CADMIC—Computer-Aided Design of Microwave Integrated Circuits

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Abstract—**A computer program for the analysis and design of distributed lumped circuits, including microwave integrated circuits, is discussed. It is capable of frequency-domain analysis, optimization of transducer power gain, reflection coefficient, and/or noise figure. Also, the program can compute the return difference with respect to any admittance parameter so that the stability of the circuit can be determined by the Nyquist criterion. The program handles complex impedances, resistors, capacitors, inductors, transmission lines, independent current sources and grounded voltage sources, voltage-controlled current sources, and multiport elements, such as transistors and circulators, described by their scattering or admittance parameters. It contains a free-format input. The implementation is based on the indefinite admittance matrix, sparse matrices, adjoint networks, the Fletcher–Powell or Fletcher minimization algorithm, and Bode's feedback theory.**

INTRODUCTION

In the past five years several general-purpose microwave ac analysis programs have been developed. Most of these programs [1]–[8] use transfer-matrix analysis techniques which work well for cascade network connections. Some of these programs include modifications to handle other special topologies [3]–[8], such as series and parallel connections, but the number of complex multiplications required per two-port section increases from 4 in the cascade case, using *ABCD* parameters, to approximately 50, in the case when the two-ports are described by their scattering parameters and connected in parallel [8]. Also, the input data for the description of the network topology increase in complexity. Other programs use a general port formulation [4], [8] or an indefinite admittance matrix formulation [9], [10] to handle general network structures. In the admittance formulation, the description of the circuit topology is simple; however, the analysis can be very time consuming since the inversion of a large matrix is required at each frequency and the number of complex multiplications is approximately $N^3/3$ where $N + 1$ is the number of nodes in the circuit. Recent studies [21] indicate that the use of sparse-matrix algorithms in the analysis of 20–30 node networks can reduce this operations count to between $4N$ and $16N$. The transfer-matrix analysis of a cascade of N two-port networks requires approximately $4N$ operations. The admittance matrix of the cascaded network is banded with a bandwidth of 3. Hence, the operations count using sparse techniques will also be approximately $4N$ [22]. Thus sparse techniques can even be competitive in the case of special structures. However, for a given network topology, sparse techniques initially require additional central-processing-unit time to order the network nodes to minimize fill-in and to generate the elimination code, but this need be done only once for a given network structure and if many analyses are required, this time becomes insignificant in comparison with the total.

Finally, only a few of the above programs include optimization. A recent study [11] indicates that the Fletcher–Powell [12] or new Fletcher algorithm [13] are superior to other types of optimization techniques. These algorithms require the calculation of a gradient. The adjoint network is the most efficient approach for the computation of the gradient of a particular response with respect to two or more parameters in the circuit [14]–[16]. None of the above programs utilize both of these results. Finally, the stability of the circuit is crucial, yet it is frequently neglected. When stability is considered in the above programs, the stability criterion is based on the real part of the input impedance [17], [18]. This criterion can lead to incorrect conclusions as will be illustrated later.

Motivated by sparse-matrix techniques [19]–[21] and the adjoint network approach for computing the gradient of a given performance index, the computer program CADMIC was developed. CADMIC performs ac analysis and optimization of the transducer power gain, noise figure, and/or reflection coefficient, and checks the stability of the circuit by means of the Nyquist criterion and Bode's return difference. It handles any circuit topology with distributed, lumped, and active elements.

AC ANALYSIS

The indefinite admittance matrix is the basis for the formulation of the network equations in CADMIC. The allowable two-terminal elements are resistors, capacitors, inductors, lossless, shorted, or open stubs, or any two-terminal element whose impedance is specified at each frequency of the analysis. In addition,

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the program allows for the specification of voltage-controlled current sources, lossless transmission lines, and any three-terminal elements whose scattering or admittance parameters are specified at each frequency of the analysis. Thus small-signal parameters for transistors, circulators, and other devices can be read into the program. The scattering or admittance parameters of the three-terminal device must be measured with respect to a common terminal. The scattering parameters must be converted to admittance parameters and the admittance parameters are added to the indefinite admittance matrix. The nodes of the circuit are numbered consecutively from O to N where O denotes the ground node and N is the total number of nodes excluding the ground node.

The method employed by CADMIC to calculate the node voltages is essentially Gaussian elimination [22]. However, in order to minimize the number of operations, a sparse-matrix routine similar to that described by Jenkins and Fan [23] is utilized. Storage is minimized by storing only the nonzero entries in the admittance matrix in vector form. Only the diagonal elements are used as pivots in the forward elimination and the nodes are reordered in order to reduce fill-in. Basically, the Markowitz [19] criterion is used to reorder the nodes since it is easier to employ and seems to be almost as effective in the admittance formulation as more complicated reordering schemes which minimize the local fill-in [19], [21]. The nodes of the indefinite admittance matrix are reordered as follows. The ground node is placed in the last row. The next N_v rows from the bottom denote the grounded voltage source nodes, and the first $L = N - N_v$ rows are ordered in ascending order with the row with the least number of nonzero elements first, etc. The ordered indefinite admittance matrix is illustrated in (1):

$$Y = \begin{array}{r} N - N_v\{ \\ N_v\{ \\ 0\{ \end{array} \left[\begin{array}{c|c|c} Y_{LL} & Y_{Ls} & Y_{Lg} \\ \hline Y_{sL} & Y_{ss} & Y_{sg} \\ \hline Y_{gL} & Y_{gs} & Y_{gg} \end{array}\right]. \tag{1}$$

