frequency component plus the DC offset for each variable.

2.2.2 Piecewise harmonic balance technique

Nakhla and Vlach [3,4] have proposed a technique seeking to avoid some of the problems involved in the harmonic balance method. The method seeks to make the fullest possible use of the system linearity, reducing the number of variables to be optimized.

Consider the system S in Figure 2.1a, consisting of two arbitrary subnetworks S_1 and S_2 , with current $i(t)$ between them and voltage $v(t)$ across their terminals A-B. S_1 and S_2 each do not contain any source dependent on the other.

If we now separate the subnetworks, and drive S_1 with a current $i_1(t)=i(t)$, then the voltage at terminals A-B will be $v(t)$. If this $v(t)=v_2(t)$ is now applied to subnetwork S_2 , the current $i_2(t)$ will equal $i(t)$. See figure 2.1b.

Now, let us suppose that $i(t)$ is unknown. We now excite S_1 with a current $i_1(t)=i(t)$. This yields a voltage $v_1(t)$ at A-B. If $v_2(t)=v_1(t)$ is applied to S_2 , the current $i_2(t)$ is obtained. Let us define the error as $E(t)=i_2(t)-i_1(t)$. Our goal is to find an $i_1(t)$ such that $E(t)=0$ for the period.

Specifically, the algorithm is as follows:

1) Decompose the networks into a minimum number of linear subnetworks L_i , $i=1, \dots, l$ and the minimum necessary nonlinear subnetworks N_i , $i=1, \dots, N$. The terminals of the linear subnetworks are excited by a periodic current of the form

$$\underline{I}_L(x) = \sum_{k=0}^m \underline{x}_k \cos k\omega t + \sum_{k=1}^m \underline{x}_k^* \sin k\omega t \quad (2.7)$$

where $\underline{x} = \begin{bmatrix} \underline{x}_k \\ \underline{x}_k^* \end{bmatrix}$

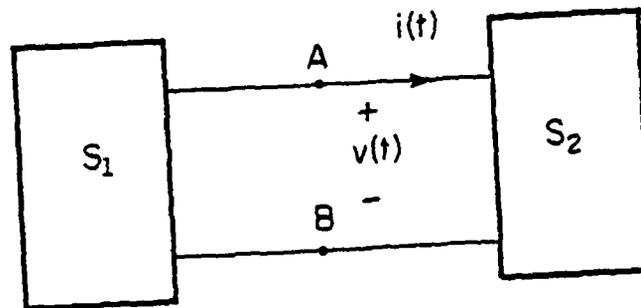
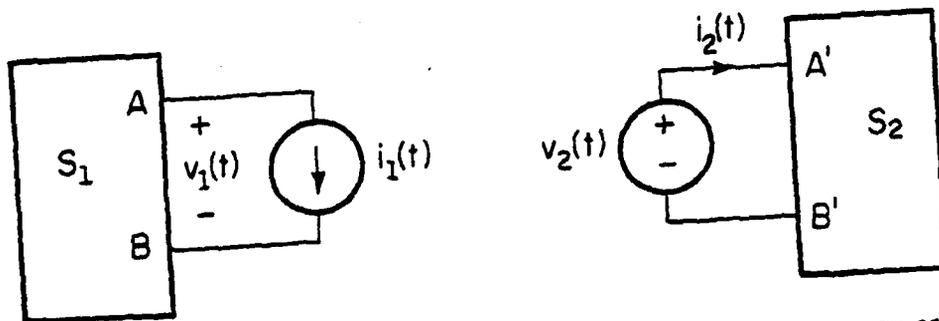


Figure 2.1a Decomposed Network



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Figure 2.1b Separated Network

Rewriting this in complex form, we have

$$\underline{I}_L(x) = \text{Re} \left\{ \sum_{k=0}^m (\underline{x}_k - j \underline{x}_k^*) e^{jk\omega t} \right\} \quad (2.8)$$

We may write the frequency domain equations of the linear system in the form

$$\underline{Y}_k(jk\omega) \underline{E}_k = \underline{A} (\underline{x}_k - j \underline{x}_k^*) + \underline{J}_k \quad (2.9)$$

where the subscript k denotes the k^{th} harmonic component; \underline{E}_k is the phasor node voltage vector and \underline{J}_k is the current excitation phasor, and \underline{A} is the node branch incidence matrix.

Using (2.9), we can write the steady state node voltage vector in the form

$$\underline{e}_k(t) = \text{Re} \left\{ [\underline{H}_k (\underline{x}_k - j \underline{x}_k^*) + \underline{b}_k] \exp(jk\omega t) \right\} \quad (2.10)$$

where

$$\underline{Y}_k \underline{H}_k = \underline{A} \quad (2.11)$$

$$\underline{Y}_A \underline{b}_k = \underline{J}_k \quad (2.12)$$

2) Choose a first estimate of the variables in (2.7).

3) Solve the linear networks in the frequency domain and use (2.11) and (2.12) to determine \underline{H}_k and \underline{b}_k , $k=0, \dots, M$, where M is the number of harmonics to be considered.

4) Using

$$\underline{e}(t) = \sum_{k=0}^m \underline{e}_k(t) \quad (2.13)$$

and (2.8) and (2.10), evaluate $I_L(t)$ and $e(t)$ at the necessary sampling points. We may use either the Discrete Fourier Transform (DFT) or Fast Fourier Transform (FFT).

Apply to the nonlinear network terminals voltage sources identical to those calculated above, and calculate the terminal currents $I_N(t)$.

5) Define an error function

$$P(x) = \int_0^T \underline{E}^t(\tau) \underline{E}(\tau) d\tau \quad (2.14)$$

where $E(t) = \underline{I}_L(t) - \underline{I}_N(t)$. Evaluate this function using numerical methods.

If P is less than some prescribed small value, stop. Otherwise continue on to step 6.

6) Use the DFT or FFT to evaluate

$$\int_0^T E(\tau) \exp(jwk\tau) d\tau \quad (2.15)$$

and

$$\int_0^T \left(\frac{\partial f}{\partial e} \right)^t E(\tau) \exp(jwk\tau) d\tau \quad (2.16)$$

7) Compute the gradient vector using

$$\underline{Z} = 2 \int_0^T \underline{E}(\tau) \exp(jwk\tau) d\tau - 2 \underline{H}_k^t \int_0^T \left(\frac{\partial f}{\partial e} \right)^t E(\tau) \exp(jwk\tau) d\tau \quad (2.17)$$

and

$$\underline{Z} = \frac{\partial P}{\partial \underline{x}_k} + j \frac{\partial P}{\partial \underline{x}_k^*} \quad (2.18)$$

If the system is autonomous, compute

$$\frac{\partial P}{\partial t} = \frac{1}{N} \sum_{i=1}^N \alpha_i \underline{E}^t(t_i, T) \underline{E}(t_i, T) + \frac{2T}{N} \operatorname{Re} \left\{ \sum_{k=0}^m (y_k^{-1} \frac{\partial y_k}{\partial T} \underline{E}_k)^t \cdot \sum_{i=1}^N \alpha_i \left(\frac{\partial f}{\partial e} \right) \underline{E}(t_i, T) \exp(jwkt_i) \right\} \quad (2.19)$$

8) Use the error function P and the gradient vector $\frac{\partial P}{\partial \underline{x}}$ as arguments in the optimization routine (Nakhla and Vlach use the Fletcher method [5]).

9) Use the correction vector $\Delta \underline{x}$ returned to compute the next estimate of \underline{x} .

For an autonomous system, use the incremental change T to readjust the period for the next iteration, and go to step (3). Otherwise go to step (4).

Nakhla and Vlach report that the best strategy was to start with only a few harmonics, obtain an optimum or near optimum solution for this value of M , then increase M and solve again, continuing until the error is small enough.

2.1.3 Volterra Series Method

A method that has found much application in the last few years is the use of Volterra Series Functionals [7]. Wiener [8] first proposed its use in 1942. The method was revived in 1967 by Narayanan [9]. It has been used extensively for distortion analysis by Narayanan [10,11] and others. Kuo and Witkowski developed a computer program using this technique to determine third-order distortion products in transistor amplifier circuits [12].

In a simple memoryless time-invariant nonlinear circuit the response can be described by the following power series,

$$y(t) = C_1 x(t) + C_2 (x(t))^2 + C_3 (x(t))^3 + \dots \quad (2.20)$$

However, in a system with memory (capacitors and inductors in an electrical circuit), the linear term is replaced by a convolution integral

$$y_1(t) = \int_0^t C_1(t-\tau) x(\tau) d\tau \quad (2.21)$$

where we have assumed $x(t)=0$ for all $t<0$, or in the transformed domain

$$Y_1(s) = C_1(s) X(s) \quad (2.21a)$$

Similarly, the second order term becomes a double convolution integral of the form

$$y_2(t) = \int_0^t \int_0^t C_2(t-\tau_1, t-\tau_2) x(\tau_1) x(\tau_2) d\tau_1 d\tau_2 \quad (2.22)$$

Note that the output depends upon products of the past values of the input, and $C_2(t-\tau_1, t-\tau_2)$ is the second-degree Volterra Kernel.

Transforming the above, we obtain

$$Y_2(S_1, S_2) = C_2(S_1, S_2) X(S_1) X(S_2) \quad (2.23)$$

In general, we may represent the n^{th} term as

$$y_n(t) = \int_0^t \dots \int_0^t c_n(t-\tau_1, t-\tau_2, \dots, t-\tau_n) \prod_{i=1}^n x(\tau_i) d\tau \quad (2.24)$$

and in the frequency domain as

$$Y_n(s) = C_n(S_1, S_2, \dots, S_n) \prod_{i=1}^n X(S_i) \quad (2.25)$$

In applying this method to the analysis of circuits, the first term is represented as a linear element, and the nonlinear terms are modeled as current sources, following the method of Bussgang, et. al. [13].

From the preceding discussion, it is obvious that this method is primarily useful where only a few terms of the expansion are required. This is the case only if the circuit is mildly nonlinear, and thus the Volterra Series approach finds use in applications such as communications systems analysis, where the system is nearly linear. Since we are concerned with highly nonlinear circuits, this method is of little help.

2.2 TIME DOMAIN METHODS

It is often more convenient in the case of highly nonlinear circuits to represent the system in the time domain.

The most obvious method available is the contraction mapping approach. First, we suppose that the system will in time reach a steady state. Then providing our analysis algorithm is sufficiently accurate, we may follow the system response from the initial state until the steady state is reached.

This "brute force" approach is not without its problems, however. To follow the system response with the necessary accuracy, we must keep the time step small compared to the period. If it then requires many periods for the steady state to be reached, it becomes very costly to proceed this way.

In order to avoid this long process, several methods have been developed which lead more or less directly to the steady state solution. We will review some of these now.

2.2.1 The Extrapolation Method

Skelboe [14] has proposed the use of the ϵ -algorithm of Wynn [15,16], an extrapolation method, for the steady state analysis of systems.

We will consider a system of N nonlinear ordinary differential equations with a periodic solution,

$$\dot{y} = \underline{f}(y, t) \quad (2.26)$$

with

$$y(t_0) = y_0 \quad (2.26)$$

We can formulate the steady state problem as finding a solution \underline{z} of the operator equation

$$\underline{z} = \underline{F}(\underline{z})$$

where

$$\underline{F}(\underline{z}) = \underline{z} + \int_0^T \underline{f}(y(\tau), \tau) d\tau \quad (2.27)$$

For the ϵ -algorithm, we define the initial values

$$\begin{aligned} \epsilon_{-1}^{(r)} &= 0 \\ \epsilon_0^{(r)} &= y_r \quad r = 0, 1, \dots \end{aligned}$$

and the recursion formula

$$\epsilon_{s+1}^{(r)} = \epsilon_{s-1}^{(r+1)} + (\epsilon_s^{(r+1)} - \epsilon_s^{(r)})^{-1} \quad (2.28)$$

6) Let $\underline{z}_n = \underline{z}_{2m}^{(k)}$

7) Terminate if $|E_{2m}^{(k)} - E_{2m}^{(k-1)}| < \delta$, where δ is estimated from an error analysis of the E-algorithm. Otherwise return to step (1) with $n=n+1$.

Several advantages are claimed for the ϵ -algorithm, including

(1) The method requires only the resultant vectors of a transient analysis routine. Thus, it is applicable without major changes to a variety of analysis algorithms, and is able to utilize sophisticated variable order and stepsize numerical integration methods efficiently.

(2) The algorithm is applied component-wise, and can utilize efficiently the separation of the time constants of different state variables. Along with the aforementioned low-pass filtering, this may significantly reduce the order of the problem.

(3) It is claimed that the ϵ -algorithm converges "faster" than the Newton algorithm, but faster in the sense of number of iterations.

The extrapolation method is not without problems, however. Among these are

(1) It is difficult to determine values of q and m without considerable prior knowledge of the circuit. An estimate of m can be made from the algorithm, but this requires that the initial iteration be made with $m=N$.

(2) If the transients involved are mostly slowly decaying, m is close to N in any case, and each application of the ϵ -algorithm requires integration of the system over $2m$ periods.

(3) It is difficult to compare the cost of computation for the E-algorithm and other methods, such as the Newton method. Some examples by Skelboe seem to show that many more forward periods are required for the ϵ -algorithm, while the Newton method adds little to the cost of the fewer periods of integration. But no direct comparisons are made.

The ϵ -algorithm seems to hold some promise for application to programs involving sophisticated numerical methods for the solution of small to medium size problems, particularly those that are mildly nonlinear. It does not seem, however, to be the final answer to the more general problem.

2.2.2 The Gradient Method

Nakhla and Branin [17,18] have proposed the use of a gradient method for the determination of the steady state solution. Here, an optimization routine is used to reduce some error measure to a zero (small) value.

Consider a system of nonautonomous differential equations, of the form

$$\dot{\underline{x}} = \underline{f}(\underline{x}, t) \quad (2.29)$$

where \underline{f} is periodic with period T .

We assume the existence of a periodic solution vector $\underline{x}_p(t)$ such that

$$\underline{x}_p(t) = \underline{x}_p(t + T) \quad (2.30)$$

We define a discrepancy vector

$$\underline{\delta}(\underline{x}_0) = \underline{f}(\underline{x}_0; T) - \underline{x}_0 = \int_0^T \underline{f}(\underline{x}, t) dt \quad (2.31)$$

where $\underline{f}(\underline{x}_0; T)$ implies dependence on the initial conditions \underline{x}_0 . This discrepancy vector is a measure of how far we are from a solution of our problem. For the steady state, the discrepancy vector vanishes; thus, our problem is to solve the nonlinear implicit system

$$\underline{\delta}(\underline{x}_0) = 0 \quad (2.32)$$

A scalar error function P is defined as the square of the Euclidian norm of the discrepancy vector, i.e.,

$$P(\underline{x}_0) = [\underline{\delta}(\underline{x}_0)]^T [\underline{\delta}(\underline{x}_0)] \quad (2.33)$$

where the superscript t denotes transposition. Clearly, we wish to minimize P , with a zero value corresponding to our desired periodic point.

The optimization method to be used (Fletcher's method [19]) requires the gradient vector

$$g(x_0) = \frac{\partial P(x_0)}{\partial x_0} = 2 \left[\frac{\partial f(x_0; T)}{\partial x_0} - I \right]^t \quad (2.34)$$

We now require a computationally efficient method of evaluating this vector. We first differentiate (2.29) with respect to x_0 , obtaining the matrix differential equation

$$\frac{\partial \dot{x}}{\partial x_0} = \frac{\partial f}{\partial x} \Big|_{x(t; x_0)} \frac{\partial x}{\partial x_0} \quad (2.35)$$

This linear time-dependent equation has the form of the variational equation

$$\dot{y} = \frac{\partial f}{\partial x} \Big|_{x(t; x_0)} y \quad (2.36)$$

This equation has as its adjoint,

$$\dot{z} = - \left(\frac{\partial f}{\partial x} \right)^t \Big|_{x(t; x_0)} z \quad (2.37)$$

which is also linear and time-dependent.

A fundamental property of linear, time-dependent differential equations and their adjoints assumes that the solutions obey

$$y(t)^t z(t) = y_0^t z_0 = \text{constant} \quad (2.38)$$

for all t .

Thus, in our case, we may write

$$\left[\frac{\partial x(T; x_0)}{\partial x_0} \right]^t z(T) = \left[\frac{\partial x(0; x_0)}{\partial x_0} \right]^t z(0) = z(0) = \text{constant, where } \frac{\partial x(0; x_0)}{\partial x_0} = I \quad (2.39)$$

Therefore, if we set $\underline{z}(T) = \delta(\underline{x}_0)$ in (2.37), i.e., we use the discrepancy vector $\delta(\underline{x}_0)$ as the "initial conditions" in (2.39) and then integrate backwards from $t=T$ to $t=0$, the vector $\underline{z}(0)$ obtained is what is required in (2.34). Thus, (2.34) is reduced to

$$\underline{g}(\underline{x}_0) = 2[\underline{z}(0; \delta(\underline{x}_0)) - \delta(\underline{x}_0)] \quad (2.40)$$

Thus, we summarize the method as follows:

- 1) Choose a first estimate of \underline{x}_0 - perhaps by the continuous integration of the system over several periods.
- 2) Using this value of \underline{x}_0 , integrate forward from $t=0$ to $t=T$, saving the trajectory $\underline{x}(t; \underline{x}_0)$ at each step.
- 3) Compute the discrepancy vector $\delta(\underline{x}_0)$ and its scalar magnitude function $P(\underline{x}_0)$. If $P(\underline{x}_0)$ is smaller than some error criterion, stop.
- 4) Otherwise, use this $\delta(\underline{x}_0)$ as initial conditions for (2.37), at $t=T$, and integrate backwards to time $t=0$, using the trajectory saved in step (2) to compute the matrix $\left(\frac{\partial \underline{f}}{\partial \underline{x}}\right)^t$ during this integration process.
- 5) Calculate the adjoint discrepancy vector, and double it to obtain the gradient vector $\underline{g}(\underline{x}_0)$. Calculate $P(\underline{x}_0)$ and use these as inputs to the optimization routine to minimize $P(\underline{x}_0)$.
- 6) Use the $\Delta \underline{x}_0$ correction vector return by the optimization routine to calculate the next \underline{x}_0 , and go to step (2).

Nakhla and Branin also present variations of this basic scheme to allow for the inclusion of variable phase and the autonomous system. Again, this method exhibits some advantages and drawbacks.

On the positive side, the gradient method

- (1) sometimes converges over a wider range than other methods, particularly the Newton method.

(2) Allows for the inclusion of arbitrary phase effects; since the solution may converge to any point on the trajectory, the method has a larger "target" to hit, and converges faster than most gradient methods.

(3) There is a possibility that the gradient method may be faster for large problems than the Newton method; however, this is somewhat fragile. As far as is known, no direct speed comparisons have been made.

Some of the problems encountered with the gradient method are

(1) More iterations are required than with most other methods; for small to medium problems, each iteration is longer than some other methods.

(2) The not-inconsequential problem of storing and recovering the trajectory of the system for the backward integration. For large problems large amounts of storage are required. Further, the trajectory must be accessed in reverse order during the backward integration. Also, with methods of variable step size, interpolation of the trajectory will be necessary.

(3) Scaling seems useful to speed convergence. However, no clear method is given in order to produce a useful scaling scheme.

(4) A large amount of computational overhead is involved with the method, optimization routine, etc.

Since the gradient method's success is tied closely to the efficiency of the method used for minimizing $P(x_0)$, the method is attractive to those possessing optimization routines. But again, it is not the solution for all problems.

2.2.3 Equivalent Linear Analysis

Several authors [20-26] have treated the highly nonlinear DC-DC power converter problem using various "equivalent" linear methods, with varying degrees of success.

These methods depend on the conversion of the nonlinear problem to an "equivalent" linear one, and applying well-known methods to the solution of this linear problem.

As an example, we will consider the "averaging" method of Middlebrook and Cuk [22,25]. Each nonlinear circuit over the period T is represented by a succession of linearized circuits. For example, in a continuous-current regulator the number of linear circuits necessary is 2. Over the period, each of these is active for an interval $d_n T$; thus, linear model 1 is active over an interval $d_1 T$; linear model 2 for $d_2 T$. Obviously, $\sum d_n = 1$. Using a state-space representation, we have

(1) over the interval $d_1 T$

$$\begin{aligned}\dot{\underline{x}} &= \underline{A}_1 \underline{x} + \underline{b}_1 u \\ \underline{y}_1 &= \underline{C}_1^t \underline{x}\end{aligned}\tag{2.41}$$

(2) Over the interval $d_2 T$

$$\begin{aligned}\dot{\underline{x}} &= \underline{A}_2 \underline{x} + \underline{b}_2 u \\ \underline{y}_2 &= \underline{C}_2^t \underline{x}\end{aligned}\tag{2.42}$$

Next, a simple averaged system is proposed, for which

$$\begin{aligned}\dot{\underline{x}} &= d_1 (\underline{A}_1 \underline{x} + \underline{b}_1 u) + d_2 (\underline{A}_2 \underline{x} + \underline{b}_2 u) \\ \underline{y} &= (d_1 \underline{C}_1^t + d_2 \underline{C}_2^t) \underline{x}\end{aligned}\tag{2.43}$$

The justification for this is basically the assumption that the fundamental matrix

$$e^{\underline{A}t} = 1 + \underline{A}t + \dots$$

may be represented by the linear term, and that the matrices \underline{A}_1 and \underline{A}_2 commute. These assumptions are not at all valid, yet the authors claim very good results.

The final result is a normal form system where

$$\begin{aligned}\dot{\underline{x}} &= \underline{A} \underline{x} + \underline{b} \underline{u} \\ \underline{y} &= \underline{C}^t \underline{u}\end{aligned}$$

with

$$\begin{aligned}\underline{A} &= d_1 \underline{A}_1 + d_2 \underline{A}_2 \\ \underline{b} &= d_1 \underline{b}_1 + d_2 \underline{b}_2 \\ \underline{C}^t &= d_1 \underline{C}_1^t + d_2 \underline{C}_2^t\end{aligned}\tag{2.44}$$

having a global state transition matrix

$$\phi [T, t_1, \underline{A}_1, \underline{A}_2; \underline{x}(0)] = e^{[d_1 \underline{A}_1 + d_2 \underline{A}_2] T}\tag{2.45}$$

Middlebrook and Cuk also go on to show that for these "averaged" equations, it is always possible to represent the averaged system by a set of appropriately averaged elements; thus, a complete analytical expression can be obtained.

In spite of the simplicity and apparent power of the method, it suffers drawbacks.

(1) Consider a system with many nonlinear elements such as diodes. The determination of switching times and the large number of equivalent linear circuits necessary requires large amounts of computer time.

(2) As a more exact analysis is desired, the two segment nonlinear device models are replaced with m segment models, further complicating the linear analysis task.

(3) In any case, the accuracy of the methods is limited by the accuracy of the linearization of the nonlinear devices. In some cases, such as power converters, where these devices may be characterized either as "on" or "off", this does not represent a problem. However, in many other cases the usefulness of the method is limited.

Thus, the equivalent linear methods find application in some special cases, but we still have not solved the general problem.

2.2.4 The Newton Method

Trick and Aprille [27-29] have proposed the use of a Newton method for the steady state analysis. This algorithm is an application of the well-known Newton method, widely used in transient analysis, to the steady state problem. Consider a system of the form

$$\dot{\underline{x}} = \underline{f}(\underline{x}, t) \quad (2.46)$$

where \underline{x} and \underline{f} are n-vectors; \underline{f} is periodic with period T; and there exists a periodic solution $w(t)$ of period T.

The problem is then expressed as a two-point boundary value problem, since in the steady state

$$\underline{x}(0) = \underline{x}(T) \quad (2.47)$$

We have

$$\underline{x}(T) = \int_0^T \underline{f}(\underline{x}_0, \tau) d\tau + \underline{x}(0) \quad (2.48)$$

and thus can express this as a mapping

$$\underline{x}(T) = F(\underline{x}_0) \quad (2.49a)$$

where

$$\underline{x}_0 = \underline{x}(0) \quad F(\underline{x}_0) = \int_0^T \underline{f}(\underline{x}_0, \tau) d\tau + \underline{x}_0 \quad (2.49b)$$

The Newton method applied to this yields

$$\underline{x}_0^{k+1} = \underline{x}_0^k - [\underline{I} - \underline{F}^1(\underline{x}_0^k)]^{-1} [\underline{x}_0^k - \underline{F}(\underline{x}_0^k)] \quad (2.50)$$

Given the initial state \underline{x}_0^1 , assume a unique trajectory $\underline{x}^1(t)$, $0 \leq t \leq T$ exists for the system (2.46). Since $\underline{x}^1(T) = \underline{F}(\underline{x}_0^1)$,

$$\underline{F}^1(\underline{x}_0^1) = \left. \frac{\partial \underline{x}^1(T, \underline{x}_0)}{\partial \underline{x}_0} \right|_{\underline{x}_0^1} \quad (2.51)$$

Trick and Aprille have shown [27] that this is equivalent to finding the state transition matrix of the system of first variation about the trajectory $\underline{x}^i(t)$.

Thus,

$$\underline{F}^1(\underline{x}_0^i) = \underline{\Phi}(T, 0; \underline{x}_0^i) \quad (2.52)$$

Trick, Colon, and Fan [30] have presented a computational technique for computing this sensitivity matrix directly, without the formulation of the state equations. Thus, the approach is easily incorporated into existing programs for the transient analysis of circuits.

Aprille and Trick have also given a method whereby the Newton algorithm may be extended to autonomous systems [28].

Colon and Trick [31,32] have also proposed a modified Newton method which improves convergence. In summary, this method includes an initial iteration over three periods without applying the Newton method in order to allow short-term transients to decay, noting that the sensitivity matrix does not necessarily change significantly for a new set of initial conditions, establishing a set of criteria for the computation of a new sensitivity matrix, and the establishment of a damping factor to be applied to the sensitivity matrix which reduces the size of the Newton step when the system is far from the steady state.

The Newton method has been shown to be very effective in reducing the computation necessary for the determination of a steady state solution. The method has been reported to have been used in several circuit analysis programs [31, 33-35] and has served as the foundation of further work in the area [35,36].

The principal drawbacks to the Newton method are

- (1) The addition of the mechanisms for the computations of the sensitivity.

networks and the formation of $\frac{1}{2}$ require extensive revision of transient analysis algorithms.

(2) It is difficult to use variable order techniques with the necessity of calculating sensitivity networks.

(3) Convergence is only guaranteed in the neighborhood of the solution.

In spite of these criticisms, the Newton method is still an efficient means of calculating the steady state response of a circuit. The goal of this work is to add some further improvements to the method in order to provide reliable and efficient analysis of nonlinear circuits.

CHAPTER 3

SOME PROBLEMS ENCOUNTERED IN THE CLASSIC
NEWTON METHOD AND POSSIBLE SOLUTIONS

3.1. THE CLASSIC NEWTON METHOD

The many advantages of the Newton method clearly indicate that this technique is worth exploring in detail. Trick, et al., [27-31] have presented a method which yields excellent results on many, but not all, problems.

Unfortunately, the application of a Newton method to a nonlinear problem does not guarantee convergence to a solution. Convergence proofs are local in nature, and we are never assured of global convergence. Further, we are not usually able to determine the region of local convergence so that we are never sure of finding a solution.

Within the region of local convergence the Newton method is quadratically convergent. If the method converges at all from outside the region as is often the case, we have no estimate on the rate of convergence.

In this chapter we shall be concerned with improvements in the reliability of the Newton method (increased region of convergence) and in the speed.

3.2. HIGH-SENSITIVITY PROBLEMS

There exist classes of problems which exhibit a high sensitivity to perturbations which leads to erratic behavior on the part of the Newton method.

