

# A Unified Theory for Frequency-Domain Simulation and Sensitivity Analysis of Linear and Nonlinear Circuits

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**Abstract**—In this paper, a unified theory for frequency-domain simulation and sensitivity analysis of linear and nonlinear circuits is presented. An elegant derivation expands the harmonic balance technique from nonlinear simulation to nonlinear adjoint sensitivity analysis. This provides an efficient tool for the otherwise expensive but essential gradient calculations in design optimization. The hierarchical approach, widely used for circuit simulation, is generalized to sensitivity analysis and to computing responses in any subnetwork at any level of the hierarchy. Therefore, important aspects of frequency-domain circuit CAD such as simulation and sensitivity analysis, linear and nonlinear circuits, hierarchical and nonhierarchical approaches, voltage and current excitations, or open- and short-circuit terminations are unified in this general framework. Our theory provides a key for the coming generation of microwave CAD software. It will take advantage of the many existing and mature techniques such as the syntax-oriented hierarchical analysis, optimization, and yield driven design to handle nonlinear as well as linear circuits. Our novel sensitivity analysis approach has been verified by a MESFET mixer example exhibiting a 90 percent saving of CPU time over the prevailing perturbation method.

## I. INTRODUCTION

IN THIS PAPER, we present a unified approach to the simulation and sensitivity analysis of linear/nonlinear circuits in the frequency domain. The linear part of the circuit can be large and can be hierarchically decomposed, highly suited to modern microwave CAD. Analysis of the nonlinear part is performed in the time domain and the large-signal steady-state periodic analysis of the overall circuit is carried out by means of the harmonic balance (HB) method. In the sensitivity analysis we exploit the concept of the adjoint network technique.

The HB method has become an important tool for the analysis of nonlinear circuits. The work of Rizzoli *et al.* [1], Curtice and Ettenberg [2], Curtice [3], [4], Gilmore and Rosenbaum [5], Gilmore [6], Camacho-Penalosa and Aitchison [7] stimulated work on HB in the microwave CAD community. The excellent paper by Kundert and Sangiovanni-Vincentelli [8] provided systematic insight into the HB method. Many others, e.g., [9]–[15], have also contributed substantially to the state of the art of the HB

technique. The first step toward design optimization was made by Rizzoli *et al.* [1], who used the perturbation method to approximate the gradients. A recent review of this area was given by Rizzoli and Neri [16].

The adjoint network approach has been a classical vehicle for sensitivity analysis of linear circuits [17], [18] and of nonlinear time-domain or dc circuits [18], [19]. The existing methods, however, are not suitable for the sensitivity analysis of nonlinear circuits operating under large-signal steady-state periodic or almost periodic conditions, especially in the context of the HB method. The HB analysis is performed in the frequency domain and generates the circuit responses through their spectra. The time-domain approach, such as that of SPICE, is very inefficient in the steady-state case since the analysis must be carried out until the transient responses vanish. Therefore, the adjoint network nonlinear time-domain sensitivity analysis, even if applicable, would be inefficient as well. On the other hand, the nonlinear dc analysis is only a part of the harmonic balance analysis; moreover, it is integrated with the calculation of all harmonics simultaneously. If there are large-signal ac sources in the circuit, the dc analysis cannot be separated from the ac analysis. Therefore, the existing nonlinear dc adjoint network technique is not directly applicable to the HB method.

In our paper, we extend to the HB technique the powerful adjoint network concept. The concept involves solving a set of linear equations whose coefficient matrix is available in many existing HB programs. The solution of a single adjoint system is sufficient for the computation of sensitivities with respect to all parameters in both the linear and nonlinear subnetworks, in the bias circuit, driving sources, and terminations. No parameter perturbation or iterative simulations are required.

To make our theory highly suitable for microwave oriented CAD programs, we have also developed a hierarchical treatment of the adjoint system analysis. Preferred by leading experts, e.g. Jansen [20], and used in circuit simulators such as Super-Compact and Touchstone, the syntax-oriented hierarchical approach has proved very convenient and efficient in analyzing linear circuits. Our theory further extends such an approach to adjoint sensitivity analysis.