Only nonzero elements are stored and nonzero operations are carried out in the solution of the equation

$$Y_{LL}\boldsymbol{V}_L = \boldsymbol{I}_L - Y_{Ls}\boldsymbol{V}_s \tag{2}$$

where $\boldsymbol{I}_L$ denotes the independent current sources connected to the first L nodes (reordered), $\boldsymbol{V}_s$ denotes the grounded independent voltage sources, and $\boldsymbol{V}_L$ denotes the unknown node voltages at the first L nodes (reordered). The matrix Y_{LL} is decomposed into a lower triangular matrix L and an upper triangular matrix U such that

$$Y_{LL} = LU. \tag{3}$$

The equation

$$LU\boldsymbol{V}_L = \boldsymbol{I}_{\text{eq}} \tag{4}$$

where $\boldsymbol{I}_{\text{eq}} = \boldsymbol{I}_L - Y_{Ls}\boldsymbol{V}_s$ is solved by forward and backward substitution [22].

Optimization

In addition to analysis, the user can request an optimization of specified parameters to approximate a desired performance of the circuit. The error function is specified as

$$\epsilon = \frac{1}{p}\sum_{\Omega} W(j\omega)|\theta(j\omega)|^p, \qquad \omega \in \Omega \tag{5}$$

where ϵ is the error function which is to be minimized, $W(j\omega)$ is a nonnegative weighting function, $p > 1$ is an integer, θ is the difference between the desired response and the actual response at the frequency ω, and Ω is a set of discrete frequencies specified by the user. The desired response can be specified as the transducer gain of a transistor or negative resistance amplifier, the reflection coefficient, or the noise figure of an amplifier. In most microwave optimization programs the function ϵ is minimized by means of direct search algorithms [1], [2], [24]. However, a recent study [11] indicates that gradient-based algorithms are superior. Thus in CADMIC the function ϵ is minimized by the Fletcher–Powell [12] or the new Fletcher [13] algorithm which require the computation of the gradient of ϵ with respect to the set of adjustable parameters Q in the circuit. For a given adjustable parameter $q_i \in Q$ we note that

$$\frac{\partial \epsilon}{\partial q_i} = \sum_{\Omega} \operatorname{Re}\left\{W(j\omega)|\theta(j\omega)|^{p-2}\left(\theta^* \frac{\partial \theta}{\partial q_i}\right)\right\}. \tag{6}$$

Below we show that θ can be expressed as a function of a particular response voltage V which depends on the desired performance function. Thus we need to compute $\partial V/\partial q_i$ at each frequency $\omega \in \Omega$ and for each adjustable parameter q_i in the set Q. For example, in the case of the design for optimal transducer power gain

$$\theta(j\omega) \triangleq G_T(j\omega) - G_T{}^d(j\omega) \tag{7}$$

where

$$G_T(j\omega) = 4R_sG_L \frac{|V_0|^2}{|V_{in}|^2}$$

V_{in} is the generator voltage and R_s is the real part of the generator impedance at the frequency ω; V_0 is the voltage across the load and G_L is the real part of the admittance of the load, and $G_T{}^d$ is the desired transducer power gain specified at each frequency $\omega \in \Omega$. Thus (6) becomes

$$\nabla\epsilon = \sum_{\Omega} \operatorname{Re}\left\{W(j\omega)|G_T(j\omega) - G_T{}^d(j\omega)|^{p-2}(G_T(j\omega) - G_T{}^d(j\omega))8R_sG_L \frac{V_0^*}{|V_{in}|^2}\frac{\partial V_0}{\partial \boldsymbol{q}}\right\} \tag{8}$$

where * denotes the complex conjugate and

$$\frac{\partial V_0}{\partial \boldsymbol{q}} = \left[\frac{\partial V_0}{\partial q_1} \cdots \frac{\partial V_0}{\partial q_n}\right]^T$$

and n denotes the number of adjustable parameters.

Other types of error functions may be easily implemented if (5) is not satisfactory, e.g., the least pth approximation suggested by Bandler [25].

Finally, let us illustrate the ease in which the gradient of ϵ in (8) can be computed. In order to calculate $\partial V/\partial \boldsymbol{q}$, where V is a specified response for each $q_i \in Q$, $i = 1,2,\cdots n$, we use the adjoint network method [14]–[16]. The adjoint network has the same topology as the original network except its admittance matrix is Y^T where superscript T denotes the transpose operation. In the case where $\partial V/\partial \boldsymbol{q}$ is desired, where $\boldsymbol{q}$ denotes the vector of adjustable parameters, then the independent sources are set to zero in the adjoint network and a 1-A current source is connected to the response terminals. In order to find the solution of the adjoint circuit, we must solve the equation

$$Y_{LL}{}^T \boldsymbol{V}_L{}^a = \boldsymbol{I}_{\text{eq}}{}^a$$

or

$$(\boldsymbol{V}_L{}^a)^T Y_{LL} = (\boldsymbol{I}_{\text{eq}}{}^a)^T. \tag{9}$$

Since we already have the LU decomposition of Y_{LL} from (4), the solution of (9) simply requires a special subroutine for the forward and backward substitution. The sensitivity of V with respect to any element in the circuit can be computed from the knowledge of $\boldsymbol{V}_L$ and $\boldsymbol{V}_L{}^a$ as illustrated in Table I. Thus not even two complete network analyses are required to compute the gradient.