Recall that the Newton method can be expressed as

$$\underline{x}_o^{k+1} = \underline{x}_o^k + [\underline{I} - \underline{\phi}]^{-1} [\underline{x}_T^k - \underline{x}_o^k] \quad (3.1)$$

This may be rewritten as

$$[\underline{I} - \underline{\phi}] [\underline{x}_o^{k+1} - \underline{x}_o^k] = \underline{x}_T^k - \underline{x}_o^k = \underline{\Delta x} \quad (3.2)$$

where $\underline{I} - \underline{\phi} = \underline{J}$, the Jacobian matrix, and

$$\underline{J} = \begin{bmatrix} 1 - \frac{\partial f_1}{\partial x_1} & - \frac{\partial f_1}{\partial x_2} & - \frac{\partial f_1}{\partial x_3} & \dots & - \frac{\partial f_1}{\partial x_n} \\ - \frac{\partial f_2}{\partial x_1} & 1 - \frac{\partial f_2}{\partial x_2} & - \frac{\partial f_2}{\partial x_3} & \dots & - \frac{\partial f_2}{\partial x_n} \\ - \frac{\partial f_n}{\partial x_1} & - \frac{\partial f_n}{\partial x_2} & - \frac{\partial f_n}{\partial x_3} & \dots & 1 - \frac{\partial f_n}{\partial x_n} \end{bmatrix} \quad (3.3)$$

We note that if the off-diagonal elements of any column of \underline{J} are significantly smaller than other columns, the change in the value of the state associated with that column can be large, although the values of $\underline{\Delta x}$ could be small. The i th state exhibits a high sensitivity to error, thus the choice of name.

Similarly, if the off-diagonal elements of any row of \underline{J} are smaller than those of other rows, large changes in all variables may be present from iteration to iteration for a given $\underline{\Delta x}$.

As a concrete example, let us consider the half-wave power supply circuit shown in Figure 3.1. Clearly, the initial state of C_1 will have little effect on the final states of either L_1 or C_2 . Similarly, neither the initial state of C_1 or L_1 will greatly affect the value of the final state of C_2 .

This is confirmed by an analysis of the sensitivities of these quantities using the Newton method. For the fourth Newton iteration, we have

$$\underline{J} = \begin{bmatrix} +1.00003 & 5.11158 \times 10^{-7} & -2.57946 \\ -6.38593 \times 10^{-8} & +1.00000 & 6.38580 \times 10^{-3} \\ -9.14239 \times 10^{-6} & 1.81169 \times 10^{-7} & 8.57791 \times 10^{-2} \end{bmatrix}$$

Calculating the inverse of the Jacobian, we have

$$\underline{J}^{-1} = \begin{bmatrix} 1.00025 & -5.96054 \times 10^{-6} & 30.078 \\ -6.16894 \times 10^{-7} & 1.00000 & -.074463 \\ 1.06607 \times 10^{-4} & -2.11268 \times 10^{-6} & 11.661 \end{bmatrix}$$

Remembering that the Newton equation for this case is

$$\begin{bmatrix} x_o^{k+1}(C_1) \\ x_o^{k+1}(L_1) \\ x_o^{k+1}(C_2) \end{bmatrix} = \begin{bmatrix} x_o^k(C_1) \\ x_o^k(L_1) \\ x_o^k(C_2) \end{bmatrix} + \underline{J}^{-1} \begin{bmatrix} x_T^k(C_1) - x_o^k(C_1) \\ x_T^k(L_1) - x_o^k(L_1) \\ x_T^k(C_2) - x_o^k(C_2) \end{bmatrix} \quad (3.4)$$

we see that a small difference in the initial and final states of C_2 will cause significant changes in the new choice of initial conditions for C_1 and L_1 .

However, the difference values for C_1 and L_1 will have little effect on the new initial value of C_2 .

3.3. COLON'S DAMPED NEWTON METHOD

Even though a system has only one stable periodic solution, the Newton algorithm may not be well-behaved. However, the contraction mapping iterates will always converge to the solution. Thus, Colon [32] proposed a modified Newton method of the form

$$\underline{x}_o^{k+1} = \underline{x}_o^k + [\underline{I} - \alpha \underline{\phi}]^{-1} [\underline{x}_T^k - \underline{x}_o^k] \quad (3.5)$$

where the scalar

$$\alpha = 1 - E^c$$

and E is the change in the ℓ_2 norm of the vector of capacitor charges and inductor fluxes. Colon states that this criterion worked well and gave good results for a large number of different circuit types. Note that if $\alpha = 1$, this is the full Newton method; if $\alpha = 0$, contraction mapping results.

3.4. VECTOR DAMPED NEWTON METHOD

In this section a vector damped Newton method is proposed which selectively damps the sensitivity matrix so that variables which could benefit most from contraction mapping techniques are separated from those best served by the Newton method.

Experimental results using variations of this idea led to the development of the following modified Newton method.

$$\underline{x}_o^{k+1} = \underline{x}_o^k + [\underline{I} - \underline{A}\underline{\phi}\underline{A}]^{-1} [\underline{x}_T^k - \underline{x}_o^k] \quad (3.6)$$

where

$$\underline{A} = \begin{bmatrix} \sqrt{a_1} & 0 & \dots & 0 \\ 0 & \sqrt{a_2} & \dots & 0 \\ \dots & \dots & \dots & \dots \\ 0 & 0 & \dots & \sqrt{a_n} \end{bmatrix} \quad (3.7)$$

$$a_i = 1 - \left| \frac{E_{Ti} - E_{Oi}}{E_{Oi}} \right|, \quad (3.8)$$

$$0.1 \leq a_i \leq 1 \text{ for all } i.$$

E_{Ti} is the energy present in the i th state variable at time T .

E_{Oi} is the energy present in the i th state variable at time 0 .

We also form a global error

$$\alpha = 1 - E \quad (3.9)$$

where

$$E = \left\| \frac{E_{Ti} - E_{Oi}}{E_{Oi}} \right\| \quad (3.10)$$

and form

$$\underline{A} = \sqrt{\alpha} \begin{bmatrix} \sqrt{a_1} & 0 & 0 \\ 0 & \sqrt{a_2} & 0 \\ 0 & 0 & \sqrt{a_n} \end{bmatrix} \quad (3.11)$$

This technique is motivated by a consideration of the behavior of heavily damped states. If we are far from the steady state solution a heavily damped state will undergo considerable change over one period. This yields large changes in the energy and small values of a_i for this i th

state. The form of the \underline{A} matrix tends to decouple these states in the Newton method and allows them to be treated by contraction mapping without unduly affecting the states acted upon by the Newton iteration.

This method also yielded good results on the circuits cited by Colon, and provided good results on additional circuits for which the single-value damped Newton method proved inadequate.

In rare cases it was necessary to provide a lower bound for a_i , as circuits which started far from the solution could spend an excessive amount of time in the contraction mapping mode due to large changes in E compared with the initial values. This was particularly true in the case of autonomous systems in which the period was an unknown. Here, an error in T could lead to phase differences in the evaluation of \underline{x}_0 and \underline{x}_T , yielding a large value for E . A good value for this lower bound was determined experimentally to be 0.1. This bound was usually invoked only once before the computed value of a_i was larger than this lower bound. For most circuits a_i never fell below this lower bound.

3.5. PROOF OF LOCAL CONVERGENCE

Aprille [29] has shown that the classic Newton method converges for an initial state \underline{x}_0 sufficiently close to the correct solution \underline{w}_0 for a specified set of conditions. This proof is based on one offered by Ortega [37].

We will use a different approach to show convergence of the modified Newton iterates, based on the theorem of Kantoravich [53] as given by Henrici [54]. An additional discussion of the theorem and its application to the

Newton method may be found in Roberts [55].

We first make the following definitions of norms which will be used throughout.

If \underline{v} is a vector, then

$$\|\underline{v}\| = \max_{1 \leq i \leq n} |v_i| \quad (3.12)$$

If \underline{B} is a matrix, then

$$\|\underline{B}\| = \max_{1 \leq i \leq n} \sum_{j=1}^n |b_{ij}| \quad (3.13)$$

We next cite the following lemma due to Banach (Henrici [54], pp. 365).

Lemma: Let \underline{B} be a matrix such that $\|\underline{B}\| = k < 1$, and let \underline{I} denote the identity matrix. Then the matrix $(\underline{I} - \underline{B})^{-1}$ exists, and

$$\|(\underline{I} - \underline{B})^{-1}\| \leq \frac{1}{1 - k} \quad (3.14)$$

From the above lemma, it is obvious that if $\underline{\phi}$ satisfies this requirement, then the matrix $\underline{B} = \underline{A} \underline{\phi} \underline{A}$ does also, and the norm for the modified Newton method is less than or equal to that of the unmodified method. This corresponds to Aprille's requirement that $\underline{\phi}$ have no unity eigenvalues.

We wish to satisfy the system

$$\underline{F}(\underline{x}_0) = \underline{x}_0 - \underline{x}_T = 0 \quad (3.15)$$

For this system, we have the Jacobian

$$\underline{J} = \underline{I} - \underline{\phi} \quad (3.16)$$

where

$$\phi_{ij} = \frac{\partial x_{Ti}}{\partial x_{oj}} \quad (3.17)$$

The classic Newton method for this system is

$$\underline{x}_o^{k+1} = \underline{x}_o^k - [\underline{I} - \underline{\phi}(\underline{x}_o^k)]^{-1} \underline{F}(\underline{x}_o^k) \quad (3.18)$$

and the modified method is

$$\underline{x}_o^{k+1} = \underline{x}_o^k - [\underline{I} - \underline{A} \underline{\phi}(\underline{x}_o^k) \underline{A}]^{-1} \underline{F}(\underline{x}_o^k) \quad (3.19)$$

For any system of the form

$$\underline{x}_o^{k+1} = \underline{x}_o^k - [\underline{B}(\underline{x}^k)]^{-1} \underline{F}(\underline{x}^k) \quad (3.20)$$

we have the following theorem (Kantorovich):

Theorem: Assume that the following conditions are satisfied:

- i) for $\underline{x} = \underline{x}^0$, the initial approximation, the matrix $\underline{B}(\underline{x}^0)$ has an inverse $\underline{\Gamma}_0 = (\underline{B}(\underline{x}^0))^{-1}$, and an estimate for its norm is known:

$$\|\underline{\Gamma}_0\| \leq B_0 \quad (3.21)$$

- ii) The vector \underline{x}^0 satisfies the system of equations (3.15) approximately, in the sense that

$$\|\underline{\Gamma}_0 \underline{F}(\underline{x}^0)\| \leq \eta_0 \quad (3.22)$$

iii) In the region defined by equation (3.25) below, the components of $\underline{F}(\underline{x})$ are twice continuously differentiable with respect to the components of \underline{x}^0 and satisfy

$$\sum_{j,k=1}^n \left| \frac{\partial^2 F_i}{\partial x_j \partial x_k} \right| \leq K \quad i = 1, 2, \dots, n \quad (3.23)$$

iv) The constants B_0 , η_0 , K introduced in (i)-(iii) above satisfy the inequality

$$h_0 \equiv B_0 \eta_0 K \leq \frac{1}{2} \quad (3.24)$$

Then the system of equations (3.15) has a solution \underline{x}^* which is located in the hypercube

$$|\underline{x} - \underline{x}^0| \leq N(h_0) \eta_0 = \frac{1 - \sqrt{1 - 2h_0}}{h_0} \eta_0 \quad (3.25)$$

Moreover, the successive approximations \underline{x}^k defined by (3.20) exist and converge to \underline{x}^* , and the speed of convergence may be estimated by the inequality

$$|\underline{x}^k - \underline{x}^*| \leq \frac{1}{2^{k-1}} (2h_0)^{2^{k-1}} \eta_0 \quad (3.26)$$

We do not repeat the details of this proof here; it may be found in Henrici [54], pp. 366.

Let us determine how this theorem applies to the modified Newton method as compared to the unmodified method. In the following discussion a prime will be used to denote the modified system. Clearly condition (iii) is satisfied for both cases within the bounds of the smaller cube if it is

satisfied for either. Condition (i) is satisfied for the modified method if it is satisfied for the unmodified method, and further by the previous lemma $B'_0 \leq B_0$. Therefore, if condition (ii) is met by the modified system, and the value for η'_0 satisfies condition (iv), then the sequence of iterates defined by (3.6) converges. We may also determine that if $h'_0 \leq h_0$, the hypercube defined in (3.25) is larger, and therefore we may approach the solution \underline{x}^* from an initial point further away from the correct solution.

3.6. A MODIFIED NEWTON METHOD FOR AUTONOMOUS SYSTEMS

For autonomous systems a technique very similar to that outlined above was proposed. However, we have no estimate of the individual error for the period of the autonomous system. Any error in the estimate of period would be evidenced by phase changes in many states. Thus, the total error would be proportional to the error in period. Therefore this error was used to compute the damping factor to be applied to the period.

Some small changes in the formulation of the modified Newton equations are also necessary to provide continuity between the Newton and contraction mapping techniques. We will not cover this in detail as this has already been shown by Colon [29], but give the necessary changes.

We first obtain K , the index of a state variable within the orbit of oscillation. We then form the series of Newton iterates

$$\underline{v}_0^{k+1} = \underline{v}_0^k - [\underline{I}_k - \underline{A} \underline{\varphi}_k \underline{A}]^{-1} [\underline{x}_0^i - \underline{x}_T^i] \quad (3.27)$$

where

$$\underline{v}_0^k = [x_{01}^k, x_{02}^k, \dots, x_{0K-1}^k, T, x_{0K+1}^k, \dots, x_{0n}^k]^T \quad (3.28)$$

\underline{I}_K is the Identity matrix of rank n with the k th diagonal element replaced by zero.

and

$\underline{\varphi}_K$ is equal to the sensitivity matrix

$\underline{\varphi}$ except that the K th column of $\underline{\varphi}$ is replaced by $\underline{f}(\underline{x}_T^k)$, and the K th row of $\underline{f}(\underline{x}_T^k)$ denoted by $\underline{f}(\underline{x}_T^k)$ is replaced by $f_K(\underline{x}_T^k)/\alpha^2$, where α is the global damping coefficient.

No proof of local convergence for this technique is yet available. As before, as $\alpha \rightarrow 0$ the method approaches contraction mapping; as $\alpha \rightarrow 1$, the full Newton step is taken.

3.7. SUMMARY OF THE DAMPED NEWTON METHOD

The final choice of method was a vector damped Newton method incorporating the reduced calculation of the sensitivity matrix suggested by Colon [32]. This method also merges some of the steps given by Colon. He defines

$$\underline{x}^*(o) = \underline{x}^i(o) \quad (3.29)$$

for any step in which $\underline{\varphi}$ is recalculated, and

$$EP = \frac{\|\underline{m}(\underline{x}^*(o) - \underline{x}^{i+1}(o))\|}{\|\underline{m}\underline{x}^*(o)\|} \quad (3.30)$$

where \underline{m} is the vector of reactance values at time 0 is calculated each time a new initial condition vector is proposed. EP and the total error E are used to determine whether a new calculation of the sensitivity matrix should be performed. SW1 and SW2 are internal flags used to modify the path taken. SW1 set indicates that the sensitivity matrix should be recomputed. SW2 set indicates that the sensitivity matrix has been calculated since the last contraction mapping period.

The algorithm is as follows:

- 1) Given $\underline{x}^0(o)$, integrate the system without calculating ϕ for some user-specified number N periods to remove initial fast transients.
- 2) Set $i = N$, $E = 0$, and set SW1 and SW2. Set $\underline{x}^i(o) = \underline{x}(NT)$, $\underline{x}^*(o) = \underline{x}^i(o)$.
- 3) Given $\underline{x}^i(o)$, calculate $\underline{x}^i(T)$. If SW1 is set, simultaneously calculate $\underline{\phi}$. If $|\underline{x}^i(o) - \underline{x}^i(T)|$ sufficiently small, stop.
- 4) Set ELAST = E, and calculate new values of E_i^V and E.
- 5) If $E \geq 1$, do a contraction mapping period; clear SW1 and SW2, set $\underline{x}^{i+1}(o) = \underline{x}^i(T)$, increment i and go to step 3. Otherwise, go to step 6.
- 6) If we have not calculated $\underline{\phi}$ since the last contraction mapping step (SW2 cleared), do so now; set SW1 and SW2, set $\underline{x}^{i+1}(o) = \underline{x}^i(T)$, increment i, set $\underline{x}^*(o) = \underline{x}^i(o)$, and go to step 3. Otherwise, go on to step 7.
- 7) Calculate a_i . Use the modified Newton method to calculate $\underline{x}^{i+1}(o)$, increment i, and calculate EP.
- 8) If $EP \geq .2$, recalculate $\underline{\phi}$ in the next iteration; set SW1 and SW2, set $\underline{x}^*(o) = \underline{x}^i(o)$, and go to step 3. Otherwise, go to step 9.
- 9) If $E < ELAST$, don't recalculate $\underline{\phi}$; clear SW1, set SW2, and go to step 3. Otherwise, recalculate $\underline{\phi}$; set both SW1 and SW2, and go to step 3.

This procedure is followed until the error vector, calculated in step 3, is smaller than the error specified.

Although it is not claimed that this method is "best" for all problems, it does yield good results on a wide class of problems. It is not significantly slower than previous Newton methods on any test problems, and produces good results on problems which are not otherwise tractable.

3.8. AN EXTERNALLY-SUPPLIED DAMPING VECTOR AND STATE ELIMINATION

It is assumed that in the design phase of circuit evolution, the design engineer has a good grasp of the operation of the circuit and can therefore make intelligent decisions concerning the convergence of state variables. It was therefore proposed to allow the designer to specify the elements of the A matrix. If these values were chosen to be either 1 or 0, the circuit could be partitioned into those states which decayed rapidly toward the steady state and those for which the Newton method would be advantageous.

This technique of separating a system into components with different classes of time constants is not new. Edsberg [38] reports on several suggestions for separating the components of a system of ordinary differential equations describing a chemical reaction into fast and slow components.

The convergence properties of such a method are those specified previously as long as the elements of A are bounded by 0 and 1 inclusively.

It is probable that in actual use the partitioning of such a system would have to be dynamic, and perhaps interactive. Initially a zero vector (contraction mapping technique) would be applied to the entire system. As the solution approached full convergence the system could move closer and closer to the full Newton step, thereby taking advantage of the quadratic rate of convergence in the neighborhood of the solution.

3.9. PROVISION OF CONTRACTION MAPPING PERIODS FOR THE DECAY OF FAST TRANSIENTS

States with a high degree of damping contribute to erratic behavior of the Newton method in two ways. First, they often exhibit a high sensitivity to error in some other state, contributing to large changes in the

initial state vector. In addition, the large degree of damping produces large values in the discrepancy vector $(\underline{x}_T^k - \underline{x}_0^k)$ unless the initial guess is close to the correct value.

Consider that these fast transients are associated with states so heavily damped that they come close to the steady state in one period. Now, let us insert a contraction mapping period after each Newton step. Thus, given a new initial vector $\underline{x}^i(0)$ calculated by the Newton method, we integrate for one period and set

$$\underline{x}^{i+1}(0) = \underline{x}^i(T) \quad (3.31a)$$

and integrate for a second period, simultaneously calculating ϕ , and find

$$\underline{x}^{i+2}(0) = \underline{x}^{i+1}(0) + [\underline{I} - \phi(\underline{x}^{i+1}(0))]^{-1} [\underline{x}^{i+1}(T) - \underline{x}^{i+1}(0)] \quad (3.31b)$$

Note that we need to calculate the Jacobian only during the second period.

During the contraction mapping period the large transients generated by high sensitivity elements in the Jacobian during the last iteration decay. Therefore they do not affect the Newton step, and the elements of the discrepancy vector are uniformly small.

Obviously, the usefulness of this technique depends on the ability to separate the states into fast and slow groups, the fast states decaying in a few periods. However, for systems which are so partitionable this method is natural and easily implemented.

All of the methods outlined above have been included in a general purpose DC and transient analysis program ("SINC"); they may be applied to a problem separately or in combination. In the next chapter we will discuss the details of this implementation.

CHAPTER 4

THE SINC PROGRAM

In this chapter we describe the program used to implement these ideas. The framework chosen was SINC, an existing program for the Simulation of Integrated Nonlinear Circuits. We first provide an introduction to SINC. Then, we describe several changes and corrections made to the basic program. Finally, major additions and changes for the implementation of these modified Newton algorithms will be described.

4.1. INTRODUCTION TO SINC (SIMULATOR FOR INTEGRATED NONLINEAR CIRCUITS)

The SINC simulation program is designed for general DC and transient analysis of moderately complex circuits. SINC originally descended from TIME [39] at the University of California, Berkeley. It then was restructured by Young and Dutton at Stanford University, where it was referred to as MSINC (A Modular Simulator for Integrated Nonlinear Circuits) [40]. Further changes were made by Kao at the University of Illinois [33].

The current version of SINC includes elements of several other programs. The dc analysis routine of SINC uses the step-limiting algorithm of CANCER [41], rather than the method of Colon [29] or that used in BIAS-3 [42].

The user has a choice of Backward Euler, trapezoidal, or second order Gear's method [43,44] for the integration of the system. The default of Gear's method provides superior results for stiff systems [44].

Further details on the program are available in Kao [33] or the User's Manual [45].

4.2. CHANGES AND CORRECTIONS TO SINC

In this section we describe several changes to SINC which do not reflect major changes or modifications. Several of these are transparent to the user and are described to inform users of prior versions of SINC of the presence of errors in the Program. Others reflect corrections or changes to the program which affect the user, either by changing input formats or output data.

4.2.1. User Transparent Changes

Changes were made to store the transformer initial conditions in the temperature coefficient initial condition vector above the element (R,L, or C) temperature coefficients or initial conditions. Previously, initial conditions for the transformer currents overwrote these locations.

The FFT algorithm of Brigham [46] was rewritten to improve its efficiency. The user was also given the option of choosing either the FFT algorithm or conventional discrete Fourier analysis. Also, the components were changed so that the sine term prints under the "sin" heading.

The current through voltage sources was calculated, thus allowing the user to return currents as output values.

A problem was corrected with the timing of the steady-state algorithm. Prior to this correction, numerical error buildup could cause the algorithm to terminate one timestep too early, resulting in an error in the value of the period of an oscillator or an erroneous steady-state solution. Also, a plot of the output did not always continue over the interval specified.

An error in the JFET modeling routine was corrected, allowing the

user to specify more than one JFET model.

4.2.2. Fourier Analysis Routines

The user can now specify the FFT algorithm or conventional Fourier analysis. The calling arguments are different. For the FFT, the format is

FFT TSTR TSTP NUX

where FFT indicates that the FFT algorithm is to be employed; TSTR is the starting time point; TSTP is the final timepoint; and 2^{NUX} is the total number of sample points. If more than 2^{NUX} sample points are available, only the first 2^{NUX} are used.

For discrete Fourier analysis using a conventional technique, the command is

FOR TSTR TSTP NHAR

where FOR identifies a conventional Fourier analysis requirement; TSTR and TSTP are the starting and stopping times respectively, and NHAR is the number of harmonics desired. In the case of an oscillator, setting both TSTR and TSTP equal to 0.0 will communicate the period found by the Newton algorithm to the Fourier algorithm.

4.2.3. Output Format Changes

The format of some outputs has been changed to provide further information or reduce confusion. For example, in the output plots off-scale and coincidence indicators are printed. Arrows indicate values off scale in either direction, while an "X" indicates coincidence of two or more characters.

If an unrecognized field is encountered while reading data, the line containing this field is printed. Below that, another line contains an up-arrow pointing to the unrecognized field or fields, and a warning message is printed.

The information provided during the verification phase of the program has been greatly increased. This includes additional information about the circuit data provided, and information on some parameters for transient and steady state analysis and special conditions which may be present.

4.2.4. Input Format Changes

The control card NDC specifies that no DC analysis is to be performed prior to the transient analysis. The time domain program begins at the zero state, or some other state specified by including initial conditions on the reactive element cards. This card is primarily useful when the the circuit is an oscillator which has an unstable solution at the DC operating point. Thus, if transient analysis is performed starting from this DC state, the time domain analysis algorithm may indicate convergence has been achieved. Use of an "NDC" card usually avoids this problem.

The type specification on the FET model card has been expanded. Rather than the specification field "CTYP = 1" or "CTYP = 2" specifying an n or p-channel device respectively, the user can simply specify "NJF" or "PJF" for n or p-channel junction FET. The format for the tolerance control card is now

TOLERANCE TOL NTRA NTRY1

where TOLERANCE is a key word; TOL is the convergence tolerance on junction voltages, expressed as a percentage of VT; NTRA is a code specifying the integration method to be used for transient analysis (-1 for Backward Euler; 0 for Gear's second order method, and +1 for trapezoidal rule); and NTRY1 is the number of tries allowed to reach convergence (minimum is 5; default is 10). These values apply to both transient analysis and each period of the steady-state analysis.

The steady state analysis card format has also been changed. The form is

```
STEADY  KSTE  FREQ  PERIOD  NST  TOL  NTRY2  NTRY0
```

where the key word STEADY indicates that a steady state analysis is required. KSTE is a code which indicates whether the circuit is nonautonomous (0) or autonomous (1). FREQ is the fundamental frequency of the system, or an estimate if the system is autonomous. The PERIOD is the period of the system; if both the frequency and period are specified as non-zero, the period is computed as $1/\text{FREQ}$. NST is the minimum number of steps allowed over the period, and it determines the maximum timestep and also the print interval. TOL is the tolerance of convergence for the steady state algorithm, expressed as a fraction of the maximum value of the state over the period. NTRY2 is the number of periods allowed to achieve the steady state. NTRY0 is the number of continuous integrations to be performed before the first Newton step is taken. Setting this equal to zero means the first full period analysis will be used to calculate the first Newton step. If a non-zero frequency is specified, any or all of the last five values can be left

out. Default values for NST, TOL, and NTRY2 and NTRY0 are 50, 1.0E-4, 20, and 3 respectively.

If both TIME and STEADY time domain control cards are included, the minimum timestep computed from the values on the STEADY card overrides the explicit values given on the TIME card.

Two new control cards are also available to users of the steady state analysis capability. These are PRINT STATES and PRINT TRANSIENT ANALYSIS. They were developed to reduce the amount of output from SINC during steady state analysis. Without these cards, SINC now prints only the state values for the final Newton iteration, and prints intermediate points as specified using TIME and VOUT or IOUT cards only during the final iteration. Plots are always of the final period only.

The inclusion of a PRINT STATES card in the input deck will cause the state values at each period to be printed. For Newton steps, this will include both initial and final states. This option is useful for checking convergence toward a solution of the Newton method chosen.

The "PRINT TRANSIENT ANALYSIS" card will cause all transient analysis output to be printed for each step of the steady state analysis algorithm. This option is capable of generating considerable amounts of output. It is included so that the user could determine the region of operation of a circuit during the steady state analysis.

4.3. NEW IMPLEMENTATIONS

4.3.1. Diode Model

Previous versions of SINC had no provision for inclusion of semiconductor diodes. We have added this feature in order to eliminate the need for using transistor junctions to serve as diodes. The model chosen is that used in SPICE2 [47], with appropriate modifications for use with SINC. This model is applicable to either junction diodes or Schottky-Barrier diodes. It includes a single linear resistance for modeling both ohmic resistance and high-level injection effects. This series resistance is denoted by R_s in Figure 4.1.