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TABLE I  
NOTATION AND DEFINITION

| Notation               | Definition   |
|------------------------|--|
| $N_t$                  | total number of nodes (internal and external) of a linear subnetwork.  |
| $N$                    | number of circuit nodes (or ports) used in harmonic analysis. Also, it is the number of external nodes for a typical subnetwork of Fig. 2.   |
| $H$                    | number of harmonics, including DC.   |
| $k$                    | harmonic index, $k = 0$ for DC, $k = 1$ for the fundamental harmonic, $k = 2, 3, \dots, H-1$ for other harmonics.  |
| $V_t(k)$ ,<br>$I_t(k)$ | complex $N_t$ -vectors indicating $k$ th harmonic voltages or currents at all nodes (both internal and external) of a linear subnetwork.   |
| $V(k)$ ,<br>$I(k)$     | complex $N$ -vectors indicating $k$ th harmonic voltages or currents at all external nodes of any linear subnetwork (at the highest level of hierarchy the nodes or ports at which the harmonic balance equations are formulated). |
| $\bar{V}_t, \bar{I}_t$ | real $2HN_t$ -vectors containing real and imaginary parts of $V_t(k)$ or $I_t(k)$ at all harmonics $k, k = 0, 1, \dots, H-1$ .   |
| $\bar{V}, \bar{I}$     | real $2HN$ -vectors containing real and imaginary parts of $V(k)$ or $I(k)$ at all harmonics $k, k = 0, 1, \dots, H-1$ .   |
| $Y_t(k)$               | $N_t$ by $N_t$ matrix representing the unreduced nodal admittance matrix of a linear subnetwork at harmonic $k$ .  |
| $Y(k)$                 | $N$ by $N$ matrix representing the reduced nodal admittance matrix of a linear subnetwork at harmonic $k$ .  |

| Notation    | Definition   |
|-------------|--|
| $\bar{Y}_t$ | $2HN_t$ by $2HN_t$ real matrix obtained by splitting the real and imaginary parts of $Y_t(k)$ for all harmonics $k, k = 0, 1, \dots, H-1$ .          |
| $\bar{Y}$   | $2HN$ by $2HN$ real matrix obtained by splitting the real and imaginary parts of $Y(k)$ for all harmonics $k, k = 0, 1, \dots, H-1$ .                |
| $\bar{J}_t$ | $2HN_t$ by $2HN_t$ real matrix representing the Jacobian defined in (A2).  |
| $\bar{J}$   | $2HN$ by $2HN$ real matrix representing the Jacobian defined by (5). The internal nodes of the linear subcircuit are suppressed.                     |
| $\bar{e}_t$ | $2HN_t$ real vector selecting the output voltage from the vector $\bar{V}_t$ .   |
| $\bar{e}$   | $2HN$ real vector selecting the output voltage from the vector $\bar{V}$ .   |
| $A(k)$      | $\begin{bmatrix} Y_t(k) & -U \\ U^T & 0 \end{bmatrix}$ where $U$ is $\begin{bmatrix} 0 \\ 1 \end{bmatrix}$ and $1$ is an $N$ by $N$ identity matrix. |
| $Z_d(k)$    | $n_g$ by $n_g$ diagonal matrix whose diagonal entries are the terminating impedances $Z_i, i = 1, 2, \dots, n_g$ .                                   |
| $Y_d(k)$    | $n_l$ by $n_l$ diagonal matrix whose diagonal entries are the terminating admittances $Y_i, i = 1, 2, \dots, n_l$ .                                  |
| $V_d(k)$    | $n_g$ -vector containing voltage excitations $E_i, i = 1, 2, \dots, n_g$ .   |
| $I_d(k)$    | $n_l$ -vector containing current excitations $I_i, i = 1, 2, \dots, n_l$ .   |

The sensitivities we propose are exact in terms of the harmonic balance method itself. Our exact adjoint sensitivity analysis can be used with various existing HB simulation techniques, e.g., the basic HB [8], the modified HB [6], and the APFT HB [15]. *The only computational effort includes solving the adjoint linear equations and calculating the Fourier transforms of all time-domain derivatives at the nonlinear element level.* Significant CPU time savings are achieved over the perturbation method.