Now in (8) we must compute $\partial\epsilon/\partial q_i$ for each component. This is accomplished as illustrated in Fig. 1. Fig. 1(a) illustrates the circuit for which the transducer gain must be computed and optimized, and Fig. 1(b) represents its adjoint where from (8) we see that it is necessary to choose

$$I^a = W(j\omega)|G_T - G_T{}^d|^{p-2}(G_T - G_T{}^d)\,\frac{8R_s}{R_L}\,\frac{V_0{}^*}{|V_{in}|^2}. \tag{10}$$

TABLE I
FIRST-ORDER SENSITIVITY WITH RESPECT TO COMPONENT VARIATIONS EXPRESSED IN TERMS OF THE ORIGINAL NETWORK AND THE ADJOINT NETWORK VOLTAGES

Element	$\partial V/\partial q$	Δq Component
k C l Capacitor	$j\omega V_{k\ell} V^a_{k\ell}$	ΔC
k R l Resistor	$-V_{k\ell} V^a_{k\ell}/R^2$	ΔR
k L l Inductor	$jV_{k\ell} V^a_{k\ell}/\omega L^2$	ΔL
l_1 k Z_0 l Lossless Short-Stub	$jV_{k\ell} V^a_{k\ell}/Z_o^2 \operatorname{tg}\beta\ell_1$	ΔZ_o
	$j\beta V_{k\ell} V^a_{k\ell}/Z_o \sin^2\beta\ell_1$	$\Delta\ell_1$
l_1 Z_0 Lossless Open-Stub	$-jV_{k\ell} V^a_{k\ell} \operatorname{tg}\beta\ell_1/Z_o^2$	ΔZ_o
	$j\beta V_{k\ell} V^a_{k\ell}/Z_o \cos^2\beta\ell_1$	$\Delta\ell_1$
k l_1 l Z_0 Lossless Series Line	$j(V^a_k V_k + V_\ell V^a_\ell)/Z_o^2 \operatorname{tg}\beta\ell_1 - j(V_k V^a_\ell + V^a V_\ell)/Z_o^2 \sin\beta\ell_1$	ΔZ_o
	$j(V^a_k V_k + V^a_\ell V_\ell)\beta/Z_o \sin^2\beta\ell_1 - j(V^a_k V_\ell + V_k V^a_\ell)\beta/Z_o \sin\beta\ell \operatorname{tg}\beta\ell_1$	$\Delta\ell_1$

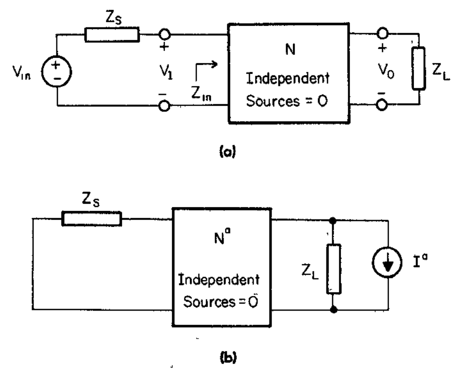


Fig. 1. (a) Computation of transducer gain. (b) Adjoint network for transducer gain.

Remember that the admittance matrix of the adjoint network is just the transpose of the admittance matrix for the original circuit in Fig. 1(a). Note that to compute the gradient we simply compute the node voltages of the original network and the adjoint network and sum the real parts of the sensitivities in Table I; that is, if the element is a lossless shorted stub, then the gradient of the error function with respect to the length of the stub is

$$\frac{\partial\epsilon}{\partial l_1} = \sum_{\Omega} \operatorname{Re}\,\{j\beta V_{kl} V_{kl}{}^a/z_0 \sin^2 \beta l_1\} \tag{11a}$$

where l_1 is the length of the lossless shorted stub connected between nodes k and l, and for a capacitor C connected between nodes k and l

$$\frac{\partial\epsilon}{\partial C} = \sum_{\Omega} \operatorname{Re}\,\{j\omega V_{kl} V_{kl}{}^a\} \tag{11b}$$

etc.

STABILITY

Typically, the stability of a microwave active network is determined by Rollet's conditions [17]. For a one-port network with driving-point impedance $Z(s)$, the conditions for short-circuit stability are: 1) the poles of $Z(s)$ are in the left-half plane (LHP); and 2) $\operatorname{Re} Z(j\omega) \geq 0$ for all ω.

The problem is that most designers neglect condition 1); an omission which can result in erroneous conclusions. For example, consider the circuit in Fig. 2 which consists of a generator with resistor R_g connected to a negative resistor with parasitic inductance and capacitance. The driving-point impedance for this circuit is given by

$$Z(s) = \frac{s^2RLC + s(RR_gC - L) + R - R_g}{sRC - 1}. \tag{12}$$

Clearly, the circuit is unstable [zeros of $Z(s)$ in the right-half plane (RHP)] if $R_g > R$, but under this condition

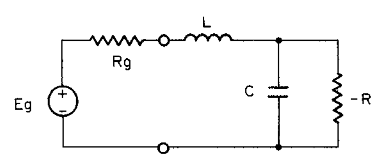


Fig. 2. Negative resistance circuit.

$$\text{Re}\,\{Z(j\omega)\} = R_g + \frac{-R}{1+(\omega RC)^2} > 0 \qquad (13)$$

for all ω. Obviously, condition 1) above is not a valid test when condition 2) is not satisfied.

Similarly, a two-port network terminated at one-port in a passive load is said to be stable provided [17]: a) the poles of the two-port parameters lie in the LHP, e.g., in the case of the short-circuit admittance parameters this implies that the two-port is short-circuit stable; and b) the real part of the immittance looking in at the unterminated port is positive for all ω with the other port terminated in a given passive impedance.