The DC diode characteristics are determined by R_s and the nonlinear current source I_d . The value of I_d is given by the diode equation

$$I_d = I_s (e^{V_d/\eta V_t} - 1) \quad (4.1)$$

where I_s is the saturation current, V_d is the diode voltage, η is the emission coefficient, and V_t is the thermal voltage, given by

$$V_t = \frac{kt}{q} \quad (4.2)$$

where k is the Boltzmann constant, t is the absolute temperature in degrees Kelvin, and q is the electronic charge. The values I_s , η , and V_s may be estimated from dc measurements of the forward-biased diode characteristics, as shown in Figure 4.2. In the straight-line ideal region below about 600 mV, Equation (4.1) may be written as

$$\log_{10}(I_d) = \log_{10}(I_s) + 2.3 \frac{V_d}{\eta V_t} \quad (4.3)$$

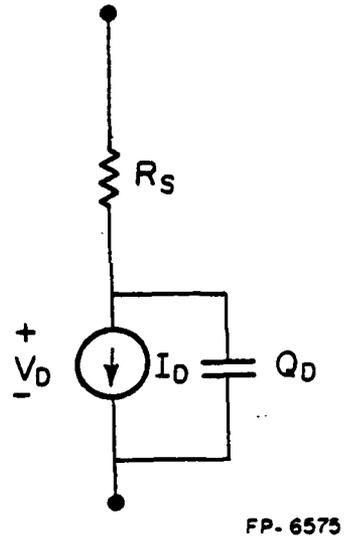


Figure 4.1 SINC Diode Model

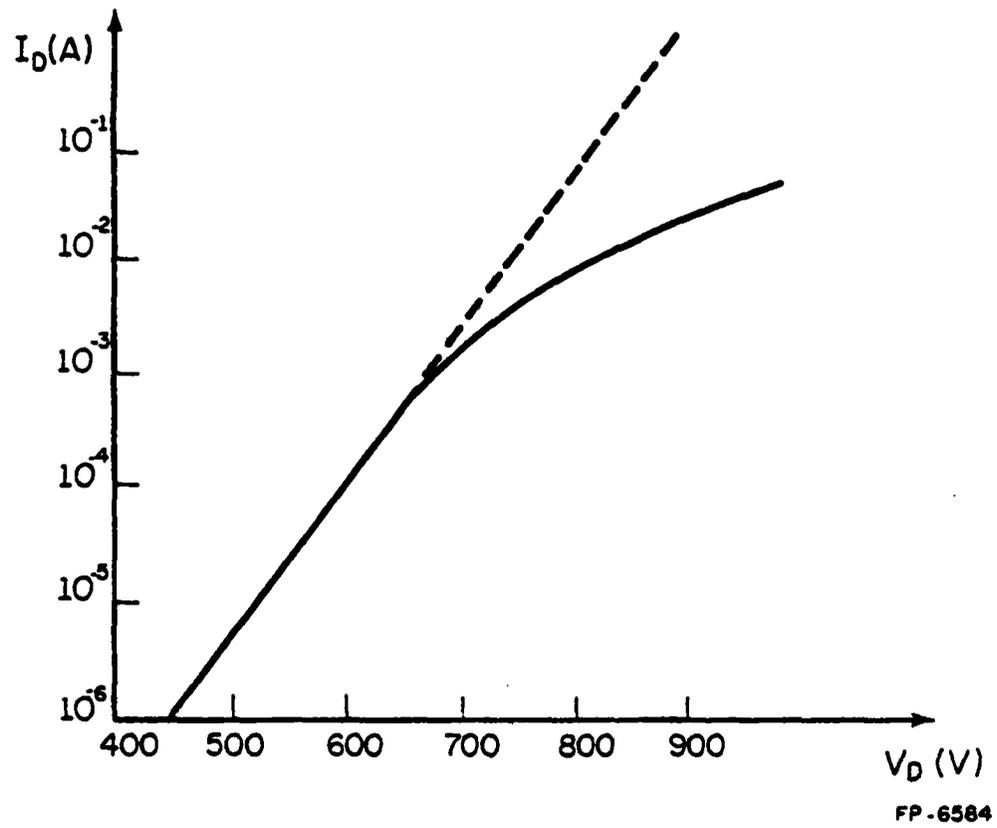


Figure 4.2 Graph of $\text{Log}(I_D)$ Versus V_D in the Forward-Bias Region

In this region, the current increases by an order of magnitude approximately every 60 mV. From Figure (4.3), we find that I_s can be determined from the intercept $V_d = 0$ of the diode current. Experimentally, several points in the forward active region are plotted and extrapolated to find the value of I_s . The emission coefficient η is determined from the shape of the characteristic line in the ideal region. In most cases $\eta = 1$.

The departure of the actual curve from the ideal straight line is due to high-level injection effects. These nonlinear effects are modeled by a single-valued linear resistance. Several points are plotted and averaged to determine the value of R_s ; if the region of operation can be narrowed, a more appropriate value of R_s can be chosen.

The time-dependent charge storage effects are modeled by C_d in Figure 4.1. The charge stored is given by the equation

$$Q_D = \tau_t I_s [e^{V_d/\eta V_T} - 1] + C_{j\phi} \int_0^{V_d} [1 - \frac{V_d}{\phi_B}]^{-m} dV \quad (4.4)$$

Equivalently, we may differentiate (4.4) to obtain the incremental capacitance

$$C_d = \frac{\partial Q_D}{\partial V_d} = \frac{\tau_t I_s}{\eta V_T} e^{V_d/\eta V_T} + C_{j\phi} [1 - \frac{V_d}{\phi_B}]^{-m} \quad (4.5)$$

The single value Q_d models both junction depletion and minority carrier storage effects. The term involving $C_{j\phi}$, ϕ_B , and m models the depletion term, and the term involving τ_t models the minority carrier storage. The values ϕ_B , $C_{j\phi}$, and m are obtained experimentally by capacitance bridge measurements in the reverse biased region of diode operation. The value of the transit time τ_t is usually determined by pulsed time-delay measure-

ments of the diode. Further details may be found in [48-52].

The small-signal linearized diode model is shown in Figure 4.3. The nonlinear current source is replaced with a linear conductance g_d determined from the characteristic equation

$$g_d = \frac{\partial I_d}{\partial V_d} /_{op} = \frac{I_s}{\eta V_T} e^{V_d / \eta V_T} \quad (4.6)$$

where the values are determined at the operating point. At each time-point, g_d is determined iteratively to achieve convergence. The value of R_s is externally specified, and C_d is as found in [4.5].

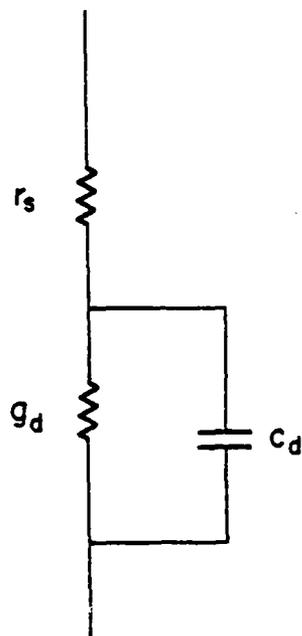
As with all other semiconductor junctions, the diode model used in SINC employs step limiting to avoid exponential overflow in the determination of diode current. At each iteration, the proposed new value of V_d^{k+1} is determined. If the previous value V_d^k is less than $10 V_t$ and V_d^{k+1} is greater than $10 V_t$, V_d^{k+1} is reset to $10 V_t$. If both V_d^k and V_d^{k+1} are greater than $10 V_t$, and $|V_d^{k+1} - V_d^k|$ is greater than $2.3 V_t$, the change is limited to $2.3 V_t$. A flowchart of this process is shown in Figure 4.4

To specify a diode for inclusion in a circuit the form is

Dxxx Zxxx N1 N2

where the initial letter D indicates a diode, xxx refers to an optional name of from 1 to 4 characters; the initial letter Z indicates that the model with name Zxxx is to be referred to for the specifications of the diode; and N1 and N2 refer to the node numbers of the anode and cathode, respectively. Each diode card refers to a specific diode; these cards can refer to the same or different model cards.

The model card format is



FP-6576

Figure 4.3 Small-Signal Linearized Diode Model

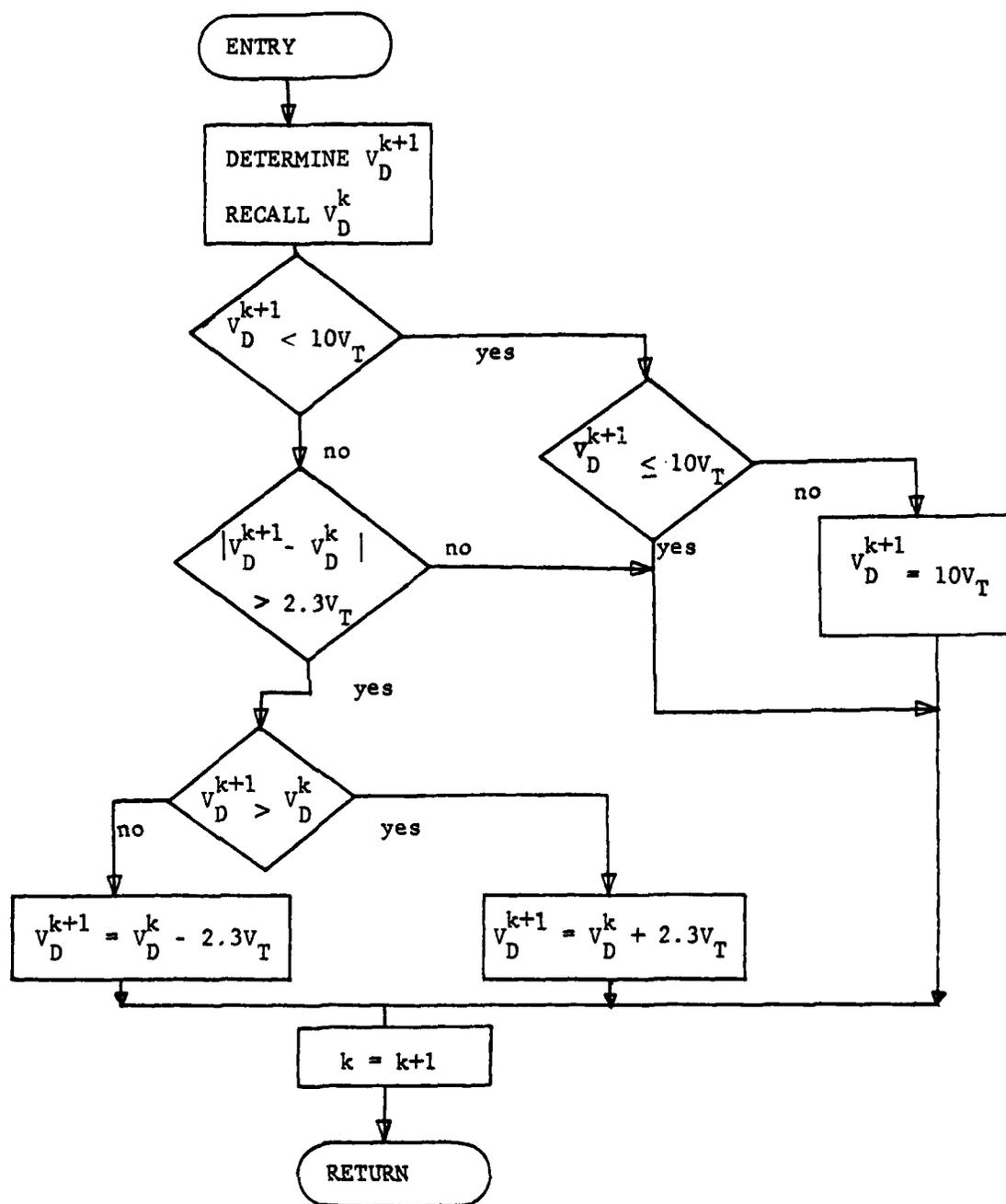


Figure 4.4 Diode Step-Limiting Flowchart

Zxxx Parameter name = value Parameter name = value

where the initial letter Z indicates that this is a model card for the model with name Zxxx, and the parameters named have the specified value. All parameters have default values; however, even if no parameters are inserted (all values are defaulted), a model card must be included. A list of parameters and typical and default values is given in Table 4.1. Current limitations on SINC allow for 10 diode and 10 model specifications.

4.3.2. Pulsed Sources

A new voltage/current source model has been added to SINC. This allows for the specification of pulsed sources. Previously, users were limited to the use of piecewise linear sources with a maximum of ten break-points. This precluded the specification of sources with a pulse repetition rate greater than the basic frequency of the circuit.

Pulsed sources are specified by a statement of the form

VPLSX N1 N2 VOFF VON TD TR TF PW PER

or

IPLSX N1 N2 IOFF ION TD TR TF PW PER

where VPLS (IPLS) refers to pulsed voltage (current) sources; X is an optional character used for differentiating separate pulse sources; N1 and N2 are the positive and negative nodes, respectively; VOFF (IOFF) is the "off" value of voltage (current); VON (ION) is the "on" value of voltage (current); TD is the time delay before the "on" period; TR is the rise time during which the pulse rises linearly from VOFF to VON (IOFF to ION); TF is the fall

Table 4.1 Diode Parameters for the SINC Model

<u>Name</u>	<u>Parameter</u>	<u>Typical Value</u>	<u>Default</u>
IS	Saturation current in Amps	10^{-14}	10^{-14}
RS	Series Resistance in Ohms	10	0
N	Emission Coefficient	1.0	1.0
TT	Transit Time in Secs	10^{-10}	0.
CJ ϕ	Zero-bias Junction Capacitance	2pf	0
PB	Junction Potential	0.6	1.0
M	Grading Coefficient	0.5	0.5

time during which the pulse falls from VON to VOFF (ION to IOFF); PW is the pulse width (on time); PER is the total period.

These sources use the piecewise linear model already present in SINC, but recompute breakpoints as required throughout the period of interest. Thus, in the case of pulsed sources whose fundamental period is much less than the period of analysis, six breakpoints are always present. These breakpoints are initialized at the beginning of each period. If the time exceeds the value of the last breakpoint computed, a new set is computed by adding PER to all the previous values except the initial one. A listing of timepoints and values is given in Table 4.2 for a voltage source. As in the SINC piecewise linear sources, linear interpolation is used between these values. A flowchart is given in Figure 4.5.

Due to the method used in SINC for timestep control (an iteration count), it is impossible to insure that very fast rise and fall times will be followed accurately. While no restriction is placed on the values of TF or TR, unlike the piecewise linear model, the user should be aware of possible problems. If possible, TR and TF should not be made smaller than the print interval. This will insure reasonably accurate values during the transition.

4.3.3. Modified Newton Methods for Steady State Analysis

Several Modified Newton methods are available in SINC for steady state analysis. These are

- a) Colon's method
- b) Vector damped method using both local and global damping factors
- c) Externally damped method
- d) Vector damped method described in Chapter 2

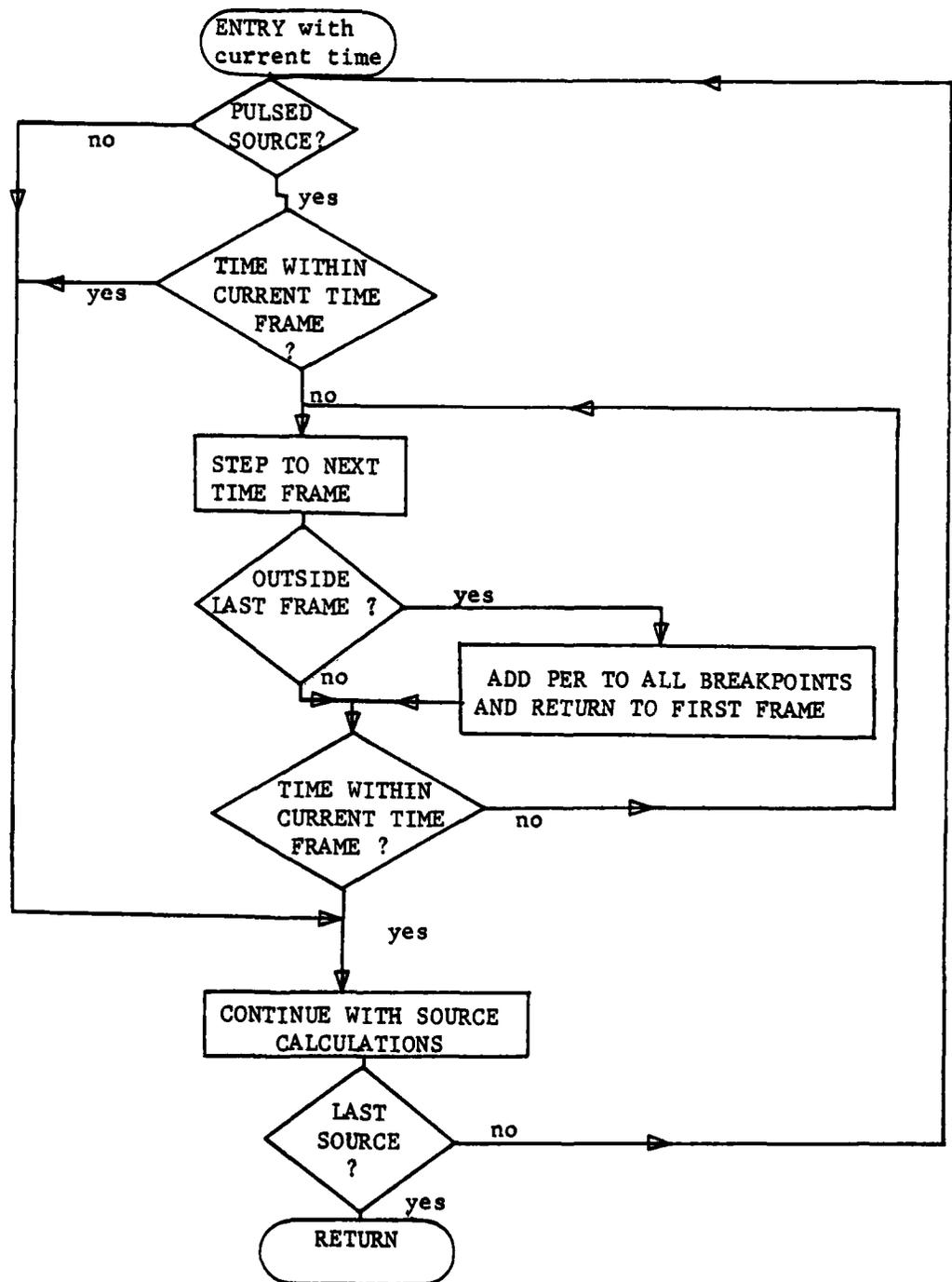


Figure 4.5 Pulsed-Source Time-Frame Flowchart

Table 4.2

<u>Timepoint</u>	<u>Value</u>	<u>Present during Period</u>
0.0	VOFF	always
TD	VOFF	1
TD+TR	VON	1
TD+TR+PW	VON	1
TD+TR+PW+TF	VOFF	1
PER	VOFF	1
PER+TD	VOFF	2
PER+TD+TR	VON	2
PER+TD+TR+PW	VON	2
PER+TD+TR+PW+TF	VOFF	2
PER+PER	VOFF	2
2 PER+TD	VOFF	3

Breakpoints and source values for pulsed voltage source in SINC

The modified Newton method to be used is specified by including a control card in the SINC input deck of the form

MODIFIED NEWTON METHOD N

where N is a number from 0 to 4. The value of N specifies the method to be used (zero corresponds to an unmodified Newton method). This allows rapid comparison of the effectiveness of the various methods. This card may be left out of the input data, in which case the default is the unmodified Newton method corresponding to $N = 0$.

The vector damped method using both global and local error factors computes an error term for each individual state and also for the system. The values of damping factor for each state and the system are computed, and the values used for the damping vector are the products of the individual damping coefficients and the global damping factor. Thus, if the entire system were far from its steady state, some damping would be applied to all states.

In order to compute these values of the error vector E, it was necessary to have available the value of the capacitance or inductance of the state. Note that in some cases, such as junction capacitance, these are time-dependent functions. Changes were made in the subroutine GETX to return these values at any time.

Using a MODIFIED NEWTON METHOD 3 card uses only the external damping values specified on an EXTERNAL card. However, using an EXTERNAL card with some other modification of the Newton method results in a damping vector that is the product of the calculated values and the values specified on the EXTERNAL card. This allows the use of a modified Newton method while

still eliminating some states by using an EXTERNAL card with zeros for the removed states.

4.3.4. The EXTERNAL Damping Specification

The elements of a user-imposed damping vector are specified using an EXTERNAL card. The form for this specification is

EXTERNAL	VALUE	VALUE	VALUE
----------	-------	-------	-------

In order to use this specification effectively, the user must be aware of the order in which the state variables are placed, as this is the order in which the damping specifications must be given. This order is as follows:

- 1) All reactive elements (inductors and capacitors) in the order in which they appear in the input data.
- 2) Transformer primary and secondary currents, in the order in which the transformers appear in the input data.
- 3) Bipolar transistor C_{eb} and C_{cb} voltages, in the order in which the transistors appear in the input data; provided however that these capacitances are non-zero.
- 4) Diode junction capacitance voltage, if this capacitance is non-zero, in the order the diodes are specified in the input data.
- 5) Junction Field-Effect Transistor Gate-Source and Gate-Drain capacitance voltage, if these capacitances are non-zero, in the order the JFET's appear in the input data.

All unspecified values default to 1.0.

A typical use of the EXTERNAL card would be to eliminate some state(s) from the Newton method. In this case, the values on the EXTERNAL card would be 1.0 or 0.0, the zero values eliminating the desired state, the full Newton step being applied to others.

In order to verify the values input via the EXTERNAL card, all externally specified and default damping factors are printed by the verification subroutine (CHEKIN) along with the damped parameters. This listing also indicates the order of the damping parameters, so any error should be easily corrected.

4.3.5. Decay Period Specification

The user may also cause SINC to insert additional periods during which the system is allowed to contract by including a card of the form

DECAY PERIODS N

where N is a single integer from 1 to 9. If no value for N is specified, the default is 1 if the DECAY PERIODS card is included. Also, DECAY PERIODS 0 has the same effect as not including the card (i.e. no contraction periods are included).

Large values of N are to be avoided, as they can greatly increase the number of periods of analysis required by the program. Each full period of analysis of the system increases the iteration counter by one. Thus, the specification of a large value for N will only allow a few Newton steps; if convergence is not reached by the limit specified on the STEADY card, a message and the final values are printed. As an example, consider an input data stream containing a DECAY PERIODS 3 card. Then each Newton iteration

would be preceded by three contraction mapping periods. Thus, four full periods of analysis would be required for each Newton step. If we wished to allow for 10 Newton steps, we would have to allow for these 40 periods plus the initial transient decay period of three or N periods, whichever is greater. In this case, the STEADY card would have to include allowance for 43 full period analysis of the system.

Obviously, the ideal case for use of this technique involves a system where "fast" states reach equilibrium in one period of time or less. The use of decay periods will not affect the Newton method in any way. The user still has his choice of the Newton method to be used (modified or unmodified) and may specify externally any desired damping vector.

No provision has been made for partial period contractions for several reasons. In general, it was felt that possible problems caused by using partial period contraction would not be compensated by the increased efficiency in a small number of cases.

4.4. CONTL-TIME DOMAIN ANALYSIS CONTROL ROUTINE

The subroutine "CONTL" has as its primary function the control of analysis in the time domain. Most of the changes to "SINC" described in the preceding pages are implemented in "CONTL". Because of the interactions of these various modifications, the logic of the subroutine has become rather complex. In order to increase the user's understanding of what his commands do, and provide a basis for change, this section will attempt to provide an explanation of the "CONTL" subroutine.

The control subroutine provides the following major functions:

- 1) Determination of a transient or steady state requirement.
- 2) Output of required variables at times required and storage of plot information.
- 3) Control of timestep size and keeping track of current time.
- 4) Control of source values.
- 5) Control of the time-domain integration subroutine.
- 6) Determination of sensitivity requirement and control of sensitivity computation subroutine.
- 7) Determination of time-step convergence.
- 8) Determination of need for contraction mapping periods and provision of these.
- 9) Determination of steady state convergence.
- 10) Determination of steady state errors and damping factors, and the switching logic that uses these values.
- 11) Newton method control and determination of an improved initial state, if necessary.
- 12) For oscillators, the necessary changes to determine the period of the oscillator.

We will discuss some of these functions in detail. Often, several of these functions have been integrated in the logic. For example, the requirement of additional contraction mapping periods suggested by Colon is easily and logically merged with a requirement for additional contraction mapping periods imposed by the user. The flowcharts included in Figures 4.6 and 4.7 are to be referenced during the following discussions.