Again, the designer always assumes condition a) is satisfied and only checks b). In the case of the scattering parameters, condition a) is equivalent to the requirement that the network be stable when terminated in the port normalization resistors, and condition b) is equivalent to the requirement that the reflection coefficient be less than one at one-port with the other port terminated in the given passive load.

Due to the difficulty in checking condition a) it was decided to use a stability criterion based on Bode's feedback theory and the Nyquist criterion. The denominator of the transfer function can be expressed in the form

$$D(s) = D_1(s) + y(s)D_2(s) \qquad (14)$$

where $y(s)$ is some admittance parameter. Bode [18] defines the return difference as

$$F(s) = \frac{D(s)}{D_1(s)} = 1 + T(s) \qquad (15)$$

where

$$T(s) = y(s)\,\frac{D_2(s)}{D_1(s)} \qquad (16)$$

is called the return ratio. The number of times that the Nyquist diagram of $T(j\omega)$ encircles the point $(-1,0)$ in the clockwise direction is equal to $(Z_F - P_F)$ where Z_F is the number of zeros of $F(s)$ in the RHP (poles of the transfer function) and P_F denotes the roots of $D_1(s)$ *and the poles of* $y(s)$ *in the RHP*. Usually the designer can choose the parameter $y(s)$ such that there is reasonable assurance that $P_F = 0$. Thus if $P_F = 0$ (the circuit is stable with $y(s) = 0$ and $y(s)$ does not have any poles in the RHP), then the circuit is unstable if the Nyquist diagram encircles or passes through the point $(-1,0)$ in the clockwise direction. Desoer [26] has shown that the Nyquist diagram is applicable to distributed systems under some very general conditions.

Finally, the computation of $T(j\omega)$ is very straightforward [18] as illustrated in Figs. 3 and 4. In Fig. 3(a) the admittance $y(s)$ is a driving-point admittance. To compute $T(j\omega)$ with respect to this element, we simply set all independent sources to zero and replace $y(s)$ by the current generator $y(j\omega)$ in Fig. 3(b), then $T(j\omega) = V_{ij}(j\omega)$. In Fig. 4(a) the admittance parameter represents

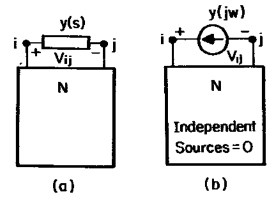


Fig. 3. Return ratio with respect to an admittance $y(j\omega)$.

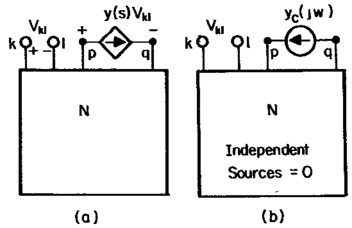


Fig. 4. Return ratio with respect to a voltage-controlled current source.

a voltage-controlled current source. Again, the independent sources are set equal to zero and the controlled sources are replaced by a current generator $y(j\omega)$ as shown in Fig. 4(b). Now, the return ratio $T(j\omega) = V_{kl}(j\omega)$, the voltage difference which controls the dependent current source. Thus it is a simple matter to generate the Nyquist diagram with respect to a given admittance parameter in the circuit.

Description of the Program

The basic flow chart of CADMIC is shown in Fig. 5. One of the most powerful features of CADMIC is the free format specifications. This is accomplished by subroutine READ whose operation is based on the use of three FUNCTION subprograms; the first one interprets a number read under an 80A1 format, the second obtains the character type,

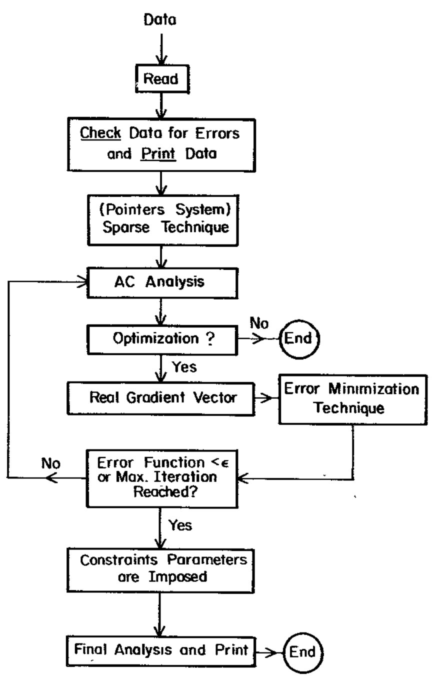


Fig. 5. Flow chart of CADMIC.

and the last checks delimiters, i.e., slash, comma, etc. Also, at subroutine READ, cards containing errors are detected and printed out with the type and location of the errors. Subroutine CHECK and PRINT, insures that the frequency points for analysis agree with the frequency data points of transistor, desired response, etc. It also does not accept a zero value for the nominal value of an element. The first part of the subroutine sparse checks that the number of nodes does not exceed 100; also, singular nodes (nodes on which fewer than two branches terminate) are identified and printed. The second part of subroutine sparse renumbers the nodes and sets up sparse-matrix indicators.

The analysis-gradient subroutine is described in some detail in Fig. 6. Observe that before entering this subroutine, the pointer system was set up without numerical values. The analysis-gradient routine is executed for each frequency with the numerical values of the parameters of the circuit.