In order to reduce the amount of output from "SINC" two flags are

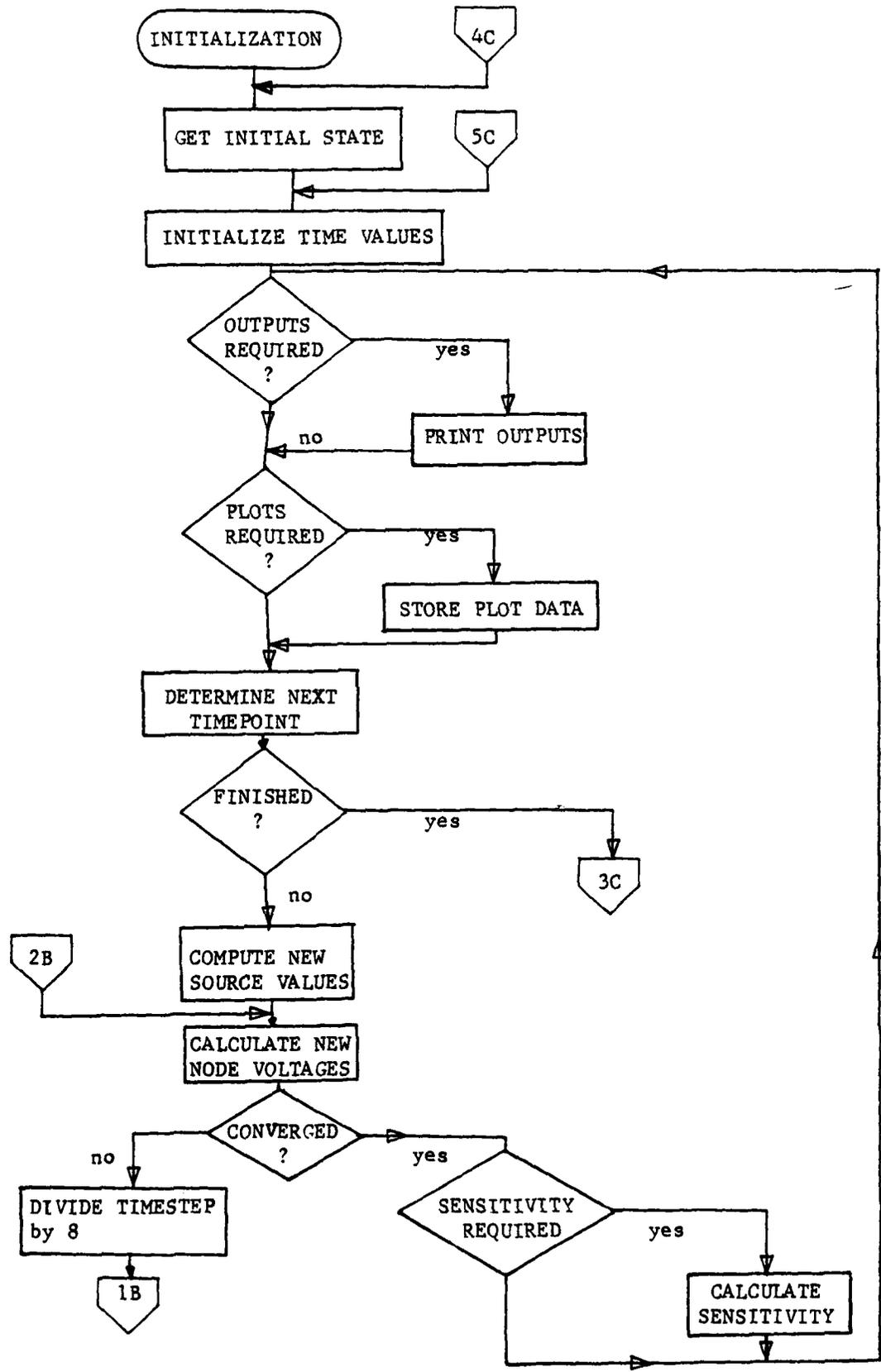


Figure 4.6a "CONTL" High-Level Flowchart

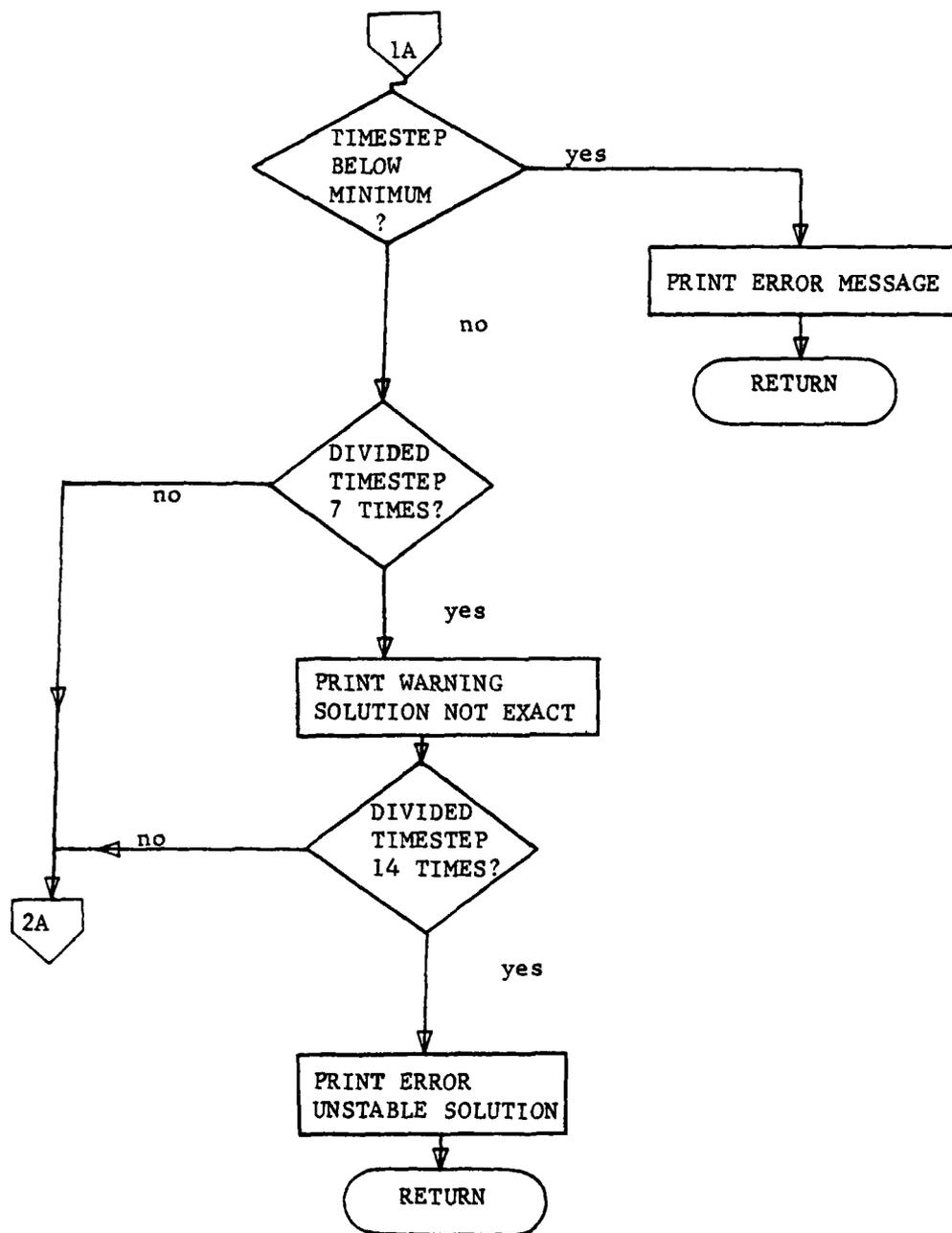


Figure 4.6b "CONTL" High-Level Flowchart

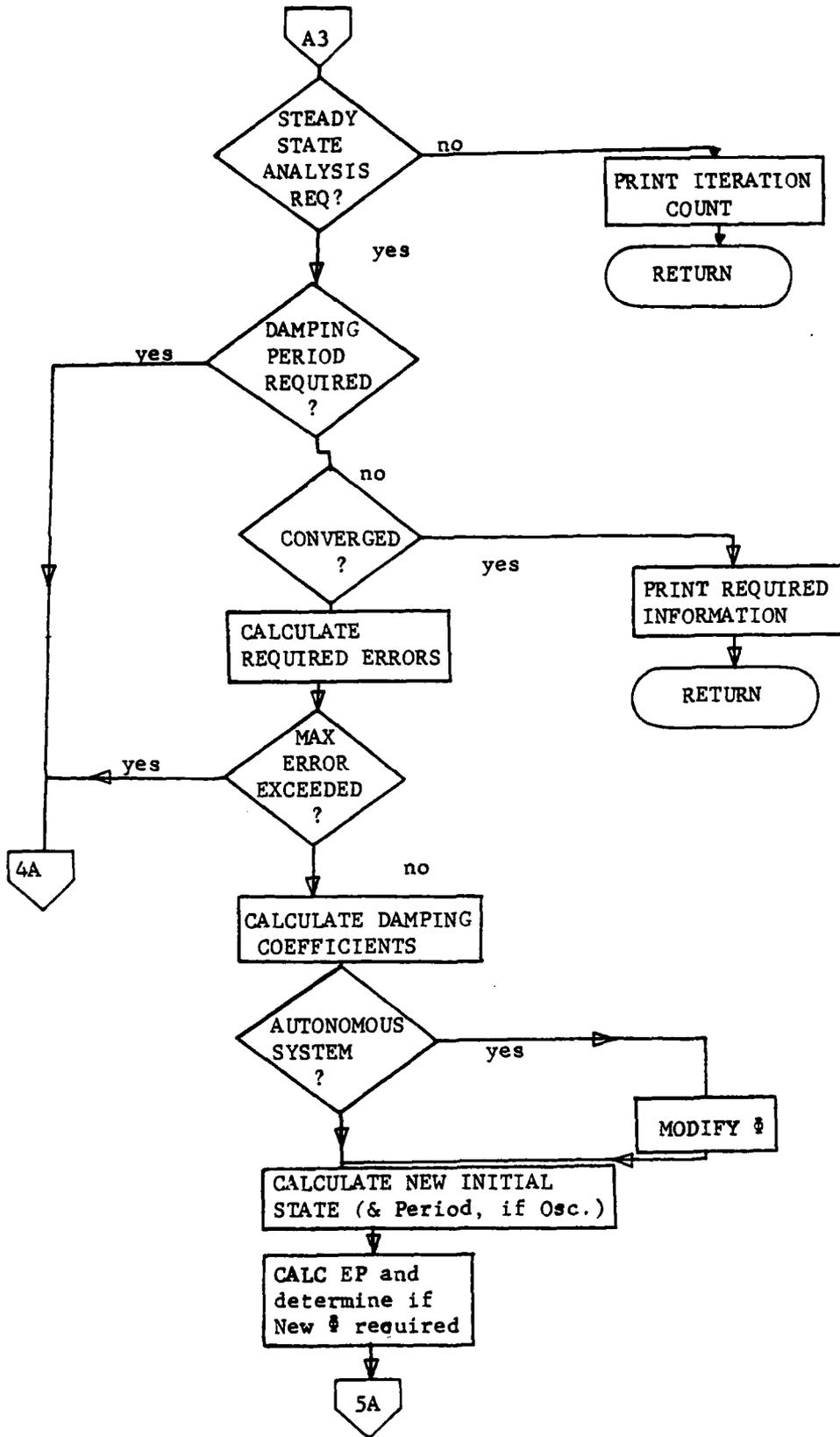


Figure 4.6c "CONTL" High-Level Flowchart

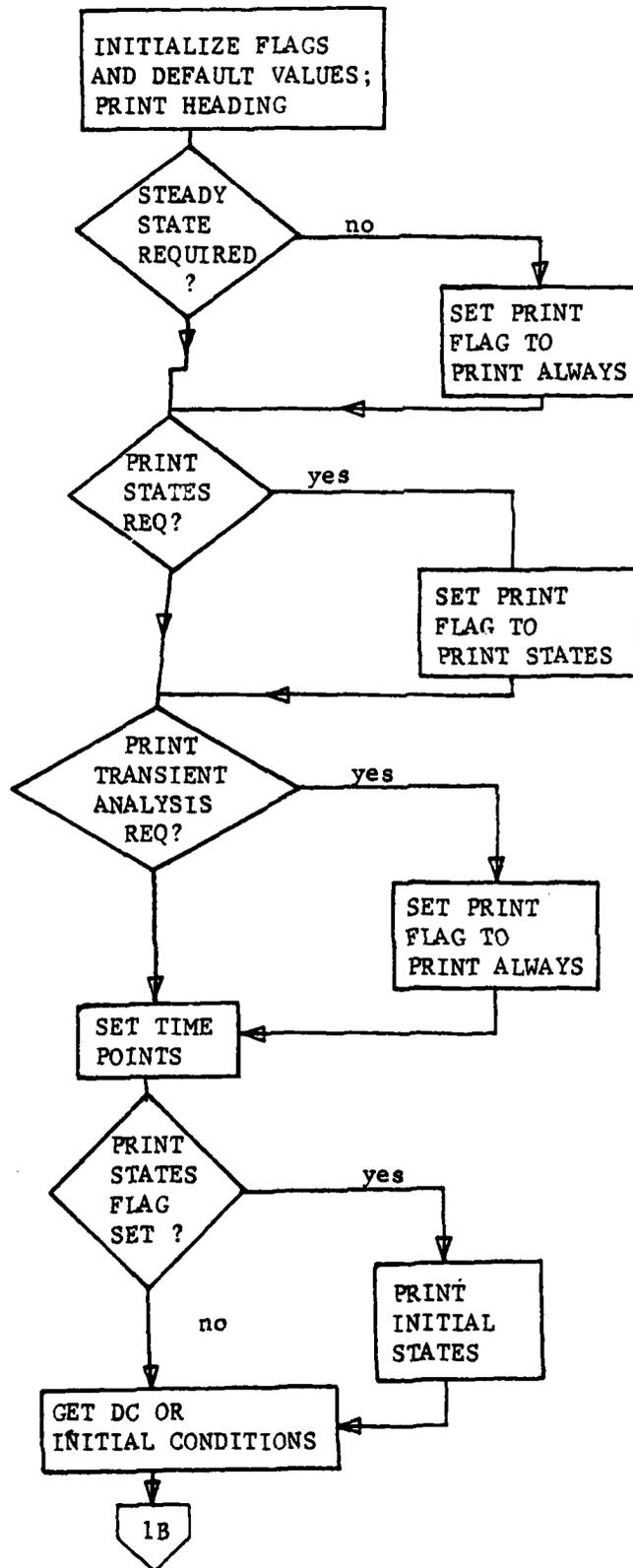


Figure 4.7a "CONTL" Detail Flowchart

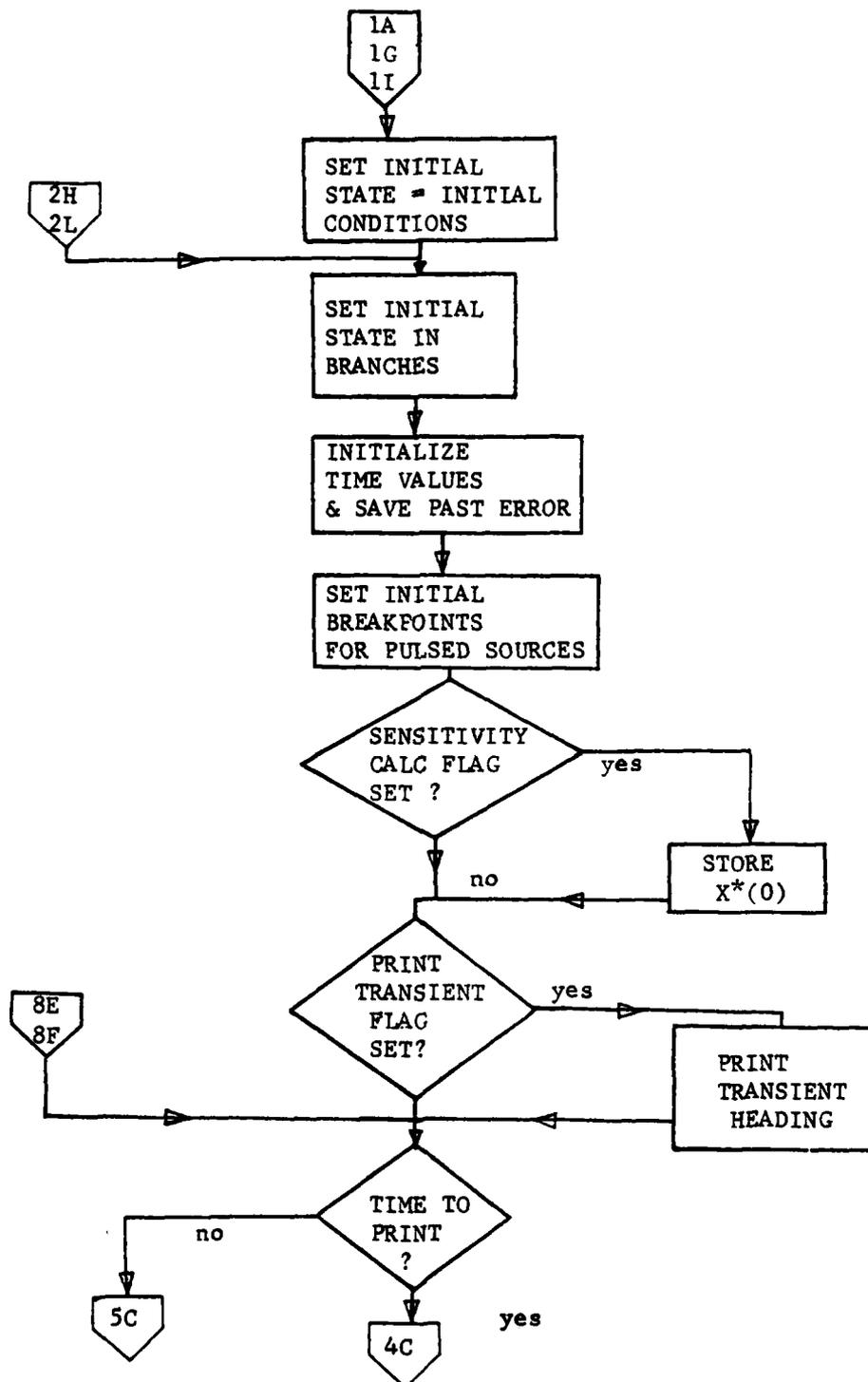


Figure 4.7b "CONTL" Detail Flowchart

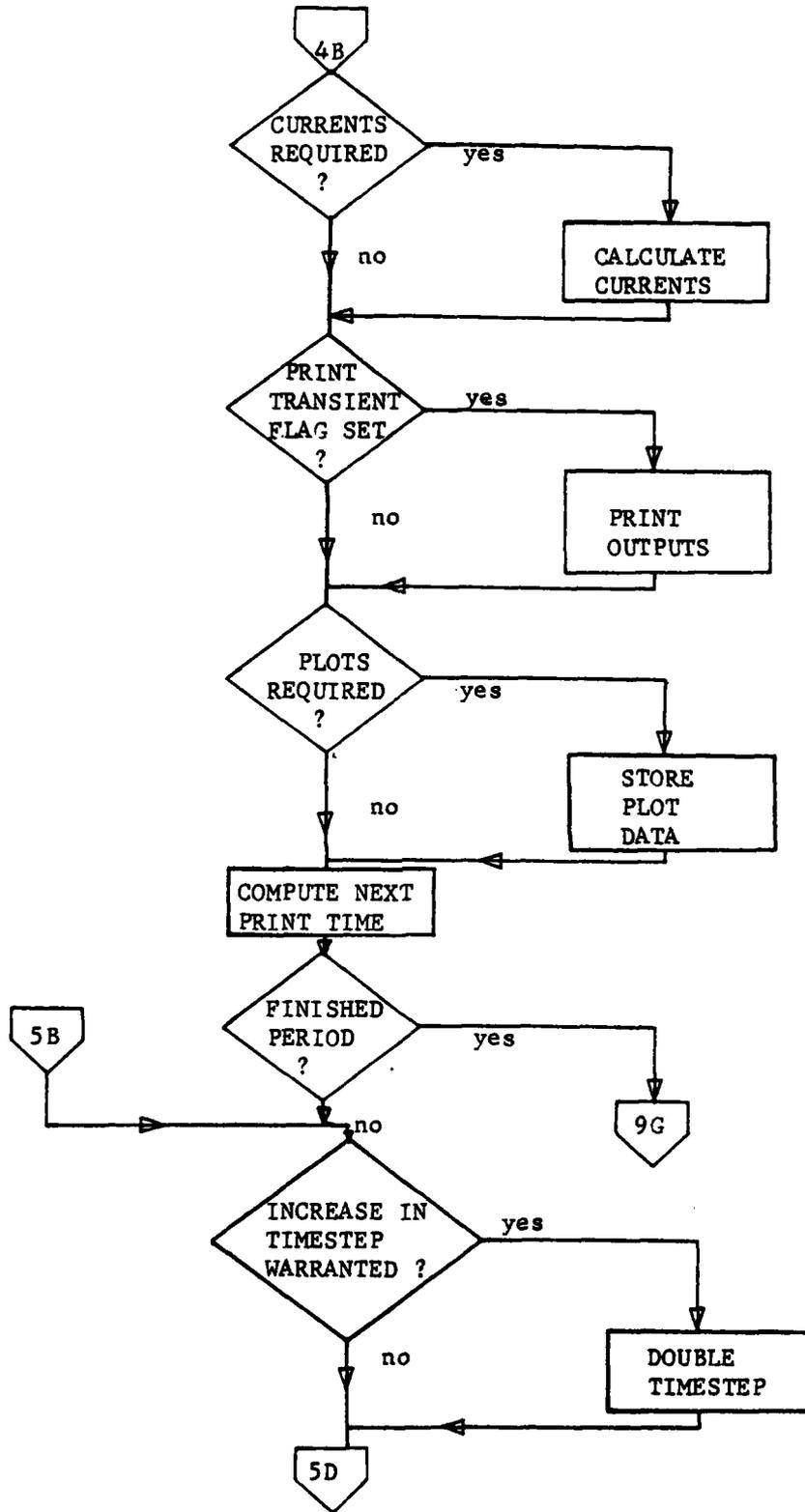


Figure 4.7c "CONTL" Detail Flowchart

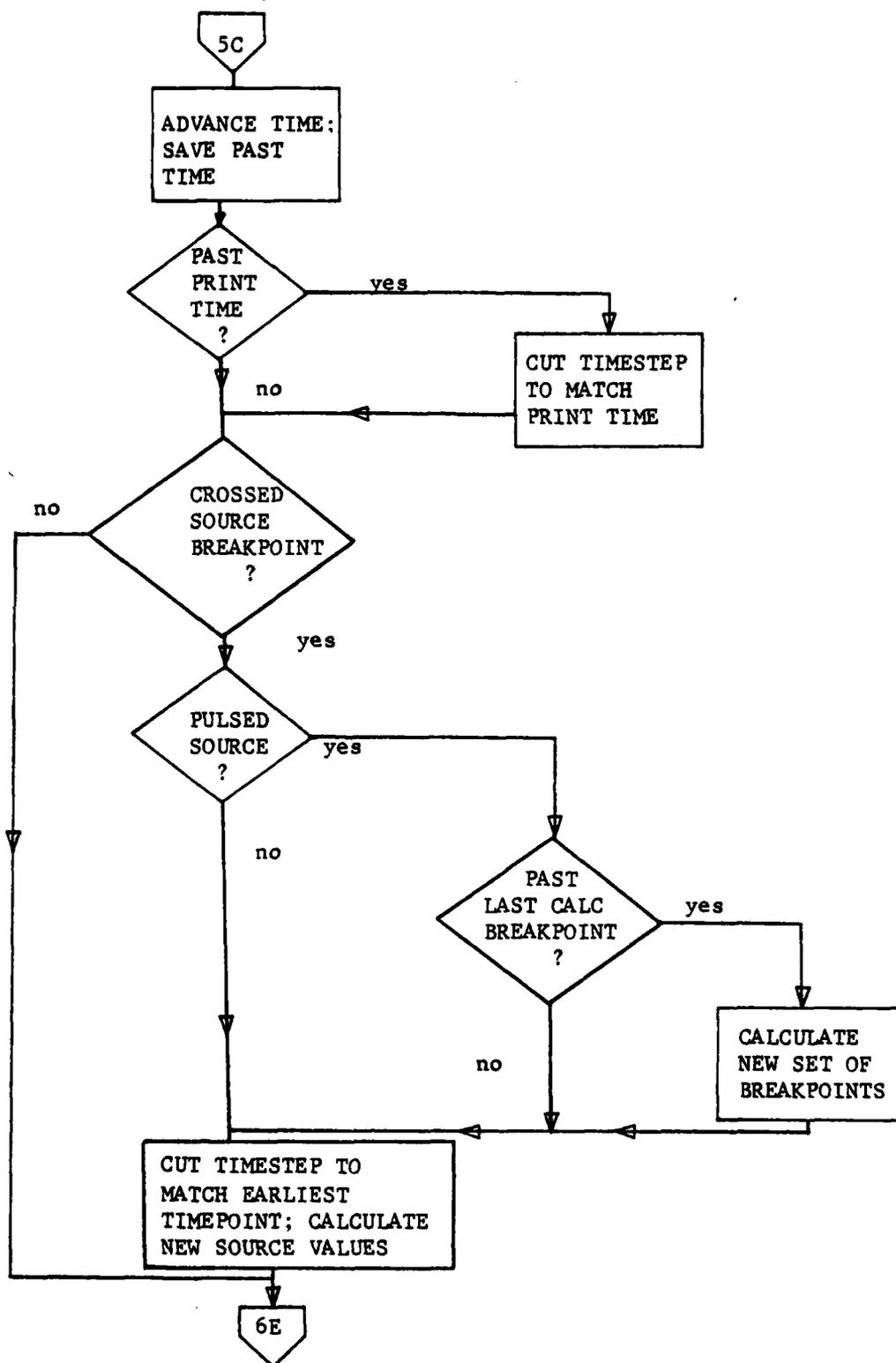


Figure 4.7d "CONTL" Detail Flowchart

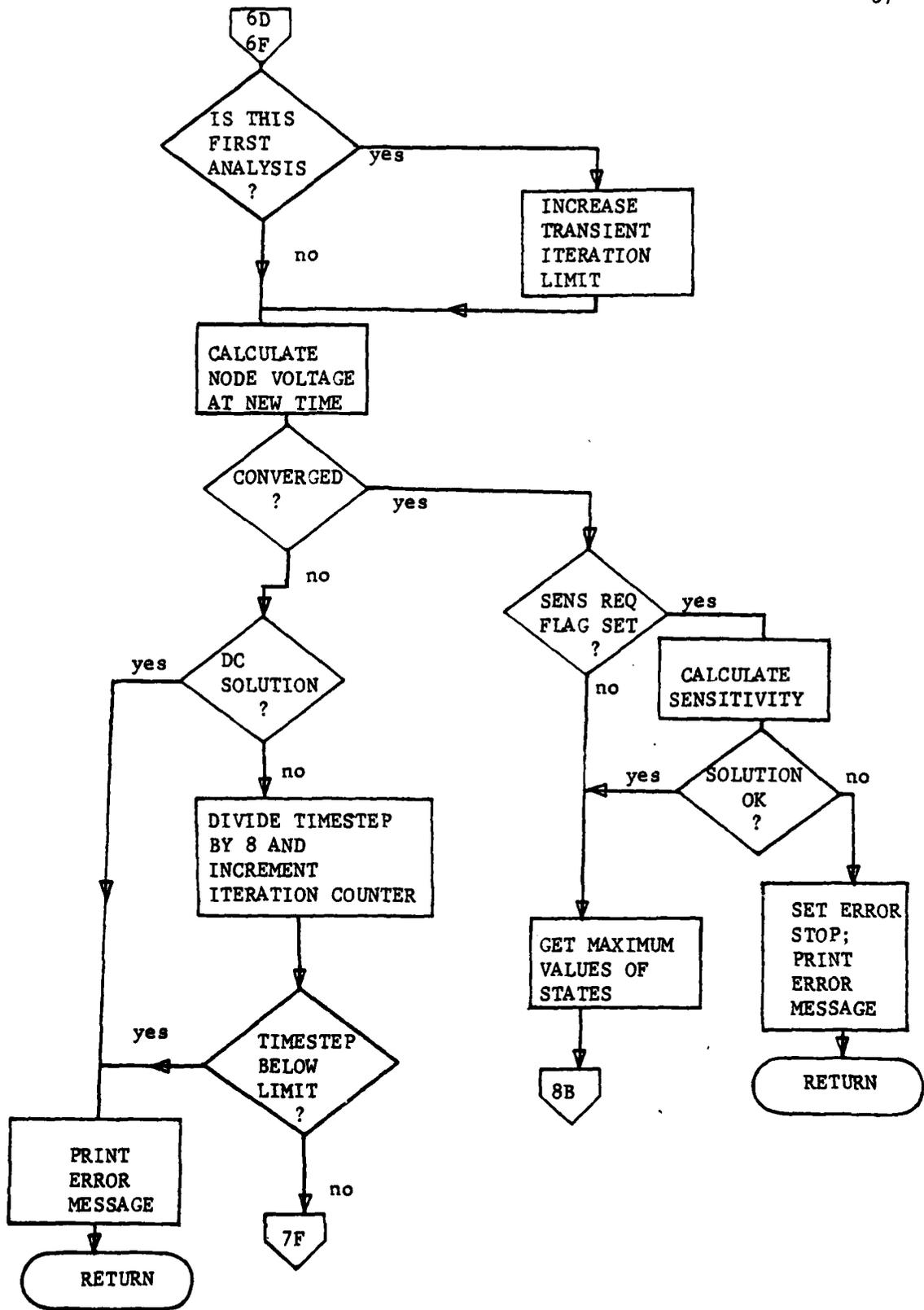


Figure 4.7e "CONTL" Detail Flowchart

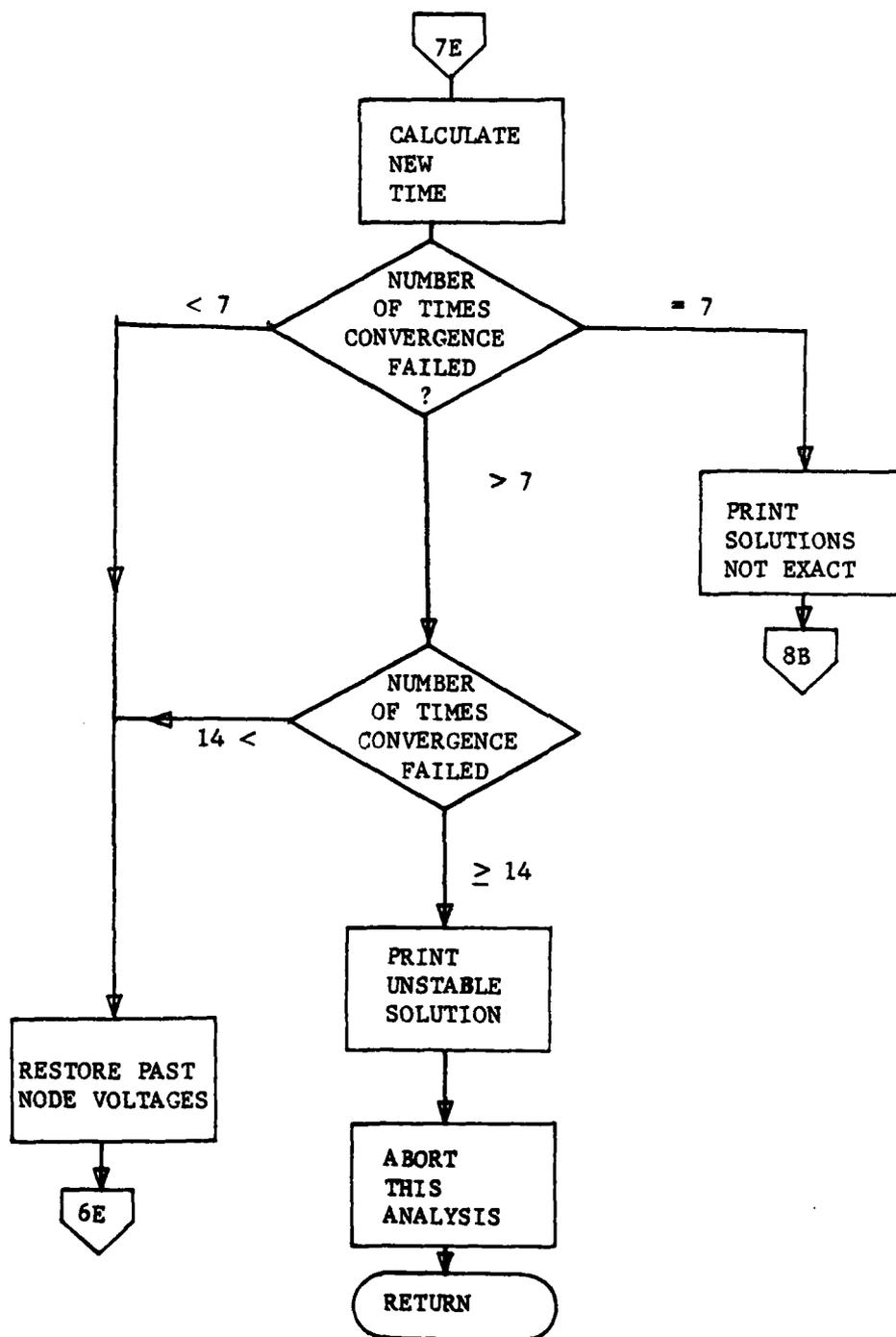


Figure 4.7f "CONTL" Detail Flowchart

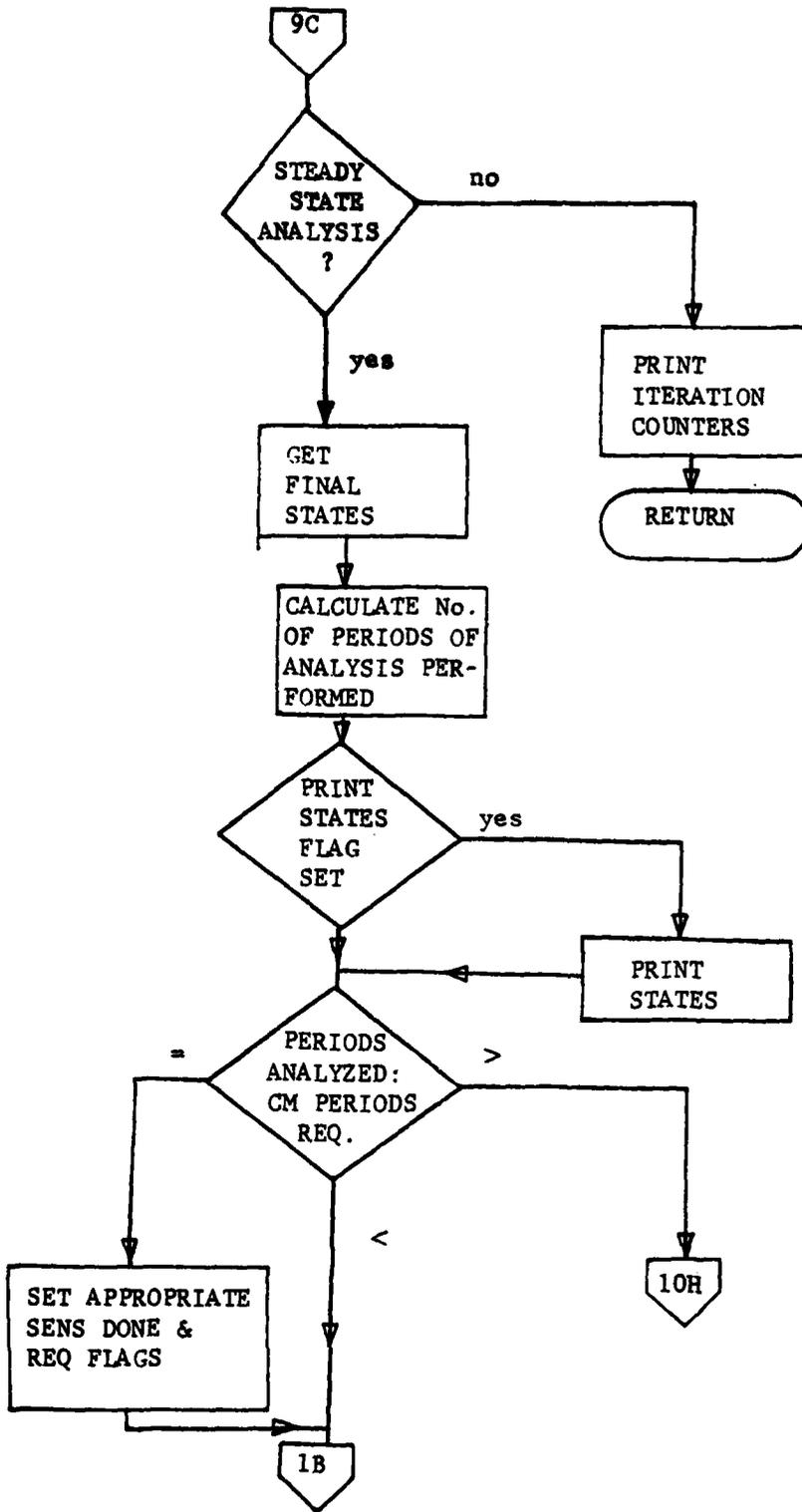


Figure 4.7g "CONTL" Detail Flowchart

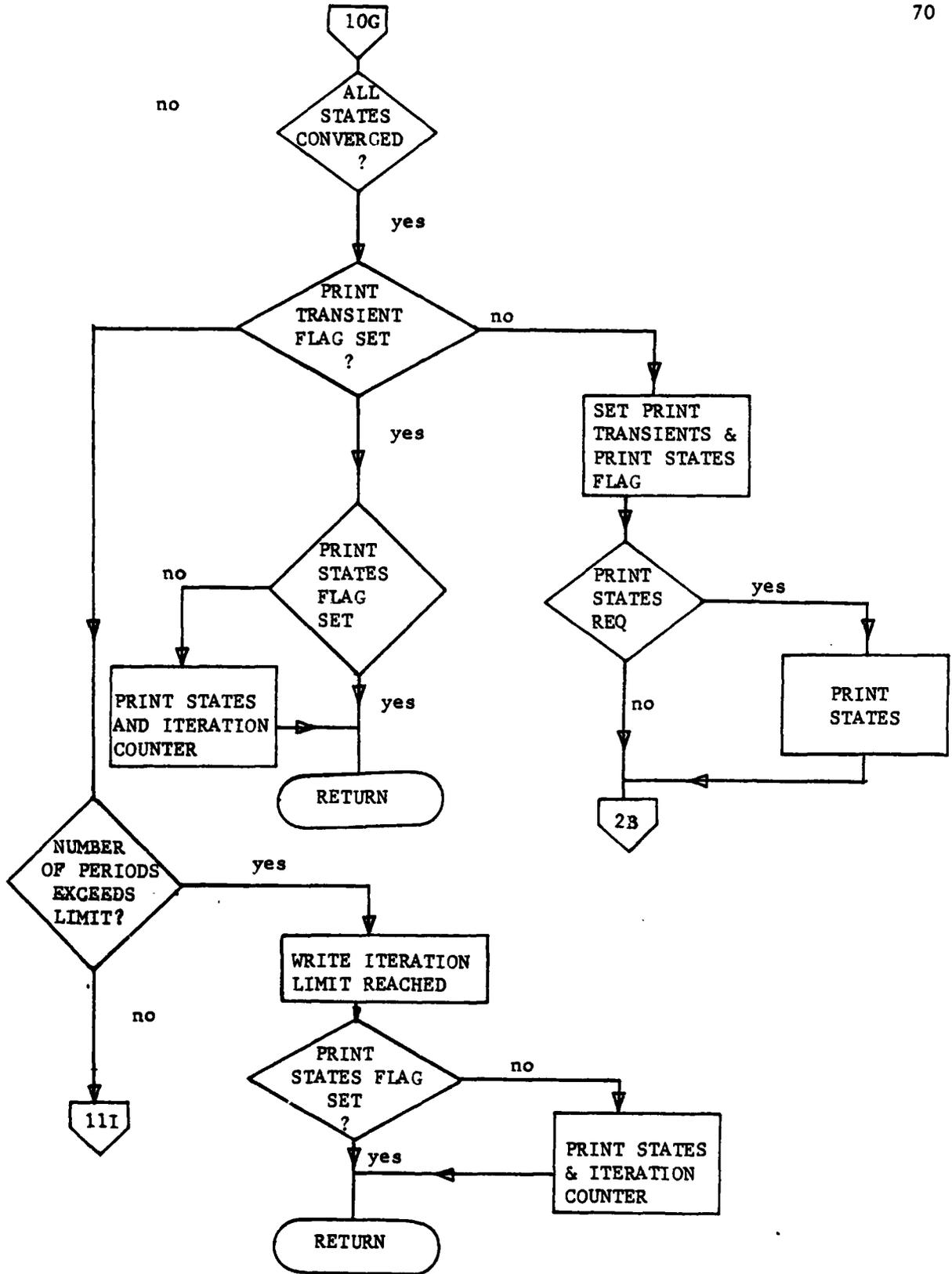


Figure 4.7h "CONTL" Detail Flowchart

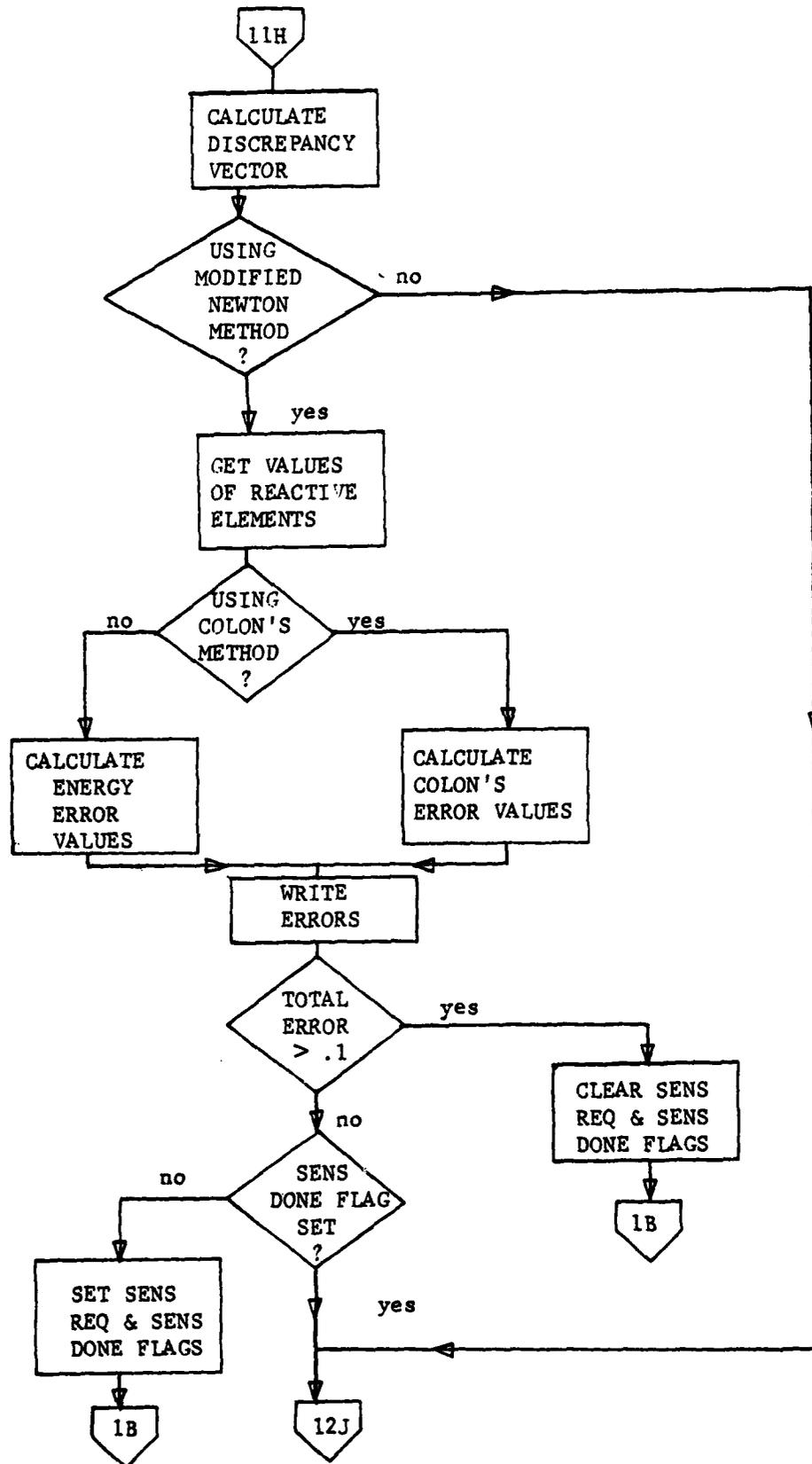


Figure 4.71 "CONTL" Detail Flowchart

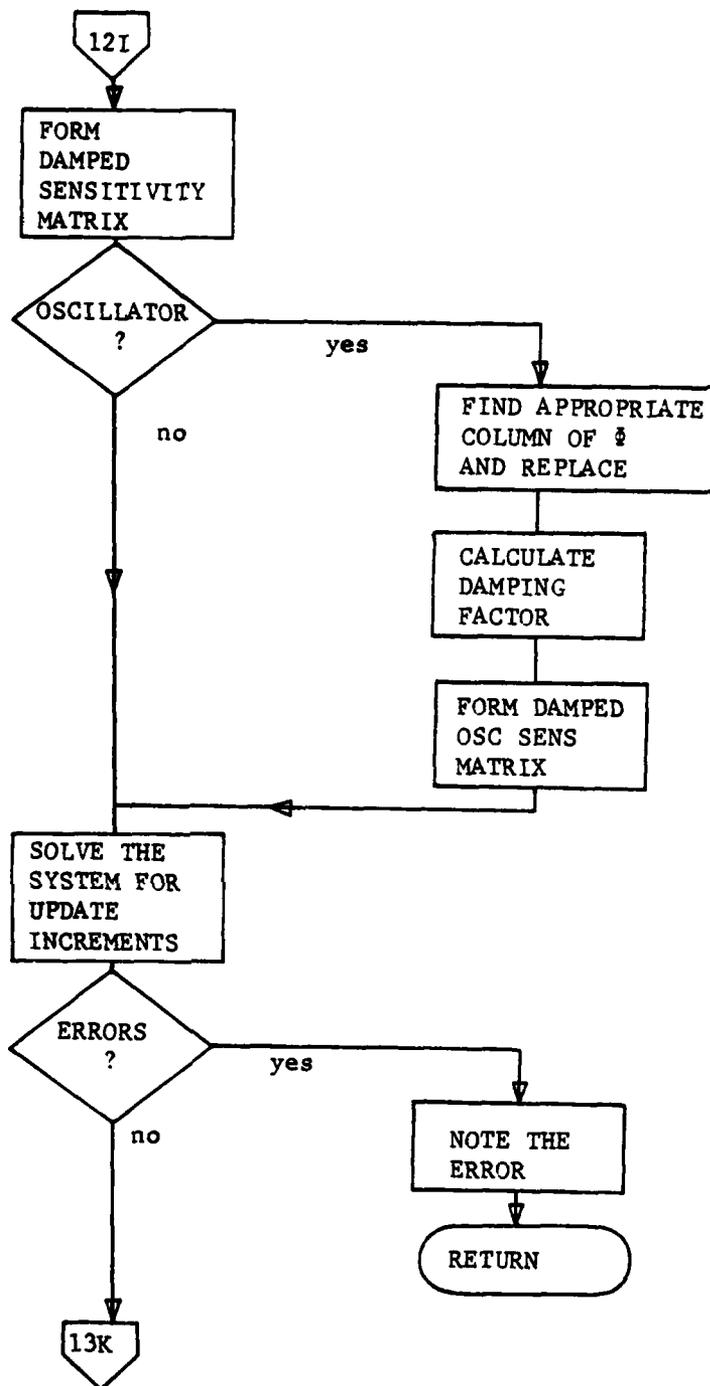


Figure 4.7j "CONTL" Detail Flowchart

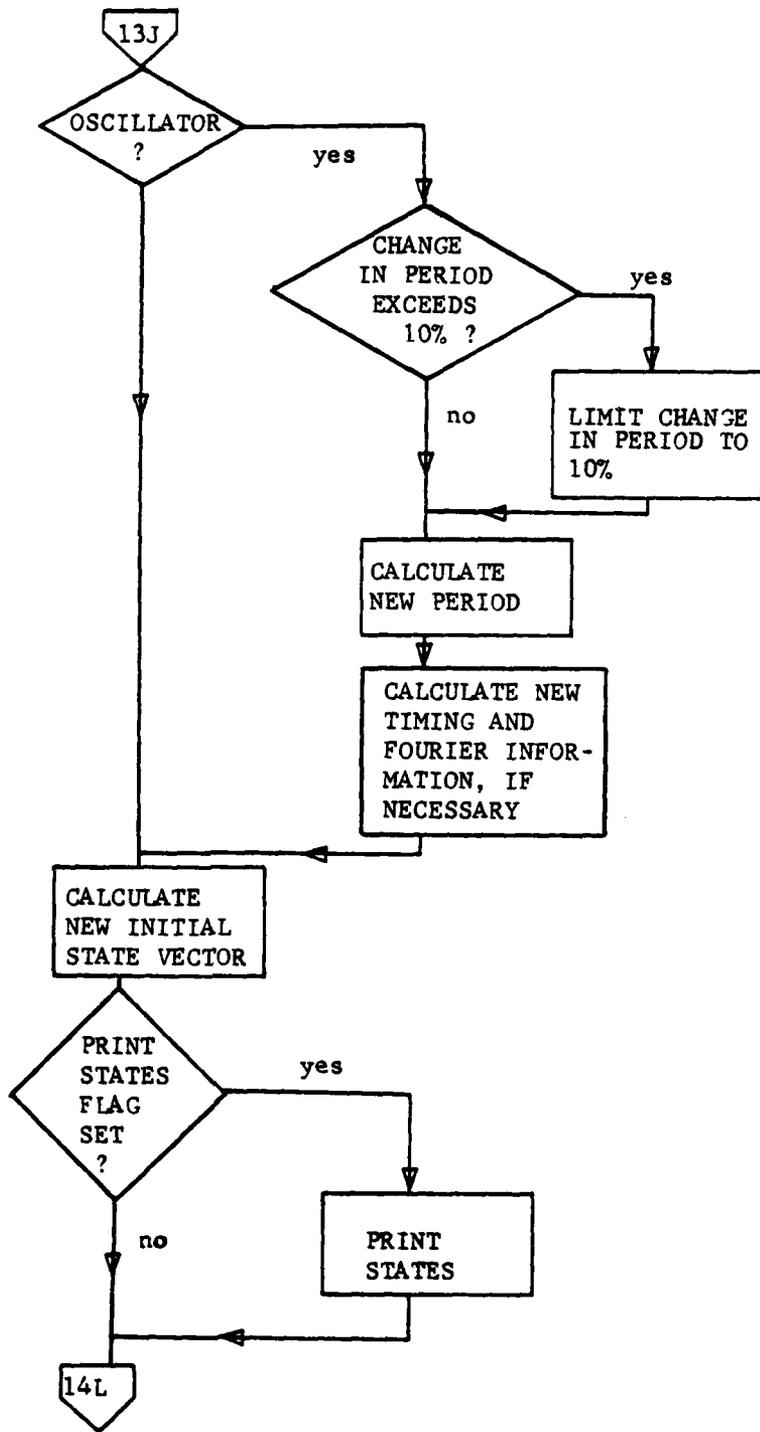


Figure 4.7k "CONTL" Detail Flowchart

introduced, controlled initially by the presence or absence of "PRINT STATES" and "PRINT TRANSIENT ANALYSIS" cards. If these flags (KOF LG and KTTO) are set, the appropriate outputs are printed for each steady state or transient iteration. Otherwise, only the final values are printed.

Timestep size is controlled internally by the program, based on an iteration count of the transient convergence routine. Initially, the user specifies a print interval using either a "TIME" or "STEADY" card; this is automatically the maximum allowable step size. The initial step size chosen by "SINC" is this value divided by 50. After each timestep that converged, the timestep is doubled if convergence was obtained in less than four iterations. Each time convergence is not obtained to the tolerance specified within the allowed number of iterations (the tolerance and number of iterations specified on the "Tolerance" card), the timestep is divided by 8, and the integration restarted from the last convergent point. Any time one step is divided in this fashion seven or more times, a message is output warning of a possible loss of significance in the answers. If one timestep is divided fourteen times, it is assumed that the solution at this point is unstable, a message to that effect is printed, and this analysis is aborted. Finally, if at any time the timestep should become smaller than some minimum value, defined as 10^{-12} times the maximum timestep, the solution is determined to be proceeding too slowly and execution of this analysis is halted.

Control of independent source values is also provided by "CONTL". No changes are necessary for sine wave sources, as the value of these may be calculated for any time by inserting the time into the formula. Pulse and piecewise-linear sources are put into a point-slope form. The timestep

control is also used to cut the timestep if necessary to match breakpoint boundaries. At each breakpoint, the voltage and time are stored, and the slope to the next breakpoint is computed. With this information available it is a simple matter to determine the voltage at any time by computing the time elapsed since the last breakpoint, computing the voltage change over this interval, and adding this difference to the value at the breakpoint. Pulsed sources are handled similarly except that the breakpoints are advanced so that the current time is always within one of the intervals stored.

Control of the sensitivity calculations is by two flags, ISENSF and JSENSF. The first is used to indicate the need for sensitivity calculations during the current period of analysis, and is referred to as the "sensitivity required" flag. The second indicates that we have available a valid matrix of sensitivity values, and is called the "sensitivity done" flag. An analysis of Colon's algorithm, for example, will indicate that by using this second flag a number of steps can be merged, reducing the algorithm from 10 to 7, reducing the length of the required code. The program therefore sets the sensitivity required flag anytime a new calculation of the sensitivity is required; at the conclusion of this period, the sensitivity done flag is set. During contraction mapping periods no sensitivity is required, so the sensitivity required flag is cleared and none is calculated. The requirement is used to set or clear the flag during the Newton period that follows the last contraction mapping period.

Convergence for each timestep is a function only of semiconductor junction voltages, as all other elements are linear. The tolerance is

determined by a "Tolerance" card or the default of 10%; during successive iterations each junction voltage must remain constant within this tolerance times V_c . The actual tolerance checking is done as part of the transient integration subroutine TIMSOL.

If the user requests contraction mapping periods preceding each Newton step, the number requested is stored. Then, following each Newton step a counter is used to determine whether the next step is to be a contraction mapping step or a Newton step. The user is also allowed to specify a larger number of contraction mapping periods before the first Newton step only through the "Steady" card.

Steady state convergence is determined by comparing the difference in initial and final values to the maximum value encountered during the period. This criterion was chosen to avoid apparent convergence failure in switching circuits. For example, the value of current through an inductance at the initial and final times might be nearly zero; let us choose $1\mu\text{A}$. Then our default steady state convergence tolerance will require that these values agree within 10^{-10}A . Yet during other portions of the period the inductor current could exceed 100A . Therefore, the error involved is buried in the roundoff noise of the computer. To avoid this problem the convergence tolerance chosen is based upon maximum state values.

The error values discussed in Chapter 3 are also computed, where appropriate. If a modified Newton method is chosen, the necessary damping factors are also computed. These values are used to compute the damped sensitivity matrix; the identity matrix is added to obtain the modified Jacobian matrix. For oscillators, only a global damping factor, similar to that obtained by Colon, is available.

An understanding of the control subroutine is essential to anyone desiring to change the time domain analysis available in "SINC". All parts of the time domain problem appear in "CONTL" and are affected by it.

CHAPTER 5

EXAMPLES

In order to test the practicality of these methods, a series of examples was constructed and run using the modified SINC program. We will give these results and show how the new techniques act on some typical circuits.

5.1 INTRODUCTION

All problems were run on the Coordinated Science Lab Digital Equipment Corporation DECSYSTEM-1070 in a time-sharing environment. Because part of the system overhead is charged to each user, the runtimes given cannot by themselves be used to determine relative efficiency. In particular, if heavy swapping is taking place, overhead can exceed 10% of total runtime.

We also give the number of Newton steps taken and the total number of full periods over which the system is integrated. Typically, the additional calculations involved for a Newton step requires 30% more time than that for a contraction mapping step. The amount of CPU time required for one full period integration also varies with the region of operation, as this affects the choice of stepsize taken.

Due to programming constraints already in SINC, certain possible changes to increase the efficiency of the modified methods were not included. For example, if an external damping value of zero was supplied for a state, the calculation of sensitivities for the column of \dot{q} associated with that state could be bypassed with an attendant savings in time. This was not done because it was felt that the resulting efficiency would not be worth the cost of restructuring SINC.

Occasionally problems arose due to the step-limiting algorithm used to limit junction voltage changes. In particular, in the transient analysis of switching circuits occasionally the iterates failed to converge at the source transition points.

In the next several sections we have tried to provide a cross-section of example circuits, in order to demonstrate the utility of these modified methods.

5.2 LINEAR CIRCUIT WITH PARASITICS

As an example of the effect of small parasitics on a linear circuit we have chosen the circuit of Figure 5.1.

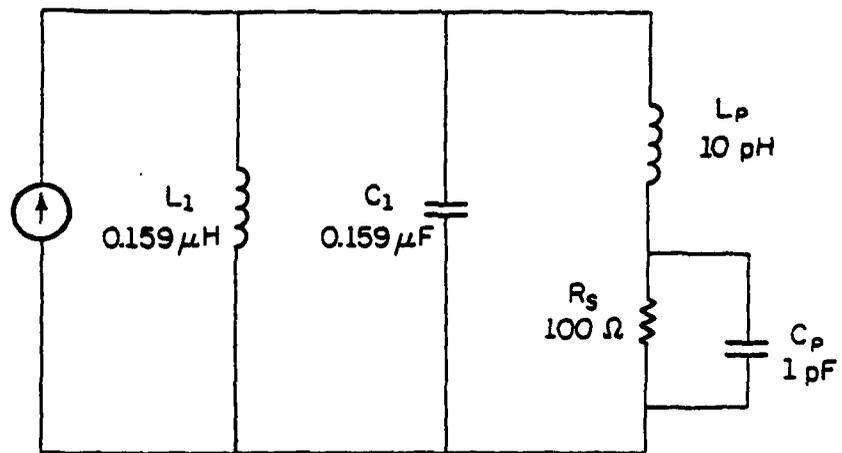
In theory, any linear circuit will converge in one Newton iterate. In practice, two iterations are usually required due to numerical errors; however, the presence of the small parasitic values greatly increases the time required.