The program is written in Fortran IV with a length of nearly 2400 Fortran statements. CADMIC is loaded on the Sigma 5 computer by using overlay. CADMIC requires less than 20 000 words of core storage, and uses single precision. It can handle 100 nodes, 250 parameter values of the following kind of components: resistors, capacitors, inductors, complex impedance; series lossless transmission lines, opened and shorted lossless stubs described by their lengths and characteristic impedances; 13 voltage-controlled current sources; 21 three-terminal devices with a common reference node, such as circulators and transistors; current sources, and grounded voltage sources. The above elements are keyed in the program as R, C, L, Z, U, O, S, GM, Q, I, and V, respectively.

CADMIC is easy to use since it contains a "free-format" input similar to that used by Jenkins and Fan [23]. The first card of the data deck is the title card and the last one is an end card. The cards between them can be arbitrarily ordered. Table II illustrates how easy circuit data can be supplied. Columns 1–4 must contain an element name of up to four characters. "A" stands for any alphanumeric character. Thus elements can be described with a meaningful name, i.e., R102, COUT, LIN. The numbers of the nodes between which the element is connected are $N1$ and $N2$; NG stands for ground node, and NCON is the controlling node for the VCCS's named GMAA's. Node numbers must be integers and all data fields must be separated by one or more blanks.

LB and UB are the lower and upper bounds specified when the element value is to be optimized; a zero value in the first LB field on the card is an instruction to the program not to optimize that parameter. The first LB and UB fields must not be left blank for distributed elements, or the program will interpret Z_0 as LB or UB. For transmission lines and stubs, default values for the bounds are supplied by the program if a slash is substituted for the LB or UB. The default lower bounds are 1° and 10 Ω and the upper bounds are 179° and 125 Ω for electric length and characteristic impedance, respectively. EL is the normalized electric length of the transmission lines and stubs; the actual length l is given by

$$l = \mathrm{EL}/[2\pi(\mu\epsilon)^{1/2}f_n] \tag{17}$$

where f_n is the normalized frequency. Thus $\mathrm{EL} = \pi/2$ is one-quarter wavelength at the frequency f_n. The permitivity of a medium may be defined as $\epsilon = \epsilon_r\epsilon_0$, where ϵ_0

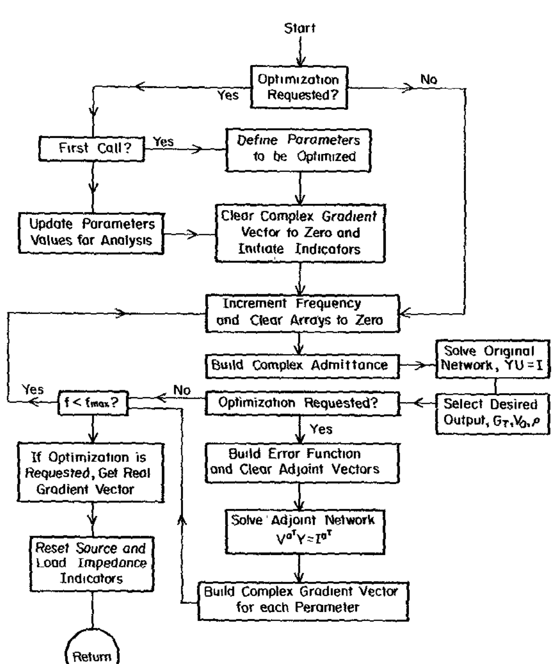


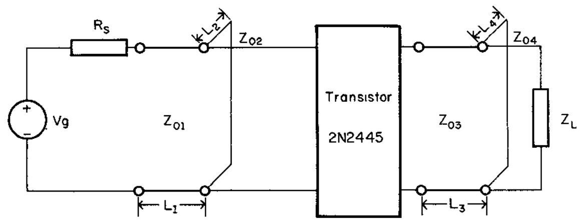
Fig. 6. Flow chart of the ANALYSIS-GRADIENT subroutine.

TABLE II
INPUT DATA INFORMATION

RAAA		N1	N2		VALUE	LB	UB			
CAAA		N1	N2		VALUE	LB	UB			
LAAA		N1	N2		VALUE	LB	UB			
ZAAA	F or V	N1	N2		Re Z	Im Z	Re Z	Im Z		
OAAA		N1	N2		EL	LB	UB	Z_o	LB	UB
SAAA		N1	N2		EL	LB	UB	Z_o	LB	UB
UAAA		N1	N2	NG	EL	LB	UB	Z_o	LB	UB
EPS		VALUE								
GMAA		N1	N2	NCON						
QAAA	DAAA	N1	N2	NR						
DAAA	Y or S	S11	/S11	S12	/S12	S21	/S21	S22	/S22	
VAAA		N+	NG	VALUE						
IAAA		N+	N-	VALUE						
VOUT		N+	N-							
GTR		NS1	NS2	NL1	NL2					
RFC		NS1	NS2							
OPT		1 or 2	IPRINT							
FRE		FNOR	F(1) F(2)							
+	F(N-2)	F(N-1)	F(N)							
WFU	P	F or V	W(1) W(2)		W(N)					
RDE		F or V	VALUE(1)	VALUE(2). . .		VALUE(N)				
STAB		I+	I-	V+	V-					
END										

TABLE III
ABBREVIATIONS ALLOWED FOR INPUTTING DATA VALUES

P	for	1E-12
N		1E-09
U		1E-06
M		1E-03
K		1E+03
ME		1E+06
G		1E+09



	L_1	Z_{01}	L_2	Z_{02}	L_3	Z_{03}	L_4	Z_{04}
Initial	06	50Ω	06	50Ω	06	50Ω	06	50Ω
Final	2 012	86 76Ω	0.976	97 57Ω	0833	125Ω	0.927	132Ω

Fig. 7. Broad-band amplifier with complex antenna load.

is the permitivity of free space and ϵ_r is the relative permitivity. In case ϵ_r is different from 1, a card EPS with the value of ϵ_r is introduced. Z_0 is the characteristic impedance of the line. On the complex impedance card ZAAA, F indicates that the impedance is constant with respect to frequency and only the real and imaginary values of Z are supplied; otherwise, values for each frequency must be specified, and a V instead of an F must be declared.