Referring to Table 5.1, we note that the contraction mapping solution requires a very large number of full period integrations. Use of the Newton method reduces the required number of iterations by an order of magnitude. However, the presence of the small parasitics introduces errors in the Newton method that still require several iterations to decay.

Next, an interval during which contraction mapping was used prior to the Newton step was inserted. One period of decay did not improve the Newton method; however, two periods provided a marked reduction in the amount of time and number of iterations required.

As a reference point for the damped methods Colon's technique was used. No change was evidenced using this technique. One Newton iteration yields the neighborhood of the solution. At this point, the error due to the parasitics and error in the initial point is so small as to provide a damping term insignificantly smaller than 1.

1 A Pulse
 $T_r = 10 \text{ ns}$
 $T_f = 10 \text{ ns}$
 $P_w = 10 \text{ ns}$
 $PER = 1 \mu\text{s}$



FP-6577

Figure 5.1 Linear Circuit

Table 5.1 Linear Circuit Example Results

METHOD	CPU SECS	NEWTON PERIODS	TOTAL PERIODS
Contraction Mapping	47.3		185
Unmodified Newton	8.3	7	11
1 Decay Period Prior to Newton Step	9.0	6	16
2 Decay Periods Prior to Newton Step	5.8	2	10
Colon's Method	6.8	7	11
Local Vector Damped	17.3	26	30
Product Vector Damped	7.4	7	11
Eliminated Parasitic States Via EXTERNAL Statement	5.0	4	8

For vector damped methods, both the method consisting solely of the local damping term and the product of the local and global terms was used. Use of a local damping vector alone produced a problem requiring a significantly longer analysis. Failure in the early Newton step to take into account the global error led to a significant error in the value of the parasitic states, particularly in the current through the parasitic inductor L_p . Once the other states have converged, this current was essentially isolated by the form of the sensitivity matrix, and proceeded to converge via contraction mapping which required 16 periods of integration.

Use of a damping vector containing both global and local terms removed this difficulty. In the early iterations when all terms were far from the steady state, a global error term provided additional damping and prevented a large initial error which required a long time to decay.

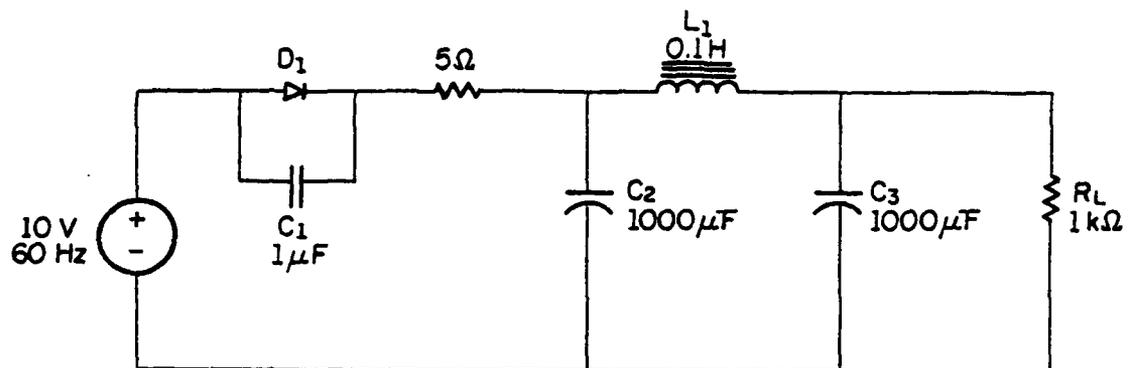
As the final modification for this circuit example, the parasitic elements were eliminated from the Newton method by using an EXTERNAL card containing a damping value of 0 for these states. This showed the largest saving in time, even though maximum efficiency was not used.

Therefore, even a linear circuit for which the Newton method should be most efficient can sometimes benefit from these techniques.

5.3 POWER SUPPLY

The power supply circuit shown in Figure 5.2 is an example used by Aprille [29]. Here the unmodified Newton method is already very successful in reducing the number of periods required to reach the steady state. Again the Newton method provides savings of an order of magnitude over contraction mapping.

Given this performance by the unmodified Newton method, it was difficult to make a significant improvement. Addition of a decay period did not reduce the number of Newton steps required, and therefore, increased the work required by the addition of four CM periods. See Table 5.2.



FP-6578

Figure 5.2 Power Supply Circuit

Table 5.2 Power Supply Example

METHOD	CPU SECS	NEWTON PERIODS	TOTAL PERIODS
Contraction Method	72.9		113
Unmodified Newton	10.8	4	8
Addition of 1 Decay Period	13.0	4	12
Colon's Method	11.2	7	11
Local Vector Damped	12.7	8	12
Product Vector Vamped	12.8	8	12
Elimination of C_1 from Newton Step	9.8	5	9

In this instance all of the damped methods increase the computation required to reach a solution. In cases like this such behavior is not unexpected; if we are proceeding along a trajectory toward a solution a damping method will slow the progress. Ideally, this loss will be minimal in such cases, and will result in increased reliability for other circuits.

Elimination of the capacitor C_1 across the diode did result in a small time savings, even though it required another period. This was due to recomputing the sensitivity matrix less often.

Therefore, some circuits are efficiently handled using an unmodified Newton technique; modified methods, although they do little or no harm, do not significantly benefit these cases.

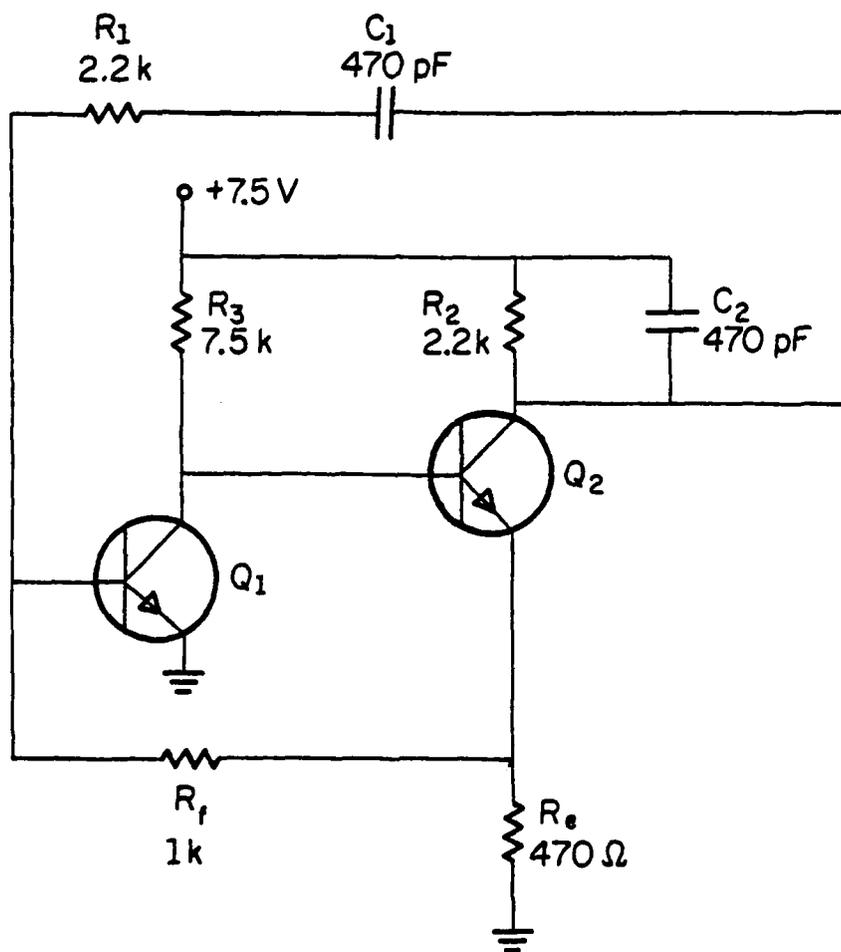
5.4 COLON WIEN BRIDGE

The Wien Bridge oscillator shown in Figure 5.3 was part of an example used by Colon [32]; an error in the value of R_e was subsequently corrected [30].

This example is used to illustrate the use of these modified techniques on autonomous systems. Unfortunately, the transient analysis scheme used in SINC occasionally exhibits convergence problems when semiconductors are connected in feedback configurations; in this case, this problem prevented analysis of the feedback-stabilized bridge circuit.

Without the diode-capacitor feedback branch, the circuit Q was rather low, and the contraction mapping technique required only a few periods to achieve convergence. Even so, the Newton method again reduced the time required for this analysis. See Table 5.3.

The time before the first Newton step was reduced from 3 to 2 to 1 periods. Reducing the number of initial periods to 2 saved one full Newton



FP-6579

Figure 5.3 Wien Bridge Oscillator Circuit

Table 5.3 Wien Bridge Example

METHOD	CPU SECS	NEWTON PERIODS	TOTAL PERIODS
Contraction Mapping	16.6		10
Unmodified Newton - 3 Initial Periods	12.6	2	6
Unmodified Newton - 1 Initial Period	11.6	3	5
Unmodified Newton - 2 Initial Periods	11.0	2	5
1 Decay Period Prior to Newton	15.5	2	8
Colon's Method	12.0	2	6
Local Vector Damped	12.3	2	6
Product Vector Damped	12.3	2	6

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SOME MODIFICATIONS TO NEWTON'S METHOD FOR THE DETERMINATION OF --ETC(U)
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iteration; but a further reduction to a single initial period prior to the first Newton step lost some of the time previously saved by converting a contraction mapping period into a full Newton step.

Addition of a decay period after each Newton period only increased the total number of periods required and the cost of the analysis.

The damped methods required the same number of periods as the full Newton method, but saved some time by requiring fewer evaluations of the sensitivity matrix.

There are no parasitic elements to be eliminated by the use of an external damping vector.

The modified methods are moderately useful with this relatively low-Q oscillator. We might logically wonder about a higher-Q system.

5.5 COLPITTS OSCILLATOR

The Colpitts oscillator circuit given in Figure 5.4 is originally taken from Clarke and Hess [56], pp. 226. This is a moderately high-Q oscillator, with $Q \approx 100$. This circuit has an unstable solution at the DC operating point; if a DC solution is computed first, and then a steady state analysis is requested, oscillations do not build up rapidly enough to prevent the algorithm from indicating convergence on the DC solution. Therefore, the approach used in this case was to do only a DC analysis initially. These values were perturbed approximately 1% and then used as the initial conditions for the steady state analysis. The perturbation was enough to insure that oscillations started rapidly, but not enough to require excessive time to reach a solution.

The high-Q of the circuit is reflected in the time required to reach a steady state solution using contraction mapping techniques. See Table 5.4.

The unmodified Newton method failed to converge in 30 iterations. The addition of 1 or 2 periods of decay after each Newton step failed to help,

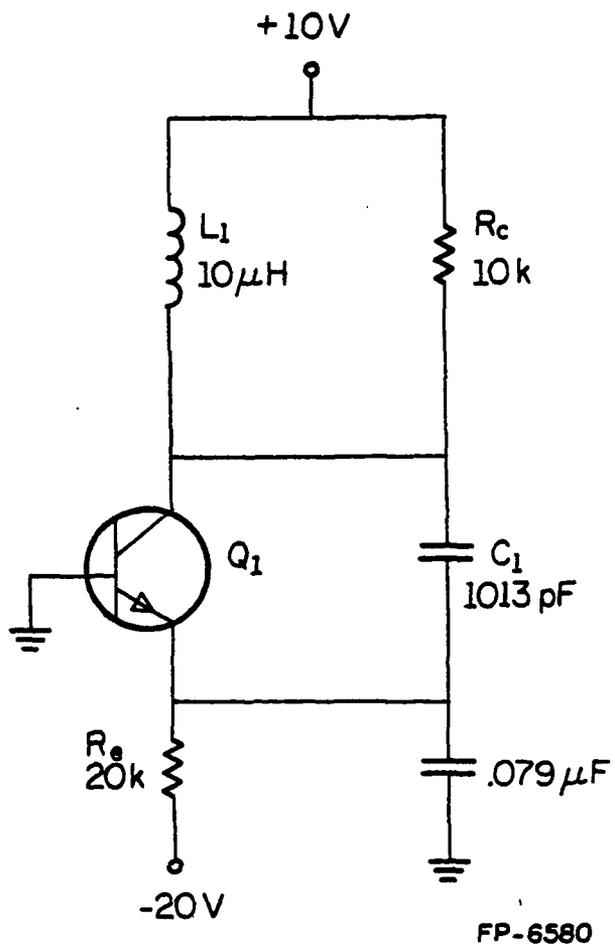


Figure 5.4 Colpitts Oscillator Circuit

Table 5.4 Colpitts Oscillator Example

METHOD	CPU SECS	NEWTON PERIODS	TOTAL PERIODS
Contraction Mapping	43.2		72
Unmodified Newton	Failed to Converge		
1 Decay Period Following Each Newton	Failed to Converge		
2 Decay Periods Following Each Newton	Failed to Converge		
Colon's Method	11.2	8	12
Local Vector Damped	11.9	9	13
Product Vector Damped	14.7	12	16

undoubtedly because of the high Q of the circuit. Here, the period of the oscillator showed a high degree of sensitivity to the state variables. The iteration limit on steady state convergence was 30 periods; however, the results showed no signs of becoming less erratic or tending to converge within a reasonable period.

Use of a damped Newton method alleviated this problem, reducing the excursions in the choice of period. All three damping techniques converged within 12 Newton iterations.

Again, there were no parasitic elements involved for which the use of an EXTERNAL card would be advantageous.

5.6 CLASS-C RF AMPLIFIER

Figure 5.5 shows the circuit of a moderate power Class-C RF Amplifier example used by Vidjkaer [57]. This circuit produces approximately 13 watts of output power from 1.7 watts input at 160 MHz.

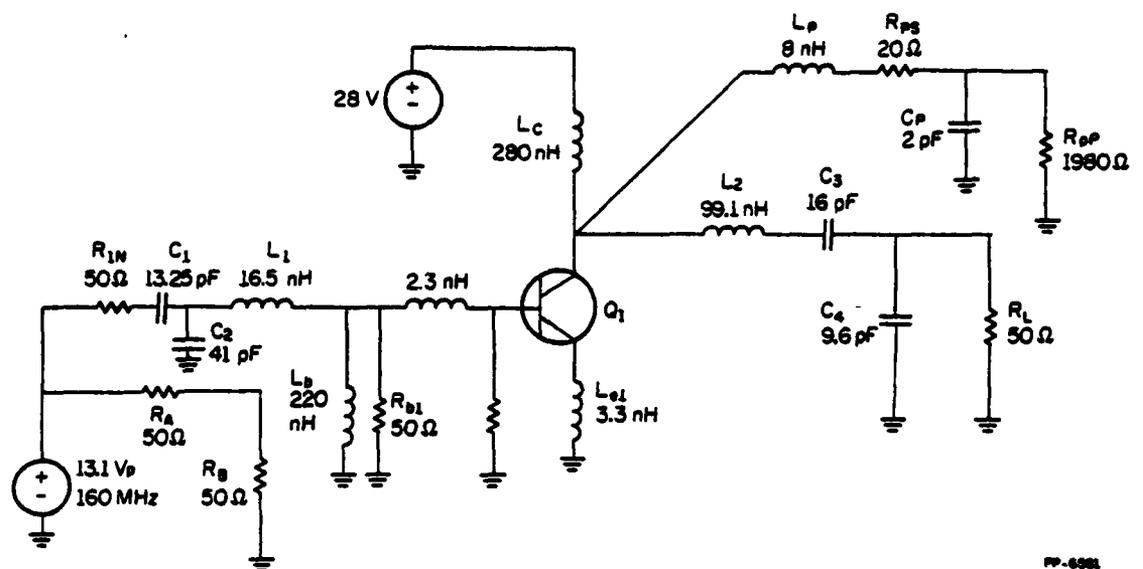
Table 5.5 compares the results of several types of analyses. As usual, continuous integration is the slowest and takes the most time. The unmodified Newton method presents some savings as compared to the contraction mapping approach.

Addition of a period contraction mapping after a full Newton step reduced further the time required for the determination of the steady state.

Colon's scaler damping produced the best time of the damped methods, but all three took approximately the same amount of time, and represented about 25% faster time than the unmodified method.

5.7 CLASS-C 1GHz RF AMPLIFIER

Hajj and Skelboe [58] use the circuit shown in Figure 5.6 as an example for a method involving the application of backward differentiation formulas to piecewise-linear systems.

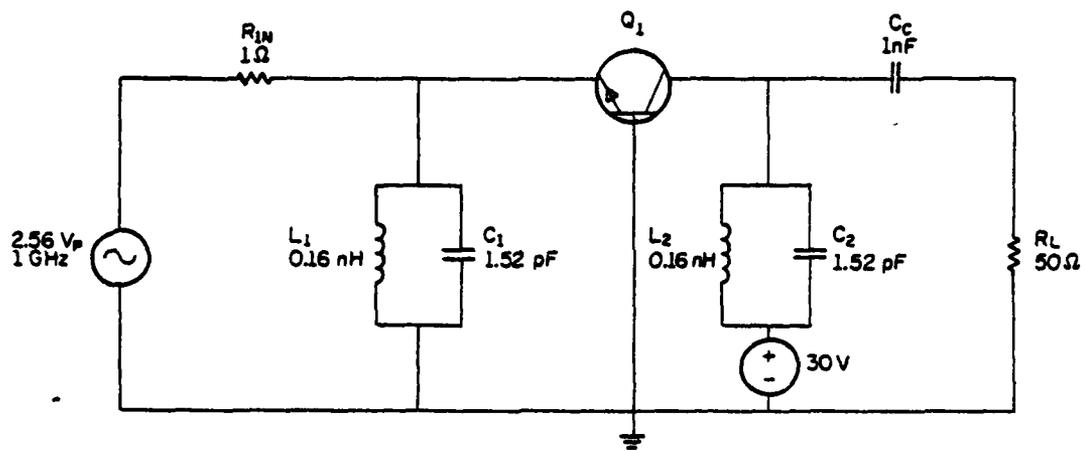


PP-6021

Figure 5.5 Class-C RF Amplifier Circuit

Table 5.5 Class-C RF Amplifier

METHOD	CPU SECS	NEWTON PERIODS	TOTAL PERIODS
Contraction Mapping	45.4		22
Unmodified Newton	40.8	3	7
1 Period of Decay After Each Newton Step	35.7	2	8
Colon's Method	29.0	3	7
Local Vector Damped	31.6	4	8
Product Vector Damped	30.5	4	8



PP-6882

Figure 5.6 1GHz RF Amplifier Circuit

The Newton method again shows a great improvement over the continuous integration approach. However, the unmodified Newton method converges well and the modifications usually require some additional computation or reduce the steps taken by the Newton method, and thus require additional time. The exception to this is the case in which the small parasitic junction capacitances were eliminated using an EXTERNAL card. This analysis required the least computation time of all methods, even though the program was not arranged for maximum efficiency in this case. Results are shown in Table 5.6.

5.8 TRANSISTOR SWITCHING POWER SUPPLY

The DC-DC power converter shown in Figure 5.7 is due to Branin [26,59]. This circuit represents a difficult problem due to its high degree of nonlinearity and large excursions in value.

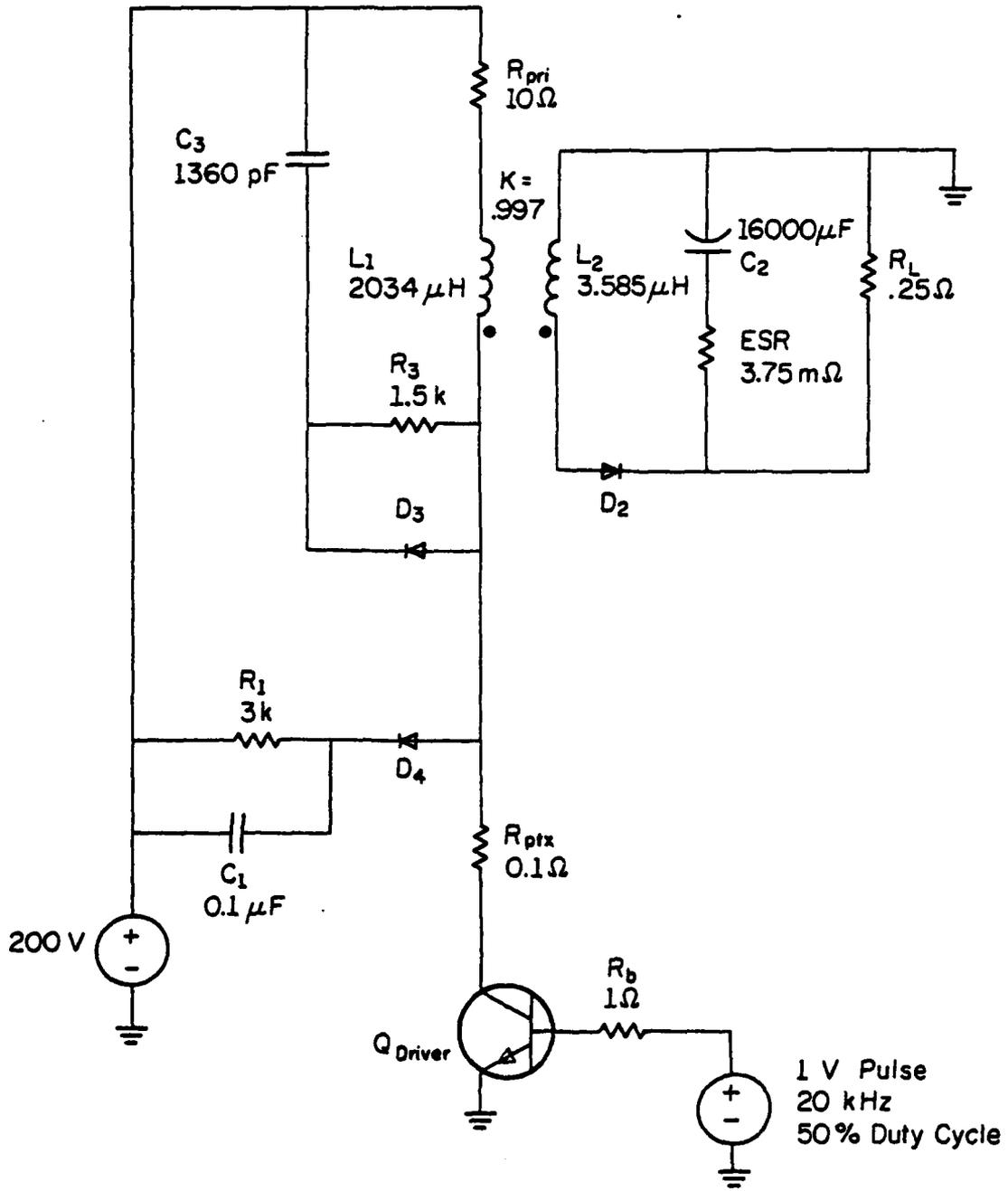
Figure 5.7 is typical of bulk DC-DC power converters. A DC voltage, often obtained by direct rectification of the power line, is chopped by a transistor switch operating at some high frequency. This high frequency allows smaller, more efficient transformers to be used for voltage conversion and isolation. This chopped DC is input to the transformer primary; the output of the secondary winding is rectified and filtered to provide the desired power source. The filter network D_4, R_1, C_1 is used to limit high voltage switching transients in the period between the turnoff of the driver transistor and the conduction of D_2 ; these voltage spikes could otherwise easily exceed the breakdown rating of the switching transistor.

The bypass circuit D_3, R_3, C_3 provides a path for the transformer primary current as the switching transistor turns off, thus minimizing the power dissipated by the switching transistor.

The resistance labeled ESR represents the equivalent series resistance of the filter capacitor C_2 . RPRI is the DC resistance of the transformer primary winding.

Table 5.6 1GHz RF Amplifier

METHOD	CPU SECS	NEWTON PERIODS	TOTAL PERIODS
Contraction Mapping	61.0		52
Unmodified Newton	30.8	8	12
1 Decay Period After Each Newton Step	32.0	6	16
Colon's Method	35.3	14	18
Local Vector Damped	32.5	13	17
Product Vector Damped	37.3	16	20
External Applied Vector Eliminated Transistor Junction Capacitances	25.0	8	12



FP-6583

Figure 5.7 TSR Power Supply Circuit

This supply has a nominal rating of 5 volts at 20 amperes. Due to the extreme nonlinearity of this circuit in operation, we have chosen a convergence tolerance of 1%, instead of the default value of .01%.

The circuit requires in excess of 100 periods of continuous integration to come within this tolerance of its steady state solution, requiring over 8 minutes of CPU time, as shown in Table 5.7.

The unmodified Newton method failed to converge to a solution and showed no tendency to do so in 300 iterations. This circuit exhibited the high sensitivity states noted before. In this case C_1 and C_3 , and to a lesser extent the secondary current I_s , exhibited erratic behavior even close to the solution of the circuit. The values of these states would approach the steady state solution and then veer off wildly, even changing sign.

Use of a single global damping scalar (Colon's method) or a local damping vector did not solve this problem, although they did reduce its severity by reducing the size of the fluctuations.

A damping vector formed by a combination of global and local damping terms did successfully damp this erratic behavior. The ability to use the Newton method at all yielded a great savings in the computer time necessary to reach the steady state.

The addition of a period of contraction mapping following each Newton step yielded the largest savings. Because of the small size of C_1 and C_3 these reach the steady state quickly. The use of a decay period for this purpose produces extremely rapid convergence of the Newton iterates.

An externally supplied damping vector was also used successfully to promote convergence. Elimination of the states associated with C_1 and C_3 provided convergence; but eliminating I_s along with C_1 and C_3 was more successful. A damping vector was also used in which C_1 , C_3 , and I_s were reduced by a factor of 10 in the Newton steps by assigning them damping

Table 5.7 TSR Power Supply

METHOD	CPU SECS	NEWTON PERIODS	TOTAL PERIODS
Contraction Mapping	483.2		112
Unmodified Newton Method	Failed	to Converge	
Colon's Method	Failed	to Converge	
1 Decay Period Following Each Newton Step	52.7	3	10
Local Damping Vector	Failed	to Converge	
Product Damping Vector	75.9	9	13
Eliminate States Associated With C_1 , C_3 , and I_s	58.8	7	11
Eliminate States Associated With C_1 , C_3	68.5	10	14
Reduced C_1 , C_3 , I_s With External Vector	53.2	6	10

factors of 0.1 via an EXTERNAL card. This was more successful than state elimination; on the other hand, it requires more work on the part of the user.

In this case, the modifications allowed successful Newton convergence when it would not otherwise have been possible.

5.9 ANALYSIS OF RESULTS

These methods satisfy the primary law of the physician, "First, do no harm." On the other hand, they are no panacea. They increase reliability, but at the price of increased computational cost in some cases.

In general, they are probably best reserved for those cases in which the unmodified method does not yield the solution in a few iterations. Even if it is determined that some improvement is desirable, some experimentation may be necessary to find the optimum method or combination of methods.

The user has to do the least work in using the damped methods - he simply chooses one of three methods and adds the statement requiring this to the input data.

Use of damping periods requires some work to find the optimum number of periods. If the user has calculated the time constants involved, then a decision could be made as to the number of periods necessary. Also, several experimental runs could be used to determine the optimum number of contraction mapping periods. This experimental approach produces savings only when a large number of runs is contemplated.

Use of the EXTERNAL card requires the most knowledge of the circuit behavior in order to make an intelligent decision as to which states should be reduced or eliminated. A poor choice of external damping vector can inhibit convergence rather than enhance it.

Although not necessary or even helpful in all cases, for many circuits these techniques increase speed and for some circuits they are necessary in order to avoid the high cost of contraction mapping.

CHAPTER 6

CONCLUSION

The usefulness of the modified Newton methods is primarily in their ability to increase the reliability of convergence and to a lesser extent to increase the computational speed. If a method is available which nearly always converges to a steady state solution then the average design engineer would be more willing to accept this tool for his work.

In order to facilitate the use of these techniques for circuit design, some observations will be made on the choice of methods to be employed. Finally, some suggestions for future development will be given.

6.1 GUIDELINES FOR THE SELECTION OF MODIFIED NEWTON METHODS

In this section some observations on the usefulness of the different techniques in dealing with the various circuit examples will be made and some general guidelines offered as to the choice of method for a circuit.

In the case of those circuits for which the unmodified Newton technique produces rapid convergence, it is often counterproductive to spend several computer runs in order to determine the modified Newton technique which yields the best results, as the computer time required for this testing can exceed that required for the analysis required by the designer. An exception to this would be the case for which many analyses are required, as the total improvement in runtime might be significant. A better approach for the case in which many runs of a slightly perturbed circuit is required (e.g., Monte Carlo analysis of component tolerance) might be to start each analysis from the steady state solution of the nominal circuit.

Although it requires the most understanding and work on the part of the user, the use of an EXTERNAL card, when appropriate, often provided the best results. Unfortunately, this method is not universally applicable. It requires a clear determination of states which are parasitic to the circuit operation and which reach the steady state quickly. Some circuits have no such states.

Since the EXTERNAL card can be used in conjunction with other methods, it is often advantageous to include it in any case for which an appropriate set of damping coefficients can be determined. Its use will not interfere with selection of a damped Newton method or the use of periods of contraction mapping.

The addition of periods for the decay of fast transients is useful when these are present. High-Q circuits and others in which no fast transients are present would derive no benefit from the increased cost incurred by requiring more than one full period of integration for each Newton step. If more than a few periods are required for transients to die away, then the method becomes very inefficient. It does not require much effort on the part of the user, however. The effect of this card should be investigated whenever the user has reason to suspect that fast transients are present in the circuit. In non-autonomous systems this is often indicated by a rapid change of state over the period. An unsuccessful attempt was made to use this effect to automate the addition of decay periods.

The use of a damped Newton method is somewhat less certain. In the case of circuits which are well-behaved with regard to steady state convergence, the shorter steps taken by the damped methods delays convergence. On the other hand, the method was successful in cases in which no other methods worked.

The choice of which damped methods is best is also dependent on the circuit. The vector damped method in which each damping term was solely dependent on the error of its associated state proved the least reliable. The problem appears to be due to cross-coupling between states; a state that changed little over the period takes a full Newton step, while a coupled state that underwent a large change undergoes a nearly contraction mapping step; i.e., it takes a large change. This large difference naturally affects the coupled state to which the full Newton step is applied.

Since most circuits exhibit the coupled characteristics described above, some form of global damping is usually desirable. Colon's method uses a single scalar damping coefficient, while the vector damping method employs a function of both the individual state error and the global error. The vector damped method is more appropriate in cases involving widely separated time constants and state values.

The formulation of the error terms and the damping coefficients was chosen experimentally to provide the best match between reliability and speed for a variety of circuit examples. However, the optimum amount of damping is a function of the circuits; for some cases the function chosen may provide more damping than desirable, slowing convergence; in other cases, the iterates might not be damped heavily enough, allowing oscillations and preventing convergence. Because of the form used, the vector damped method is more heavily damped than Colon's method; in the case of well-behaved circuits this is evidenced as an increase in the time required for convergence. In other cases this damping prevents convergence failure.

The final choice of a modified method rests with the user; his judgment will determine whether or not the methods are successful.

6.2 FUTURE WORK

If further use of the SINC program is desired or widespread distribution is to be undertaken, two changes are highly desirable.

The first is that the timestep error control be rewritten. Instead of a convergence iteration count, truncation error control should be undertaken. This will require rewriting large portions of the transient analysis subroutines but should also provide a large increase in reliability and speed. The variable error control of Lourenco-Fernandes and Nichols [34] could be included at this time.

A better method of semiconductor junction voltage iteration should also be incorporated. At the present time this is possibly the weakest point in SINC. In the case of tightly coupled semiconductor junctions this often causes transient convergence to fail. Some method other than brute force application of an absolute limit on junction voltage changes should be implemented.

It would also be desirable to develop a method of relieving the user of some of the decisions about the modifications to be employed. An obvious candidate would be to automate the insertion of decay periods whenever appropriate. Perhaps a better function for the choice of the damping vector could be found which would provide reliable, fast operation on most or all problems; then this could be incorporated as a default and a more specific method chosen only on rare occasions.

An alternative to the major revisions to SINC outlined above would be to incorporate these modifications into a widely used program such as SPICE2. Only when it has been used by design engineers in their daily work with a wide variety of circuits would it become possible to evaluate these methods and determine the best course for future development.

APPENDIX
SINC USER'S MANUAL

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APRIL 1979

THE FOLLOWING DOCUMENTATION DESCRIBES THE CURRENT VERSION AS OF THIS DATE (VERSION 5.6) OF THE PROGRAM 'SINC', AS MODIFIED.

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MARCH 1972

THE FOLLOWING PROGRAM WAS DEVELOPED AT THE UNIVERSITY OF CALIFORNIA, BEKELEY. A GENERAL PROGRAM DESCRIPTION OF A SIMILAR PROGRAM 'TIME' CAN BE FOUND IN THE IEEE JOURNAL OF SOLID STATE CIRCUITS AUGUST 1971 SPECIAL ISSUE ON COMPUTER-AIDED CIRCUIT ANALYSIS AND DEVICE MODELING.

UNIVERSITY OF CALIFORNIA
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DEPARTMENT OF ELECTRICAL ENGINEERING
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FALL 1971

USER GUIDE TO PROGRAM - S I N C -

GENERAL PROGRAM DESCRIPTION

-SINC- IS A SIMULATOR OF NON-LINEAR ELECTRONIC CIRCUITS. THE PROGRAM CALCULATES THE NODE VOLTAGES AND BRANCH CURRENTS AS A FUNCTION OF TIME. INITIAL CONDITIONS ARE DETERMINED AS THE DC OPERATING POINTS AT TIME ZERO. ALLOWED CIRCUIT ELEMENTS ARE BIPOLAR TRANSISTORS, RESISTORS, CAPACITORS, INDUCTORS, CURRENT SOURCES, GROUNDED VOLTAGE SOURCES, JUNCTION FIELD-EFFECT TRANSISTORS (JFETS), JUNCTION DIODES, AND TRANSFORMERS.

PROGRAM LIMITATIONS AND SPECIAL FEATURES

CIRCUIT SIZE -
100 USER EXPLICITLY DEFINED NODES INCLUDING DATUM NODE TO BE NUMBERED 0 (ZERO).
200 NODES WITH INCLUSION OF TRANSISTOR MODEL CONTAINING RB AND/OR RC, JFET AND DIODE INTERNAL NODES. NO CONTROLLED SOURCES ARE ALLOWED.

LIMITS -

100 BIPOLAR TRANSISTORS
30 SETS OF BIPOLAR MODEL SPECIFICATIONS
100 R, L, AND C ELEMENTS
10 INDEPENDENT CURRENT SOURCES
10 GROUNDED INDEPENDENT VOLTAGE SOURCES
5 TRANSFORMERS
10 JUNCTION DIODES
10 DIODE MODELS
10 JFETS
5 JFET MODELS

TRANSIENT ANALYSIS OUTPUT -

9 OUTPUT SPECIFICATIONS TO BE PRINTED
6 OUTPUT PLOTS WITH UP TO 2 WAVEFORMS PER PLOT

DC ANALYSIS -

THE PROGRAM GIVES ALL DC NODE VOLTAGES AT TIME ZERO. SEMICONDUCTOR OPERATING CONDITIONS ARE ALSO GIVEN, IF NO 'NDC' CARD IS INCLUDED (SEE 'NDC' CARD, BELOW). DURING DC ANALYSIS, ALL CAPACITORS ARE OPEN-CIRCUITED, AND ALL INDUCTANCES ARE REPLACED BY 1 OHM RESISTANCES. NO COUPLING EXISTS BETWEEN TRANSFORMER WINDINGS.

ANALYSES AT DIFFERENT TEMPERATURES -

A CIRCUIT CAN BE ANALYZED AT UP TO FIVE TEMPERATURES (USE THE TEMP CARD, SEE BELOW). TEMPERATURE COEFFICIENTS CAN BE SPECIFIED FOR ELEMENTS AND TRANSISTOR MODEL PARAMETERS.

TIME STEP CONTROL -

PROGRAM CONTROL OF THE TIME STEP IS BY AN ITERATION COUNT. THE USER SPECIFIES THE PRINT INTERVAL/MINIMUM TIMESTEP BY A 'TIME' OR 'STEADY' CARD. HOWEVER, THE ACTUAL NUMBER OF STEPS USED BETWEEN PRINT INTERVALS IS UNDER PROGRAM CONTROL AND IS A FUNCTION OF THE TOLERANCE ON CONVERGENCE OF THE TRANSIENT ROUTINE AND THE SYSTEM BEING EVALUATED. A WARNING OR ERROR MESSAGE IS PRINTED SHOULD THE TIMESTEP BECOME TOO SMALL; IN SOME CASES THE PROGRAM WILL TERMINATE EXECUTION TO AVOID EXCESSIVE USE OF COMPUTER TIME.

INPUT DATA CARD FORMAT

THE INPUT IS IN A SEMI-FREE FORMAT. ANY KEY CHARACTER OR WORD MUST BEGIN IN COL. 1. ALL DATA FIELDS ARE SEPARATED BY ONE OR MORE DELIMITERS (EXCEPT THE MODEL CARD) (DELIMITERS ARE BLANKS, COMMA, LEFT AND RIGHT PARENTHESES). ALL FIELDS ENCLOSED IN [] ARE OPTIONAL; HOWEVER, TO PRESERVE ORDER NO FIELDS PRECEEDING THE LAST SPECIFIED CAN BE SKIPPED. THUS, IF A CARD CONTAINED THREE OPTIONAL FIELDS AND WE WISHED TO SPECIFY THE VALUE OF THE LAST FIELD, EXPLICIT VALUES FOR THE FIRST TWO FIELDS MUST BE GIVEN AS WELL.

\$ TITLE CARD -- THE FIRST CARD OF THE INPUT CARD DECK. COLS. 1-4 MUST CONTAIN THE LETTERS SINC, COLS. 5-72 COMMENTS. FOR STUDENT JOBS, THE STUDENT CLASS AND NAME SHOULD APPEAR.

\$ COMMENT CARD -- THE GENERAL FORM IS

* COMMENTS

COL. 1 MUST CONTAIN THE CHARACTER * (ASTERISK).

\$ TIME SPECIFICATION -- THE GENERAL FORM IS

TIME DELT TSTP [TSTR]
OR
TRAN DELT TSTP [TSTR]

WHERE DELT IS THE TIME INTERVAL AT WHICH THE OUTPUTS ARE PRINTED, AND
TSTP IS THE TOTAL ANALYSIS TIME.
TSTR IS THE START TIME, IF NOT ZERO.

NOTE: IF TSTR IS NOT ZERO, THE CIRCUIT IS ANALYZED FROM ZERO TO TSTR, BUT NO OUTPUT IS GENERATED; OUTPUT IS GENERATED FROM TSTR TO TSTP.

NOTE: IF A STEADY STATE ANALYSIS IS REQUESTED (SEE STEADY CARD, BELOW), THOSE VALUES OVERRIDE THE VALUES SPECIFIED ON THE TIME CARD. THE START TIME WILL BE 0.0, THE STOP TIME THE PERIOD, AND THE PRINT INTERVAL WILL BE COMPUTED AS DPRT=PERIOD/NST. IN GENERAL, A TIME CARD SHOULD NOT BE PRESENT IN A DECK CONTAINING A STEADY CARD.

EXAMPLE

TIME .5US 100US

\$ PIECEWISE LINEAR SOURCE SPECIFICATION -- THE GENERAL FORM IS

```
VXXX N1 N2 V1 [T1 V2 [T2 V3 [ ... VI ]]]
IXXX N1 N2 I1 [T1 I2 [T2 I3 [ ... II ]]]
```

XXX REPRESENTS AN OPTIONAL NAME OF ONE TO FOUR CHARACTERS.

(VSINX AND ISINX ARE RESERVED FOR SINUSOIDAL SOURCES; VPLSX AND IPLSX FOR PULSED SOURCES.)

WHERE N1 AND N2 ARE THE POSITIVE AND NEGATIVE NODES RESPECTIVELY, AND VI AND TI ARE THE VOLTAGE AND TIME AT THE ITH BREAKPOINT. II AND TI ARE THE CURRENT AND TIME AT THE ITH BREAKPOINT, RESPECTIVELY.

FOR VOLTAGE SOURCES N2 MUST BE 0 (ZERO).

A MAXIMUM OF 9 BREAKPOINTS ARE ALLOWED FOR EACH SOURCE. LINEAR INTERPOLATION IS USED BETWEEN SPECIFIED POINTS. T1 MUST BE ZERO. FOR A CONSTANT SOURCE, SPECIFY V1 OR I1 AND LEAVE THE REST BLANK.

EXAMPLE

```
VIN 2 0 -5.0 0.0 -5.0 1.25MS +5.0 1.50MS +5.0
VCC 9 0 10V
```

THE FIRST SPECIFICATION HAS A VALUE OF -5.0 V FROM TIME 0 UNTIL 1.25 MILLISECONDS, THEN RISES LINEARLY TO +5.0 V AT 1.5 MILLISECONDS, WHERE IT REMAINS UNTIL THE FINAL TIME SPECIFIED; THE SECOND SOURCE HAS A DC VALUE OF 10 VOLTS.

NOTE: THE MINIMUM TRANSITION TIME BETWEEN ANY ADJOINING PAIR OF BREAKPOINTS IS DPRT, THE PRINT INTERVAL. DPRT IS 'DELT' SPECIFIED ON THE TIME CARD, OR IS COMPUTED AS DPRT=PERIOD/NST IF A STEADY STATE ANALYSIS IS REQUESTED.

\$ SINUSOIDAL SOURCE SPECIFICATION -- THE GENERAL FORM IS

```
VSINX N1 N2 VM F [PHI] [DC]
ISINX N1 N2 IM F [PHI]
```

WHERE X IS AN ARBITRARY ALPHANUMERIC CHARACTER,
VM OR IM IS THE AMPLITUDE OF THE SOURCE,
F IS THE FREQUENCY OF THE SOURCE IN HZ,
PHI IS THE PHASE ANGLE IN DEG.,
DC IS THE DC LEVEL (VOLTAGE SOURCES ONLY).

THE SOURCE VALUE AT TIME T IS CALCULATED ACCORDING TO

$$V(T) = VM * \sin(2 * \pi * (F * T + PHI / 360)) + DC$$

OR

$$I(T) = IM * \sin(2 * \pi * (F * T + PHI / 360))$$

EXAMPLE

```
VSIN 2 0 10M 100KHZ 0
VSIN 2 0 0.5V 10MEHZ 0.0 1.5V
ISIN 3 5 1M 2MEGHZ 0
```

NOTE: THE DC VALUE FOR VSIN SOURCES IS USED FOR THE DC ANALYSIS; FOR ISIN SOURCES THE VALUE IS ZERO, NOT $IM * \sin(2 * \pi * (F * T + PHI / 360))$.

\$ PULSED SOURCE SPECIFICATION -- THE GENERAL FORM IS

VPLSX N1 N2 VOFF VON TD TR TF PW PER

IPLSX N1 N2 IOFF ION TD TR TF PW PER

WHERE X IS AN OPTIONAL CHARACTER IN THE NAME

N1 AND N2 ARE THE POSITIVE AND NEGATIVE NODES, RESPECTIVELY

VOFF (IOFF) AND VON (ION) ARE THE OFF AND ON VALUES

TD IS THE DELAY TIME

TR IS THE RISE TIME

TF IS THE FALL TIME

PW IS THE PULSE WIDTH

PER IS THE PERIOD

THE VOLTAGE (CURRENT) IS CALCULATED FROM A SERIES OF LINEAR SEGMENTS; THE SOURCE HAS A VALUE VOFF FROM 0.0 TO TD, VON FROM TD+TR UNTIL TD+TR+PW, A VALUE VOFF FROM TD+TR+PW+TF UNTIL PER, AND REPEAT EVERY PER SECONDS. DURING THE RISE AND FALL TIMES A LINEAR FIT IS USED. FOR DC CALCULATIONS THE OFF VALUE IS USED.

EXAMPLE

VPLS1 1 0 0.0 1.0 1.0US 0.1US 0.3US 5.0US 20.0US

IPLSA 3 5 -1.0 3.5 5.0MS 0.0 0.0 5.0MS 10.0MS

\$ TEMPERATURE SPECIFICATION -- THE GENERAL FORM IS

TEMP T1 [T2] [T3]

WHERE T1, T2, ... ARE THE TEMPERATURES AT WHICH THE ANALYSIS ARE TO BE PERFORMED, IN UNITS OF DEG K.

UP TO A TOTAL OF 5 (FIVE) TEMPERATURES CAN BE SPECIFIED. T1 IS TAKEN AS THE NOMINAL TEMPERATURE. IF THIS CARD IS ABSENT THE TEMPERATURE IS ASSUMED TO BE 300 DEG K.

EXAMPLE

TEMP 300 295 315

\$ R, L, AND C ELEMENT SPECIFICATION -- THE GENERAL FORM IS

RXXX N1 N2 VALUE [TC1 [TC2]]
 LXXX N1 N2 VALUE [TC1 [ICN]]
 CXXX N1 N2 VALUE [TC1 [ICN]]

WHERE VALUE IS THE ELEMENT VALUE,

TC1 IS THE 1ST ORDER TEMPERATURE COEFFICIENT OF THE ELEMENT VALUE, IN PARTS PER DEGREE KELVIN; FOR RESISTORS, TC2 IS THE 2ND ORDER TEMPERATURE COEFFICIENT, AND FOR CAPACITORS AND INDUCTORS ICN IS THE INITIAL VOLTAGE AND CURRENT, RESPECTIVELY.

EXAMPLE

R1 1 2 100K
 CS 5 0 2.3P
 RB 2 3 152OHMS
 RC 2 6 5K .002 5.E-5

\$ MUTUAL INDUCTANCE SPECIFICATIONS -- THE GENERAL FORM IS

UXXX N1 N2 N3 N4 L11 L22 K [IP] [IS]

WHERE UXXX IS THE TRANSFORMER NAME,

N1 AND N2 ARE THE POSITIVE AND NEGATIVE NODES OF THE PRIMARY,
 N3 AND N4 ARE THE POSITIVE AND NEGATIVE NODES OF THE SECONDARY,
 L11 IS THE SELF-INDUCTANCE OF THE PRIMARY,
 L22 IS THE SELF-INDUCTANCE OF THE SECONDARY,
 K IS THE COEFFICIENT OF COUPLING,
 IP IS THE INITIAL CONDITION (CURRENT) FOR THE PRIMARY, AND
 IS IS THE INITIAL CONDITION (CURRENT) FOR THE SECONDARY.

\$ BIPOLAR TRANSISTOR SPECIFICATION -- THE GENERAL FORM IS

QXXX BXXX NC NB NE

WHERE BXXX IS THE MODEL NAME,

NC NB NE ARE THE COLLECTOR, BASE AND EMITTER NODES RESPECTIVELY.

THE MODEL BXXX IS TO BE DEFINED SEPARATELY AND CAN BE REFERENCED BY OTHER TRANSISTORS.

EXAMPLE

QA B1 2 3 4
Q12 BPNP 1 3 6

\$ BIPOLAR MODEL SPECIFICATION -- THE MODEL CAN BE SPECIFIED IN FULL AS FOLLOWS -

BXXX TYPE BF=BFMAX,ICMAX,BF,IC,VCE,TC1,TC2 BR=VALUE ISS=VALUE
+ RB=VALUE,TC1,TC2 RC=VALUE,TC1,TC2 RO=VALUE,IC FT=VALUE,IC,VCE
+ TSAT=VALUE CJE=VALUE,VBE,PHIE,NE CJC=VALUE,VBC,PHIC,NC,RATIO
+ CSUB=VALUE TEMP=VALUE

THE PARAMETERS CAN APPEAR AS A GROUP IN ANY ORDER. ANY PARAMETER CAN BE OMITTED. THE VALUE SPECIFICATION TO THE RIGHT OF THE PARAMETER KEY WORD SHOULD BE PUT IN THE ORDER INDICATED (SEE EXPLANATION BELOW). OMITTING LEADING VALUE SHOULD BE INDICATED BY A SLASH (/). A CONTINUATION CARD SHOULD HAVE A PLUS (+) IN COLUMN 1.

THE MODEL PARAMETERS HAVE THE MEANING AND DEFAULT VALUES AS FOLLOWS -

TYPE	NPN OR PNP	NPN
BF=BFMAX	MAXIMUM FORWARD BETA	100.
ICMAX	IC AT WHICH BFMAX OCCURS	IGNORED
BF	SOME VALUE OF BETA AT IC<ICMAX	IGNORED
IC	IC FOR ABOVE VALUE OF BETA	IGNORED
VCE	VCE FOR ABOVE VALUE OF BETA	IGNORED
TC1	1ST ORDER TEMP.	0.
TC2	2ND ORDER TEMP. COEFF.	0.
BR=VALUE	REVERSE BETA	1.
ISS=VALUE	SATURATION CURRENT	1.E-14
RB=VALUE	BASE RESISTANCE	0.
TC1	1ST ORDER TC	0.
RC=VALUE	COLLECTOR RESISTANCE	0.
RO=VALUE	COMMON-EMITTER OUTPUT RESISTANCE	INFINITE
IC	IC FOR ABOVE VALUE	1.E-3
FT=VALUE	SMALL SIGNAL UNITY GAIN FREQ.	LARGE
IC	IC FOR ABOVE VALUE	IGNORED
VCE	VCE FOR ABOVE VALUE	IGNORED
TSAT=VALUE	SATURATION TIME CONSTANT	0.

CJE=VALUE	BE JUNCTION CAPACITANCE	0.
VBE	VBE FOR ABOVE VALUE	0.
PHIE	BE JUNCTION POTENTIAL	.7
NE	GRADIENT FACTOR	.33333
CJC=VALUE	BC JUNCTION CAPACITANCE	0.
VBC	VBC FOR ABOVE VALUE	0.
PHIC	BC JUNCTION POTENTIAL	.5
NC	GRADIENT FACTOR	.33333
RATIO	SPLIT RATIO	0.
CSUB=VALUE	SUBSTRATE CAPACITANCE	0.
TEMP=VALUE	PARAMETER DEFINITION TEMPERATURE	300.

THE MODEL USED IS THE FAMILIAR EBERS-MOLL MODEL, WITH THE INCLUSION OF BASE AND COLLECTOR OHMIC RESISTANCES, JUNCTION CAPACITANCES, AND AN OUTPUT CONDUCTANCE MODELING THE EARLY EFFECT (SEE J. J. EBERS AND J. C. MOLL, "LARGE SIGNAL BEHAVIOR OF JUNCTION TRANSISTORS", PROC. IRE, V 42, DEC. 1954).

EXAMPLE

B1 BF=120,1M,2,1E-6,5 FT=350MEGHZ CJE=2.5P/.8 TSAT=5NS
 BA BF=120 BR=0.5 FT=600MEG TSAT=5NS CJE=1.5P CJC=2P
 + RB=250 RC=15
 BPNP PNP BF=5 BR=1 FT=25MEGHZ CJE=3P CJC=4P
 BDC BF=120 BR=2 RB=250 RC=15
 BSML ISS=1.E-15 RB=1K RC=100 TEMP=295

\$ JFET TRANSISTOR SPECIFICATION -- THE GENERAL FORM IS

JXXX XXXX ND NG NS
OR
JXXX ND NG NS XXXX

WHERE JXXX IS THE TRANSISTOR NAME
XXXX IS THE MODEL NAME (FIRST LETTER MUST BE 'X')
ND NG NS ARE THE DRAIN, GATE, AND SOURCE NODES RESPECTIVELY.

THE MODEL XXXX IS DEFINED SEPARATELY, AND MAY BE REFERENCED BY OTHER TRANSISTORS.

\$ JFET MODEL SPECIFICATIONS -- THE MODEL CAN BE FULLY SPECIFIED AS FOLLOWS -
XXXX CTYP=TYPE CODE VTO=VALUE BTA=VALUE LMDA=VALUE RD=VALUE RS=VALUE
+ CGS=VALUE CGD=VALUE PB=VALUE IS=VALUE

THE MODEL PARAMETERS HAVE MEANING AND DEFAULT VALUES AS FOLLOWS -

CTYP	TYPE CODE (1 FOR N-CHANNEL, 2 FOR P-CHANNEL)	N-CHANNEL
VTO	PINCHOFF VOLTAGE	0.0
BTA	TRANSCONDUCTANCE	1.0E-4
LMDA	CHANNEL-LENGTH MODULATION PARAMETER	0.0
RD	DRAIN OHMIC RESISTANCE	0.0
RS	SOURCE OHMIC RESISTANCE	0.0
CGS	ZERO-BIAS GATE-SOURCE CAPACITANCE	0.0
CGD	ZERO-BIAS GATE-DRAIN CAPACITANCE	0.0
PB	GATE JUNCTION POTENTIAL	1.0
IS	GATE JUNCTION SATURATION CURRENT	1.0E-14

NOTE: RATHER THAN SPECIFY CTYP=1 OR CTYP=2, WE CAN USE NJF OR PJF (NOT CTYP=NJF). THESE ARE EQUIVALENT, BUT ARE MORE EASILY RECOGNIZED.

THE JFET MODEL USED IS BASED ON THAT OF SHICHMAN AND HODGES (SEE H. SHICHMAN AND D. A. HODGES, "MODELING AND SIMULATION OF INSULATED-GATE FIELD-EFFECT TRANSISTOR CIRCUITS", IEEE JSSC, V SC-3, P285-289).

EXAMPLE

XJFET NJF VTO=1.0 RD=10 RS=10

\$ DIODE SPECIFICATION -- THE GENERAL FORM IS

DXXX ZXXX NA NC
OR
DXXX NA NC ZXXX

WHERE DXXX IS THE DIODE NAME
ZXXX IS THE MODEL NAME
NA AND NC ARE THE ANODE AND CATHODE NODES, RESPECTIVELY.

THE MODEL ZXXX IS DEFINED SEPARATELY AND CAN BE REFERENCED BY OTHER DIODES.

\$ DIODE MODEL SPECIFICATIONS -- THE MODEL CAN BE FULLY SPECIFIED AS -

ZXXX IS=ISAT RS=RS N=VALUE TT=VALUE CJO=VALUE PHIB=VALUE M=VALUE

THE PARAMETERS APPEAR IN ANY ORDER. ANY OR ALL CAN BE OMITTED.
CONTINUATION CARDS MUST HAVE A '+' IN COL. 1.