The card Q defines a three-terminal device such as a transistor or circulator and DAAA denotes its data field. NR is the common reference node. The card DAAA gives a data set in scattering or admittance parameters for each frequency; one set per card is allowed. Continuation cards specified by the symbol + in column 1 can be used as necessary. For current sources, $N+$ and $N-$ denote that the current flows from $N+$ to $N-$. The output specification cards are VOUT, RFC, and GTR for output voltage, reflection coefficient, and transducer power gain, respectively. NS1 and NS2 are the nodes associated with the source impedance and NL1 and NL2 with the load impedance. In card VOUT, $N+$ and $N-$ are the positive and negative nodes, respectively. The frequency card FRE specifies f_{n_y} defined in (17) and the frequencies for which the circuit is to be optimized or analyzed. The OPT card is used when optimization is required; a 1 for Fletcher–Powell algorithm and 2 for Fletcher is specified in the second field. If I PRINT = 1 the error function will be printed at each iteration, only the final error function will be printed if I PRINT = 0. Also, the weighting function card WFU and the desired response must accompany the OPT card. In card WFU, the second field shows p, defined in (5). The weighting function can be constant over the frequency band, indicated by an F in the next field in which case only one value need be specified. If variable, a V in the third field is used and a value for each frequency must be specified. The card RDE is very similar to WFU; the value or values are specified at each frequency for the magnitude (not in decibels) of the transducer power gain or the reflection coefficient. The card STAB is used when a stability check of the circuit is desired. $I+$ and $I-$ denote the nodes where the current source is connected. The program automatically sets the current source equal to the value of the admittance which was initially connected between those nodes. The return ratio is the voltage between nodes $V+$ and $V-$. For the case of a stability check with respect to the short-circuit forward-transfer admittance of a three-terminal device, set $I+ = N2$, $I- = N1$, $V+ = N1$, and $V- = $ NR. Engineering abbreviations to describe noninteger values in this program are shown in Table III. The END card must contain letters END in columns 1–3, columns 4–80 may contain any comment.

AN EXAMPLE

As an example, the broad-band amplifier with a complex antenna load shown in Fig. 7 is given. The problem is to obtain a flat response in the 150–300-MHz band. The listing of the input data and output results are given in Fig. 8(a) and (b). The design obtained has less ripple than the one obtained by Mokari-Bolhassan and Trick

```
                        **********************************
                        *                                *
                        *          C A D M I C           *
                        *                                *
                        **********************************

         BROAD - BAND  AMPLIFIER  WITH  COMPLEX  ANTENNA  LOAD
  GTR  5 4   1 0
  OPT 1 1
  RDE F 10.0
  VG 5 0 1.00
  RS  5 4  50.
  WFU  12  F  0.00006
  U1  4 3 0  .6 //  50 //
  S2  3 0  .6 //  50 / 132
  U3  2 1 0  0.6 //  50 //
  S4  1 0  .6 //  50 / 132
  Q  D 3 2 0
  D .39812 -78.4078  .0499625  67.0273  4.2535  137.8  .838804  -.1002E+02
+ .36942  -81.1241  .0518854  67.6859  4.00991  105.18  .8346  -9.9702
+  .332181  -83.43  .538516E-01  68.1986  .37919E+01  102.65  .8304 -9.91605
+ .30088 -85.61 .0557857  68.77  3.59918 100.41 .828263  -9.87173
+ .28017  -87.9546  .0578  68.934 3.4464  98.35 .82522 -9.87324
+ .263  -90.   .06  68.8186 3.29997   96.27  .822 -9.8404
+ .259 268.23 .06224 68.71 3.16 94.54 .819  -9.81
+ .2523 267.04 .064365 68.7771 3.0353 93.4  .81477 -9.751
+ .24573 265.56 .0664831  68.83  2.92207 92.16 .810745  -9.76439
+    .2391 264.48 .06852 69.0525 2.8202 90.82  .808605 -9.7186
+ .231698 263.05 .0705 69.409  2.74 90.  .805563 -9.71957
+  .222315 261.724 .07245 69.8142 2.67047 88.9272  .803423 -9.67335
+ .2132 260.07 .0734577 69.93 2.59193 87.7889  .799312 -9.6508
+   .2041 258.41 .0744 70.2 2.504  86.79 .7971  -9.57
+  .1950 256.96    .0753061  70.5304  2.4566  85.798  .793883  -9.49792
+ .18629 255.06  .781089E-01 71.3331   2.41006 84.7625   .791664  -9.41473
  FRE .23G .15G .16G .17G .18G .19G .2G .21G .22G .23G .24G .25G
+ .26G   .27G   .28G   .29G    .3G
  ZL  V  1 0    8.5 52.5  10. 56.5 13. 62.  16.  68. 20. 75 25  83  32.5 92.5
+ 40. 100.  55. 109.  70.  115. 92.5 120.   120.   120.  150.  110.
+  175.  85.     190.  55.     195. 15.
  END
```