MODEL PARAMETER MEANINGS AND DEFAULT VALUES ARE AS FOLLOWS-

IS	SATURATION CURRENT	1.0E-14
RS	OHMIC RESISTANCE	0.0
N	EMISSION COEFFICIENT	1.0
TT	TRANSIT TIME	0.0
CJO	ZERO-BIAS JUNCTION CAPACITANCE	0.0
PB	JUNCTION POTENTIAL	1.0
M	GRADING COEFFICIENT	0.5

THE DIODE MODEL USED IS BASED ON THAT OF SPICE, AND MAY BE USED FOR BOTH JUNCTION AND SCHOTTKY-BARRIER DIODES. (SEE MEMORANDUM ERL-M520, 9 MAY 1975, UNIVERSITY OF CALIFORNIA, BERKELEY).

EXAMPLES

ZPWR IS=1.0E-10 RS=0.05

\$ PLOT SPECIFICATION -- THE GENERAL FORM IS -

PLOT MIN MAX YXXX N1 N2 YXXX N1 N2
Y IS EITHER V OR I.

WHERE MIN AND MAX ARE THE COORDINATE SPECIFICATIONS OF THE PLOT, IN VOLTS FOR VOLTAGES AND MILLIAMPERES FOR CURRENTS.

NOTE THAT CURRENT OUTPUTS REQUIRE THAT A UNIQUE BRANCH (R,L,C,TRANSFORMER PRIMARY OR SECONDARY,CURRENT OR VOLTAGE SOURCE) EXIST BETWEEN THE NODES SPECIFIED. IF TWO OR MORE BRANCHES ARE CONNECTED BETWEEN THE SPECIFIED NODES, ONLY THE CURRENT FOR THE FIRST BRANCH ELEMENT ENCOUNTERED IN A SEARCH WILL BE GIVEN, POSSIBLY LEADING TO ERRORS.

THE MAXIMUM NUMBER OF POINTS TO BE PLOTTED IS LIMITED TO THE FIRST 150 TIME INTERVALS AS SPECIFIED ON THE TIME CARD. UP TO TWO OUTPUTS CAN BE PRINTED PER PLOT.

EXAMPLE

PLOT -5. 5. VOUT 10 0 VC2 3 0
PLOT 0 5 VOUT 3 5
PLOT 0 1. IRS 1 2

\$ PRINT SPECIFICATION -- THE GENERAL FORM IS

PRINT YXXX N1 N2 YXXX N3 N4
Y IS EITHER V OR I

WHERE YXXX IS THE NAME OF THE VALUE SOUGHT
NA AND NB ARE THE POSITIVE AND NEGATIVE NODES, RESPECTIVELY.
VOLTAGES ARE IN VOLTS, CURRENTS IN MILLIAMPERES. UP TO NINE PARAMETERS CAN BE SPECIFIED.

EQUIVALENT RESULTS CAN BE OBTAINED BY LISTING OUTPUT REQUIREMENTS ON SEPERATE LINES IN THE FORM -

VOUXX N1 N2
OR
IOUXX N1 N2

WHERE XX ARE TWO OPTIONAL CHARACTERS
N1 AND N2 ARE THE POSITIVE AND NEGATIVE NODES, RESPECTIVELY.

VOUT 1 3 IS EXACTLY EQUIVALENT TO PRINT VOUT 1 3

EXAMPLES

PRINT VIN 1 0 VOUT 8 0
PRINT V2 2 0 IDRAIN 4 5
VOUT8 8 0

\$ FOURIER ANALYSIS SPECIFICATION --

THE CURRENT VERSION OF SINC ALLOWS THE USER A CHOICE OF TWO METHODS OF OBTAINING THE SPECTRUM OF A WAVEFORM; USING EITHER THE FFT OR STANDARD FOURIER ANALYSIS.

FOR FOURIER ANALYSIS AT AN ARBITRARY NUMBER OF POINTS, THE FORM IS-
FOR TSTR TSTP NHAR

WHERE TSTR AND TSTP ARE THE START AND STOP TIMES FOR THE PERIOD TO BE ANALYZED (THE PERIOD), AND NHAR IS THE NUMBER OF HARMONICS DESIRED.

NHAR MUST BE LESS THAN OR EQUAL TO 49. FOURIER ANALYSIS IS PERFORMED FOR ALL PLOT OUTPUTS.

FOR FOURIER ANALYSIS USING THE FFT, THE FORM IS -

FFT TSTR TSTP NUX

WHERE TSTR AND TSTP ARE THE START AND STOP TIMES FOR THE INTERVAL TO BE ANALYZED (ASSUMED TO BE THE PERIOD), AND 2**NUX IS THE NUMBER OF SAMPLE POINTS AVAILABLE.

NOTE: IF THE PERIOD SPECIFIED CONTAINS MORE THAN 2**NUX POINTS, ONLY THE FIRST 2**NUX ARE USED.

NOTE: WHEN USING EITHER FOR OR FFT, SPECIFYING 0.0 FOR BOTH TSTR AND TSTP WILL FORCE THE PROGRAM TO USE THE PERIOD DETERMINED BY THE STEADY-STATE ANALYSIS ALGORITHM (USED FOR OSCILLATORS).

\$ NO DC ANALYSIS SPECIFICATION -- THE FORM IS -

NDC

THIS SPECIFIES THAT NO DC ANALYSIS IS TO BE PERFORMED PRIOR TO STARTING TRANSIENT ANALYSIS. THE STATE OF THE CIRCUIT IS THE ZERO STATE UNLESS INITIAL CONDITIONS ARE SPECIFIED. THIS IS PRIMARILY USEFUL FOR OSCILLATORS WHICH ARE NEARLY STABLE AT THEIR DC POINT, SUCH AS A WIEN BRIDGE.

IN SOME CASES USING THE 'NDC' CARD WITH ZERO INITIAL CONDITIONS CAN LEAD TO EXCESSIVE COMPUTATION TIME, TRANSIENT CONVERGENCE FAILURE, OR OTHER PROBLEMS. IN SUCH CASES, THE BEST PROCEEDURE IS USUALLY TO RUN SINC TWICE. THE FIRST RUN IS ONLY TO ESTABLISH THE DC OPERATING POINT OF THE CIRCUIT (NO 'STEADY' OR 'TIME' CARD IS INCLUDED IN THE DECK). THE VALUES OF THE STATES ARE THEN DETERMINED FROM THE NODE VOLTAGES AND PERTURBED SLIGHTLY. THESE PERTURBED VALUES ARE THEN USED AS THE INITIAL CONDITIONS FOR THE SYSTEM, ALONG WITH AN 'NDC' CARD. THE PERTURBATION IS USUALLY SUFFICIENT TO ENSURE OSCILLATION; HOWEVER, THE SYSTEM IS NEAR ENOUGH TO THE STEADY STATE TO CONVERGE QUITE RAPIDLY. FOR AN EXAMPLE OF SUCH A CASE, SEE THE COLPITTS OSCILLATOR IN THE EXAMPLES SECTION.

\$ GAIN CARD -- THE GAIN CARD IS USED TO REQUIRE THE DETERMINATION OF THE DC TRANSFER FUNCTION OF THE CIRCUIT. THE GENERAL FORM IS -

GAIN N1 N2 N3 N4

WHERE N1 AND N2 ARE THE INPUT NODES
N3 AND N4 ARE THE OUTPUT NODES

THIS SPECIFICATION CAUSES THE DC SMALL-SIGNAL CHARACTERISTICS TO BE COMPUTED AND LISTED WITH THE DC OPERATING POINT IN THE FORM

SMALL-SIGNAL GAIN** N1 N2 N3 N4 K10
RIN=VALUE RM=VALUE AV=VALUE

WHERE K10 = 0 IF NO VOLTAGE SOURCE IS CONNECTED TO N1 AND N2
1 IF A VOLTAGE SOURCE IS CONNECTED TO N1 AND N2
RIN IS THE INPUT RESISTANCE
RM IS THE RESISTANCE MUTUAL TO THE INPUT AND OUTPUT PORTS, AND
AV IS THE VOLTAGE GAIN.

THE EQUIVALENT CIRCUIT REPRESENTATION USED FOR THE GAIN CALCULATION IS THE 'T' MODEL; THE VALUE GIVEN FOR RIN IS THE SUM OF THE SERIES INPUT AND MUTUAL (PARALLEL) RESISTANCES.

NOTE: SINCE THIS IS DC ANALYSIS, ALL INDUCTORS ARE REPLACED WITH 1 OHM RESISTORS, AND ALL CAPACITORS ARE TREATED AS OPEN CIRCUITS.

\$ STEADY-STATE ANALYSIS SPECIFICATION -- THE GENERAL FORM IS -
STEADY KSTE FREQ [PERIOD] [NST] [TOL] [NTRY2] [NTRY0]

WHERE KSTE = 0 FOR NON-AUTONOMOUS (DRIVEN) CIRCUITS
1 FOR AUTONOMOUS (OSCILLATOR) CIRCUITS
FREQ IS THE FREQUENCY OF THE FUNDAMENTAL COMPONENT PRESENT
PERIOD IS THE PERIOD OF THE FUNDAMENTAL COMPONENT PRESENT
NST IS THE MINIMUM NUMBER OF STEPS PER PERIOD
TOL IS THE STEADY-STATE CONVERGENCE TOLERANCE, EXPRESSED AS A FACTOR TIMES THE MAXIMUM STATE VALUE OVER THE PERIOD
NTRY2 IS THE NUMBER OF FULL PERIOD INTEGRATIONS OF THE SYSTEM ALLOWED IN ATTEMPTING TO ACHIEVE STEADY STATE CONVERGENCE.

NTRY0 IS THE NUMBER OF PERIODS OF CONTRACTION MAPPING COMPUTED BEFORE TAKING THE FIRST NEWTON STEP, AND EXCLUDES THE PERIOD REQUIRED FOR THE NEWTON STEP. THEREFORE, THE MINIMUM VALUE FOR NTRY0 IS 0. HOWEVER, IF THE SPECIFIED VALUE IS SMALLER THAN THE NUMBER OF DECAY PERIODS SPECIFIED (SEE 'DECAY PERIODS' CARD, BELOW), THE LARGER VALUE IS USED.

THE VARIABLES MUST BE LISTED IN ORDER; HOWEVER, NOT ALL ARE NEEDED.
DEFAULT VALUES ARE NST=50
TOL=1.0E-4
NTRY2=20
NTRY0=3

NOTE THAT IF BOTH FREQUENCY AND PERIOD ARE SPECIFIED, AND THE FREQUENCY IS NOT 0.0, THE PERIOD WILL BE FORCED TO BE 1./FREQ.

\$ TOLERANCE SPECIFICATION -- THE GENERAL FORM IS -

TOLERANCE TOL NTRA NTRY1

WHERE TOL IS THE TOLERANCE ON JUNCTION VOLTAGES, EXPRESSED AS A PERCENTAGE OF VT ($VT=KT/Q$; FOR ROOM TEMPERATURE THIS IS 26 MV. THUS, 10 % TOL WOULD BE 2.6 MV.)

NTRA = -1 IF THE BACKWARD-EULER METHOD IS TO BE USED
 0 IF GEAR'S 2ND ORDER METHOD IS TO BE USED
 1 IF THE TRAPEZOIDAL RULE IS TO BE USED FOR THE TRANSIENT ANALYSIS

NTRY1 IS THE NUMBER OF ATTEMPTS ALLOWED FOR CONVERGENCE TO THE SPECIFIED TOLERANCE.

IF THE CARD OR ANY VALUES ARE OMITTED, THE DEFAULTS ARE 10%, GEAR'S METHOD, AND 10 ATTEMPTS.

THE VALUES SPECIFIED ON THIS CARD APPLY TO EACH TIMESTEP. THIS IS IRRESPECTIVE OF WHETHER THE TIMESTEP IS PART OF A TRANSIENT ANALYSIS OR PART OF ONE PERIOD IN THE STEADY STATE ANALYSIS.

\$ SPECIAL CONDITIONS SPECIFICATIONS -- THE FOLLOWING SPECIAL CONDITIONS CONTROL CARDS ARE ALSO AVAILABLE -

PRINT STATES
 PRINT TRANSIENT ANALYSIS
 MODIFIED NEWTON METHOD N

THESE HAVE THE FOLLOWING EFFECTS-

PRINT STATES - CAUSES THE PRINTING OF THE INITIAL AND FINAL STATES AT EACH ITERATION DURING STEADY-STATE ANALYSIS. USED FOR OBSERVING CONVERGENCE. DEFAULT: THE INITIAL AND FINAL STATES FOR THE LAST ITERATION ONLY ARE PRINTED.

PRINT TRANSIENT ANALYSIS - CAUSES THE PRINTING OF THE VALUES REQUESTED BY A 'PRINT' CARD AT EACH TIMEPOINT FOR EACH ITERATION IN THE STEADY-STATE ANALYSIS. (WARNING: THE USE OF THIS OPTION MAY GENERATE CONSIDERABLE OUTPUT.) DEFAULT: THE OUTPUTS ARE PRINTED ONLY FOR THE FINAL ANALYSIS.

MODIFIED NEWTON METHOD N - WHERE N IS AN INTEGER FROM 1 TO 4; CAUSES A MODIFIED NEWTON METHOD TO BE USED FOR THE STEADY STATE ANALYSIS. N=0 CORRESPONDS TO AN UNMODIFIED METHOD. N=1 INVOKES COLON'S METHOD, USING A GLOBAL SCALAR DAMPING TERM. N=2 AND N=4 ARE VECTOR DAMPED METHODS; N=4 USES A VECTOR PRODUCED FROM LOCAL ERRORS ONLY, I.E., EACH STATE'S ERROR IS COMPUTED INDIVIDUALLY AND THE DAMPING TERM FOR THAT STATE IS A FUNCTION ONLY OF THAT ERROR. FOR N=2 BOTH LOCAL AND GLOBAL ERRORS ARE CALCULATED, AND THE DAMPING VECTOR IS A FUNCTION OF THEM BOTH. N=3 TAKES A FULL NEWTON STEP WHILE ALLOWING OTHER MODIFICATIONS, SUCH AS AN EXTERNAL CARD. ALL N>0 INCORPORATE REDUCED CALCULATION OF THE SENSITIVITY MATRIX. IF NO MODIFIED NEWTON METHOD CARD IS INCLUDED, THE DEFAULT IS THE UNMODIFIED NEWTON METHOD.

\$ EXTERNAL DAMPING SPECIFICATION -- THE GENERAL FORM IS -

EXTERNAL X1 X2 X3

WHERE X1,X2,... ARE USER-SPECIFIED DAMPING FACTORS WHEN USING THE MODIFIED (DAMPED) NEWTON METHOD. EACH ELEMENT I,J OF THE SENSITIVITY MATRIX IS MULTIPLIED BY $\text{SQRT}(X_I * X_J)$. THEREFORE, ALL X MUST BE NON-NEGATIVE. EXTREME CARE MUST BE USED IN SETTING ANY $X > 1$. SETTING ANY $X = 0$ ELIMINATES THAT ELEMENT FROM THE NEWTON METHOD.

NOTE THAT THE EFFECTIVE USE OF THIS OPTION REQUIRES THAT THE USER BE AWARE OF THE ORDER IN WHICH SINC SPECIFIES STATE VARIABLES, AS THIS IS THE ORDER IN WHICH THE DAMPING FACTORS MUST BE SPECIFIED ON THE EXTERNAL CARD. THIS ORDER IS -

ELEMENTS - INDUCTOR CURRENTS AND CAPACITOR VOLTAGES IN THE ORDER THEY APPEAR IN THE INPUT DATA

TRANSFORMERS - PRIMARY AND SECONDARY CURRENTS, IN THE ORDER THE TRANSFORMERS APPEAR IN THE INPUT DATA

BIPOLAR TRANSISTORS - CEB AND CBC VOLTAGES, IF THESE CAPACITANCES ARE NON-ZERO; TRANSISTORS IN THE ORDER THEY APPEAR IN THE INPUT DATA

DIODES - JUNCTION CAPACITANCE VOLTAGE, IF THE CAPACITANCE IS NON-ZERO

JFET TRANSISTORS - CGS AND CGD VOLTAGES, IF THESE CAPACITANCES ARE NON-ZERO.

ANY VALUES NOT SPECIFIED DEFAULT TO 1.0. IN ORDER THAT THE USER CAN VERIFY THESE SPECIFICATIONS, A SET OF THE STATES AND THEIR DAMPING FACTORS ARE PRINTED IF AN 'EXTERNAL' CARD IS USED.

\$ DECAY PERIOD SPECIFICATION -- THE GENERAL FORM IS

DECAY PERIOD N

WHERE N IS AN INTEGER FROM 1 TO 9.

IF THIS CARD IS PRESENT, DURING STEADY STATE ANALYSIS THE SYSTEM IS FORCED TO CONTRACT FOR N PERIODS PRIOR TO EACH APPLICATION OF THE NEWTON METHOD. THUS, IF BOTH VERY FAST AND VERY SLOW TRANSIENT COMPONENTS ARE PRESENT, USE OF AN APPROPRIATE VALUE FOR N WILL ALLOW THE FAST TRANSIENTS TO DECAY AND THEREBY REDUCE THEIR EFFECT ON THE NEWTON ALGORITHM. THIS TECHNIQUE IS PRIMARILY USEFUL WHERE A SYSTEM HAS TWO DISTINCT GROUPS OF EIGENVALUES, ONE OF WHICH IS ASSOCIATED WITH VERY SHORT TIME CONSTANTS.

NOTE: USE OF THIS TECHNIQUE MAY PRODUCE A LARGE INCREASE IN RUNTIME IF INAPPROPRIATELY APPLIED.

EXAMPLE

DECAY PERIOD 2

\$ ALTER CARD -- AN ALTER CARD IS USED TO SEPARATE MODIFICATIONS OF A CIRCUIT. IT INDICATES THE END OF DATA FOR A PARTICULAR ANALYSIS AND THAT ANOTHER ANALYSIS IS DESIRED WITH SOME CHANGES TO THE DATA. FOR EXAMPLE, WE COULD HAVE A RESISTOR NAMED R13 WITH A VALUE OF 10K OHMS. AFTER FIRST ANALYZING THE DATA WITH THIS VALUE OF R13, IT IS DESIRED TO DETERMINE THE EFFECT OF REDUCING THE VALUE TO 5K. THEN AN 'ALTER' CARD WOULD MARK THE END OF THE ORIGINAL DATA SET; FOLLOWING THE 'ALTER' CARD WOULD BE A NEW R13 CARD. THIS IS USEFUL IN DETERMINING THE EFFECT OF MINOR CHANGES IN A PROGRAM. NO PROVISION IS MADE FOR ELIMINATING RESISTORS COMPLETELY OR SHORTING THEM OUT. HOWEVER, MOST CHANGES THAT DID NOT CHANGE THE NODE STRUCTURE ARE ALLOWED, SUCH AS TOLERANCE CHANGES, CHANGES IN SEMICONDUCTOR MODELS, ETC. IF A COMPLETE RESTRUCTURING OF THE CIRCUIT IS DESIRED, THEN A NEW PROBLEM SHOULD BE SUBMITTED. NOTE THAT PROBLEMS CAN BE BATCHED BY PLACING A NEW TITLE CARD FOLLOWING THE PREVIOUS END CARD.

\$ END CARD -- THE END CARD MUST BE THE LAST CARD OF THE INPUT DECK. COLS. 1-3 MUST CONTAIN THE LETTERS END. COLS. MAY CONTAIN ANY COMMENTS.

REMARKS

THE FOLLOWING ARE GENERAL COMMENTS CONCERNING CHARACTERISTICS AND LIMITATIONS OF THE PROGRAM.

- 1) JOBS MAY BE BATCHED BY PLACING THEM SEQUENTIALLY -- A TITLE CARD FOR EACH NEW JOB FOLLOWING THE END CARD OF THE PRECEDING JOB.
- 2) SCALE FACTOR CAN BE USED FOR NON-INTEGERS. THE CONVENTION IS
G FOR 1E9
ME 1E6
K 1E3
M 1E-3
U 1E-6
N 1E-9
P 1E-12

ANY NON-SCALE FACTOR CHARACTER FOLLOWING A REAL NUMBER IS IGNORED (EXCEPT A SLASH). ANY CHARACTERS FOLLOWING A SCALE FACTOR ARE ALSO IGNORED.

- 3) DECIMAL POINT IS NOT NECESSARY FOR NON-INTEGERS. ALL INTEGERS MUST BE NON-NEGATIVE, AND SHOULD NOT HAVE DECIMAL POINT, E PART OR SCALE FACTOR.
- 4) DO NOT USE THE LETTER K TO INDICATE THE TEMPERATURE UNIT. THE LETTER K, UNLIKE O, F, H, A, AND V, CANNOT BE USED AS COMMENT, IT IS A SCALE FACTOR.

RUNNING SINC ON THE CSL DECSYSTEM-10

USUALLY DATA INPUT/OUTPUT WILL BE VIA DISK FILES. THE FORTRAN UNIT NUMBERS 2 AND 3 (USUALLY THE CARD READER AND LINE PRINTER) HAVE BEEN ASSIGNED TO THE DISK USING OPEN AND CLOSE STATEMENTS IN THE SINC MAIN PROGRAM. WHEN RUNNING SINC FROM A TIME-SHARING TERMINAL THE USER WILL BE ASKED FOR THE INPUT AND OUTPUT FILE NAMES. THESE USUALLY CONSIST OF A 1 TO 6 CHARACTER NAME, A PERIOD, AND A THREE LETTER EXTENSION; BY CONVENTION, THE EXTENSION 'DAT' IS USED TO INDICATE DATA FILES. IF THE PERIOD AND EXTENSION ARE NOT TYPED, THE EXTENSION 'DAT' IS ASSUMED; IF THE PERIOD IS TYPED BUT NO EXTENSION IS SPECIFIED, A NULL EXTENSION IS USED.

THE USER CAN CREATE A DISK FILE CONTAINING THE INPUT DATA BY USING ONE OF THE EDITORS, SUCH AS SOS OR TECO. FOR INFORMATION ON THE EDITORS, CONSULT THE APPROPRIATE MANUAL. ONCE AN OUTPUT FILE HAS BEEN CREATED BY SINC, IT CAN BE SPOOLED TO THE LINE PRINTER BY USING THE 'PRINT' COMMAND. FOR FURTHER INFORMATION ON THE DECSYSTEM-10, CONSULT THE APPROPRIATE DEC MANUALS.

EXAMPLES

THE FOLLOWING ARE EXAMPLES OF SOME TEST PROGRAMS FOR SINC:

```
SINC      RESISTOR VOLTAGE DEVIDER
VS  1  0  10.
R1  1  2  1K
R2  2  0  3K
VOUT 2  0
END
```

```
SINC      POWER-SUPPLY EXAMPLE
VSIN 1  0  10.  60.  0.  0.
D1  Z1  1  2
Z1  IS=1.0E-12
C1  1  2  1.0UF
L1  2  3  0.1
C2  3  0  1000.UF
RLOAD 3  0  100.
PRINT STATES
STEADY 0  60.  0  100
PLOT  0  20  VOUT  3  0  VIN  1  0
PLOT  0.  200.  ICHG  2  3
END
```

```
SINC      ASTABLE MULTIVIBRATOR/S P FAN      FBG      4-27-79
VCC  5  0  5.0
RC1  1  5  1500
RC2  4  5  1.5K
RB1  2  5  100K
RB2  3  5  1.0E5
C1  1  2  145.PF
C2  3  4  145PF
B1  NPN  FT=250MEG  TSAT=5NS  CJE=1PF  CJC=1P
Q1  B1  1  3  0
Q2  B1  4  2  0
PLOT  -5.  5.  VOUT  1  0  VB2  2  0
STEADY  1  50.KHZ  0  100  1.0E-2  12
END
```

SINC RCA 3040 WIDEBAND AMPLIFIER TEST FBG 4-27-79
VSIN 1 0 0.1 50.MEG
VCC 2 0 15.
VEE 3 0 -15.
RS1 30 1 1K
RS1 31 0 1K
R1 6 3 4.8K
R2 5 3 4.8K
R3 9 3 811.
R4 8 3 2.17K
R5 8 0 820.
R6 2 14 1.32K
R7 2 12 4.5K
R8 2 15 1.32K
R9 16 0 5.25K
R10 17 0 5.25K
Q1 BNL 2 31 6
Q2 BNL 2 30 5
Q3 BNL 10 5 7
Q4 BNL 11 6 7
Q5 BNL 14 12 10
Q6 BNL 15 12 11
Q7 BNL 12 12 13
Q8 BNL 13 13 0
Q9 BNL 7 8 9
Q10 BNL 2 15 16
Q11 BNL 2 14 17
PLOT 2 10 V17 17 0
BNL NPN BF=80. RB=100 RO=50K,1MA FT=325MEG,0.684MA,5.812 TSAT=6.47NS
+ CJE=3PF,0,1,0.5 CJC=2PF,0,1,0.5 CSUB=2PF
TIME 0.5NS 50.NS
END

SINC COLPITTS OSCILLATOR/S P FAN

```

BA NPN
Q1 BA 2 3 4
RB 3 0 65.
RC 1 2 20
RE 4 6 4.7K
RL 5 1 750
L1 5 1 50.NH 0. -.002
C1 1 4 5.PF 0 10.6
C2 4 5 45.PF 0. -10.6
VCC 5 0 10.
VEE 6 0 -10.
PLOT 5 15 VOUT 1 0
PLOT -0.8 -0.2 VE 4 0
PLOT 0. 15. IOUT 1 2
STEADY 1 300.MEGHZ 0 100 1.0E-2 8
PRINT STATES
NDC
END

```

SINC TTL INVERTER (7400 SERIES)

```

VIN 1 0 0. 0. 0. 1NS 3.5 2NS 3.5 42NS 0. 43NS 0.
VCC 13 0 5.
RS 1 2 50
Q1 BND 4 3 2
RB1 13 3 4K
Q2 BND 5 4 6
RC2 13 5 1.4K
RE2 6 0 1K
Q3 BND 7 5 8
RC3 13 7 100
QD1 BD1 8 8 9
Q4 BND 9 6 0
Q5 BND 11 10 9
RB2 13 10 4K
QD2 BD1 11 11 12
QD3 BD1 12 12 0
PLOT -2. 6. V9 9 0 V5 5 0
TIME 1NS 100NS
BND NPN BF=50. RB=70 RC=40 RO=50K,1MA FT=1.17GHZ,1MA,3.655 TSAT=9.8NS
+ CJE=0.9PF,0,1,0.5 CJC=1.5PF,0,0.85,0.5 CSUB=2PF
BD1 NPN BF=50 RC=40 FT=1.37GHZ,1MA,3.655 CJE=0.9PF,0,1,0.5
END

```

SINC TSR DC-DC POWER CONVERTER
* 200 VDC IN, 5V AT 20 A NOMINAL OUT
VS 1 0 200.V
R1 1 4 3.0K
C1 1 4 0.1UF
C3 1 2 .00136UF
R3 2 3 1.5K
D1 ZPW1 3 4
D1 ZPW1 3 2
D3 ZPW2 5 7
ZPW1 IS=1.0E-14 RS=0.05
ZPW2 IS=1.0E-12 RS=0.01
C2 6 0 16000.UF
RSER 6 7 3.75MILLIOHM
RLOAD 7 0 0.25
RPRI 1 10 10.
UFLY 3 10 5 0 2034.UH 3.585UH .997
RPTX 3 11 0.1
QDRIV BSW 11 8 0
BSW NPN BF=100.
RBSW 8 9 1.0
VPLSW 9 0 0.0 1.0 24.9US 0.1US 0.1US 24.9US 50.US
TOLERANCE 10. 0 100
STEADY 0 20.0KHZ 0. 150 1.0E-2 20
PLOT 5 10 VOUT 7 0 VCAP 6 0
PLOT 0 500 VC1 4 0 VC3 2 0
PRINT STATES
DECAY PERIODS 1
END

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