(a)

```
         BROAD - BAND  AMPLIFIER  WITH  COMPLEX  ANTENNA  LOAD
 GTR      5  4


FREQUENCY                 GAIN                 ERROR FUNCT.
   HZ             DECIBEL      MAGNITUDE

1.5000E 08      9.5963E 00     9.1124E 00     1.1955135E-06
1.6000E 08      1.0283E 01     1.0674E 01     1.2398150E-06
1.7000E 08      1.0365E 01     1.0878E 01     2.2872564E-06
1.8000E 08      1.0192E 01     1.0453E 01     2.2876311E-06
1.9000E 08      1.0108E 01     1.0252E 01     2.2876311E-06
2.0000E 08      9.9967E 00     9.9925E 00     2.2876311E-06
2.1000E 08      1.0044E 01     1.0101E 01     2.2876311E-06
2.2000E 08      1.0040E 01     1.0093E 01     2.2876311E-06
2.3000E 08      1.0243E 01     1.0576E 01     2.2942522E-06
2.4000E 08      1.0333E 01     1.0798E 01     2.6253520E-06
2.5000E 08      1.0359E 01     1.0862E 01     3.4613040E-06
2.6000E 08      1.0286E 01     1.0680E 01     3.5098983E-06
2.7000E 08      1.0054E 01     1.0126E 01     3.5098983E-06
2.8000E 08      9.7725E 00     9.4897E 00     3.5114554E-06
2.9000E 08      9.5477E 00     9.0109E 00     7.8929643E-06
3.0000E 08      9.2892E 00     8.4902E 00     7.0945336E-04
 ERROR FUNCTION =   .7094534E-03
 PARAMETER VALUES       .97605E 00  .97573E 02  .92750E 00  .13200E 03  .20115E 01  .86759E 02  .83297E 00  .12500E 03
```

(b)

Fig. 8. (a) Listing of input data. (b) Listing of the output results.

[16]. The ripple was reduced over the frequency band by using $p = 12$ [refer to (5)]. The Fletcher–Powell algorithm was twice restarted. In Fig. 9(a) only the initial gain and the optimized one are shown. The stability of this circuit was checked by plotting the magnitude and phase of the return ratio with respect to $y_{21}(j\omega)$ of the transistor [see Fig. 9(b)]. Fig. 9(b) indicates that the circuit is stable by a wide margin in the frequency range over which the analysis was performed. One should also model the circuit and compute the return difference at frequencies outside this band.

Conclusion

CADMIC is a versatile program able to deal with ac analysis and optimization of general topologies. It incorporates a number of recent advances such as the use of sparse-matrix storage and computation techniques, the adjoint network approach for evaluating the gradient vector of suitable performance indices related to network responses. Also, a Nyquist stability criterion has been implemented in the program which offers more flexibility than the negative input resistance criterion used in some microwave programs.

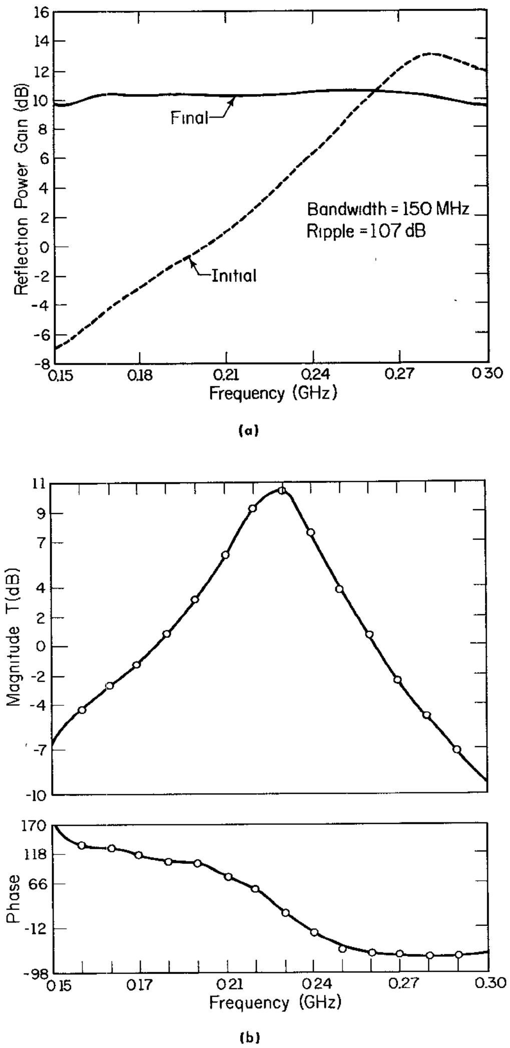


Fig. 9. (a) Transducer power gain versus frequency. (b) Return ratio for $y_{21}(j\omega)$ of transistor 2N2415.

The description of very complex topologies to CADMIC can be supplied in an effortless manner. It is not required to redraw the circuit or decompose it. Meaningful mnemonic abbreviations are used for the input data allowing the user to describe his circuit in a simple, concise, and significant way.

The example given in this paper was chosen mainly to illustrate the design techniques, although circuits with more involved topologies have been run successfully with CADMIC [27]. Research is continuing on the development and documentation of this program and its comparison with other analysis techniques with respect to speed and accuracy. When the topology is restricted and the network is small, CADMIC is not expected to be faster than some of the current smaller programs for microwave circuit analysis. However, there is a definite trend toward sparse-matrix techniques for the analysis of large networks [20], [28].

References

[1] T. N. Trick and J. Vlach, "Computer-aided design of broad-band amplifiers with complex loads," *IEEE Trans. Microwave Theory Tech.*, vol. MTT-18, pp. 541–547, Sept. 1970.
[2] V. G. Gelnovatch and I. L. Chase, "DEMON—an optimal seeking computer program for the design of microwave circuits," *IEEE J. Solid-State Circuits* (*Special ISSCC Issue on Solid-State Microwave Circuits*), vol. SC-5, pp. 303–309, Dec. 1970.
[3] P. Penfield, Jr., "Description of electrical networks using wiring operators," *Proc. IEEE* (*Special Issue on Computers in Design*), vol. 60, pp. 49–53, Jan. 1972.
[4] V. A. Monaco and P. Tiberio, "Automatic scattering matrix computation of microwave circuits," *Alta Freq.*, Feb. 1970.
[5] L. Besser, C. Hsieh, and F. Ghoul, "Computerized optimization of transistor amplifiers and oscillators using COMPACT," in *Proc. 1973 European Microwave Conf.*
[6] T. W. Houston and L. W. Read, "Computer-aided design of broad-band and low-noise microwave amplifiers," *IEEE Trans. Microwave Theory Tech.* (*Special Issue on Computer-Oriented Microwave Practices*), vol. MTT-17, pp. 612–614, Aug. 1969.
[7] "S-Parameters . . . circuit analysis and design," Hewlett-Packard Co., Hewlett-Packard Note 95, Sept. 1968.
[8] P. Bodharamik, L. Besser, and R. W. Newcomb, "Two scattering matrix programs for active circuit analysis," *IEEE Trans. Circuit Theory* (*Special Issue on Active and Digital Networks*), vol. CT-18, pp. 610–619, Nov. 1971.
[9] P. E. Green, "General purpose programs for the frequency domain analysis of microwave circuits," *IEEE Trans. Microwave Theory Tech.* (*Special Issue on Computer-Oriented Microwave Practices*), vol. MTT-17, pp. 506–514, Aug. 1969.
[10] M. A. Murray-Lasso and E. B. Kozemchak, "Microwave circuit design by digital computer," *IEEE Trans. Microwave Theory Tech.* (*Special Issue on Computer-Oriented Microwave Practices*), vol. MTT-17, pp. 514–526, Aug. 1969.
[11] D. M. Himmelblau, *Applied Nonlinear Programming.* New York: McGraw-Hill, 1972.
[12] R. Fletcher and M. J. D. Powell, "A rapidly convergent descent method for minimization," *Comput. J.*, vol. 6, pp. 163–168, June 1963.
[13] R. Fletcher, "A new approach to variable metric algorithms," *Comput. J.*, vol. 13, pp. 317–322, Aug. 1970.
[14] S. W. Director and R. A. Rohrer, "Automated network design—The frequency-domain case," *IEEE Trans. Circuit Theory*, vol. CT-16, pp. 330–337, Aug. 1969.
[15] J. W. Bandler and R. E. Seviora, "Current trends in network optimization," *IEEE Trans. Microwave Theory Tech.*, vol. MTT-18, pp. 1159–1170, Dec. 1970.
[16] M. E. Mokari-Bolhassan and T. N. Trick, "Computer-aided design of distributed-lumped-active networks," *IEEE Trans. Circuit Theory* (Corresp.), vol. CT-18, pp. 187–190, Jan. 1971.
[17] J. M. Rollet, "Stability and power-gain invariants of linear twoports," *IRE Trans. Circuit Theory*, vol. CT-9, pp. 29–32, Mar. 1962.
[18] E. S. Kuh and R. A. Rohrer, *Theory of Linear Active Networks.* San Francisco, Calif.: Holden-Day, 1967.
[19] R. P. Tewarson, *Sparse Matrices.* New York: Academic, 1973.
[20] G. D. Hachtel, R. K. Brayton, and F. G. Gustavson, "The sparse tableau approach to network analysis and design," *IEEE Trans. Circuit Theory* (*Special Issue on Computer-Aided Design*), vol. CT-18, pp. 101–114, Jan. 1971.
[21] R. D. Berry, "An optimal ordering of electronic circuit equations for a sparse matrix solution," *IEEE Trans. Circuit Theory* (*Special Issue on Computer-Aided Circuit Design*), vol. CT-18, pp. 40–50, Jan. 1971.
[22] G. Forsythe and C. B. Moler, *Computer Solution of Linear Algebraic Systems.* Englewood Cliffs, N. J.: Prentice-Hall, 1967.
[23] F. S. Jenkins and S.-P. Fan, "TIME-a nonlinear DC and time-domain circuit-simulation program," *IEEE J. Solid-State Circuits* (*Special Issue on Computer-Aided Circuit Analysis and Device Modeling*), vol. SC-6, pp. 182–188, Aug. 1971.
[24] A. D. Olmstead, "Comparison of some direct search optimization methods for the design of distributed-lumped-active networks," Coordinated Sci. Lab., Univ. Illinois, Urbana, Rep. 570, June 1972.
[25] J. W. Bandler and C. Charalambous, "Practical least *p*th optimization of networks," *IEEE Trans. Microwave Theory Tech.*, vol. MTT-20, pp. 834–840, Dec. 1972.
[26] C. A. Desoer, "A general formulation of the Nyquist criterion," *IEEE Trans. Circuit Theory*, vol. CT-12, pp. 230–234, June 1965.
[27] E. Sánchez-Sinencio, "Computer-aided design of microwave circuits," Coordinated Sci. Lab., Univ. Illinois, Urbana, Rep. 617, Aug. 1973.
[28] F. Bonfatti, V. A. Monaco, and P. Tiberio, "A microwave circuit analysis by sparse matrix techniques," presented at the IEEE G-MTT Int. Microwave Symp., Boulder, Colo., 1973.