In Section II, we define the notation used throughout this paper. In Section III, the simulation of linear and nonlinear circuits is reviewed under a general circuit hierarchy. In Section IV, a new and unified treatment to adjoint systems for linear and nonlinear circuits is introduced. Novel sensitivity formulas for nonlinear circuits are derived in Section V. Finally, in Section VI, a MESFET mixer example is used to verify our theory.

## II. NOTATION AND DEFINITION

We follow the notation of [8]. Real vectors containing voltages and currents at time  $t$  are denoted by  $v(t)$  and  $i(t)$ . Capitals  $V(k)$  and  $I(k)$  are used to indicate complex vectors of voltages and currents at harmonic  $k$ . A subscript  $t$  at  $V_t(k)$  indicates that the vector contains the nodal voltages at all  $N_t$  nodes (both internal and external) of a linear subnetwork. If there is no subscript, then the vector corresponds to the port voltages (currents) at all  $N$  ports of the reduced subnetwork. A bar denotes the split real and imaginary parts of a complex vector. In particular,  $\bar{V}$  or  $\bar{I}$  are real vectors containing the real and the imaginary parts of  $V(k)$  or  $I(k)$ , respectively, for all harmonics  $k, k = 0, 1, \dots, H-1$ . The total number of harmonics taken into consideration, including dc, is  $H$ . The hat distinguishes quantities of the adjoint system. For example,  $\hat{V}_t(k)$  represents adjoint voltages at internal and external nodes of a subnetwork at harmonic  $k$ . A detailed definition of the notation is given in Table I.

## III. LINEAR AND NONLINEAR SIMULATION

### A. Circuit Structure

Our exact adjoint sensitivity analysis can be used for hierarchically structured linear subcircuits. Consider the arbitrary circuit hierarchy of Fig. 1. A typical subnetwork containing internal and external nodes is shown in Fig. 2. A general representation of a terminated circuit is depicted in Fig. 3. An unpartitioned or nonhierarchical approach is a special case of Fig. 1 when only one level exists.

For a completely linear circuit, the sources and loads are applied at the highest level of the hierarchy, as depicted in Fig. 3. For a nonlinear circuit, the linear part of the overall circuit can have an arbitrary hierarchy as illustrated by Fig. 1 while the nonlinear part is connected directly at the highest level to the linear part. Therefore, in any case we consider an unterminated  $N$ -port circuit at the highest level of hierarchy. Such an approach simultaneously facilitates both the effect of the reference plane in microwave circuits and the need for the harmonic balance equations.

### B. Hierarchical Simulation of the Linear Network

Hierarchical simulation of linear circuits has been successfully used in many microwave CAD packages. It is summarized and expanded here into a set of formulas, enabling voltage responses at any nodes (internal or external) for any subnetwork at any level to be systematically computed. Firstly, we solve the terminated circuit at the

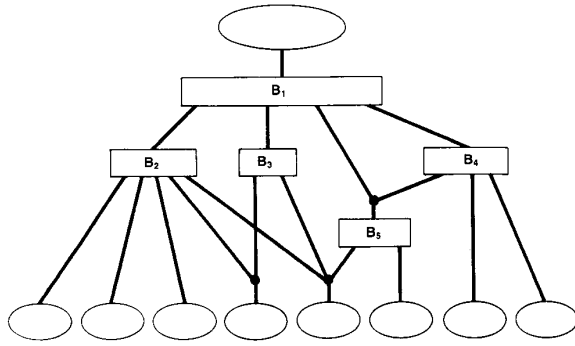


Fig. 1. An arbitrary circuit hierarchy. Each thick line represents a group of nodes. Each rectangular box represents a connection block for a subcircuit. Each bottom circular box represents a circuit element and the top circular box represents the sources and loads.

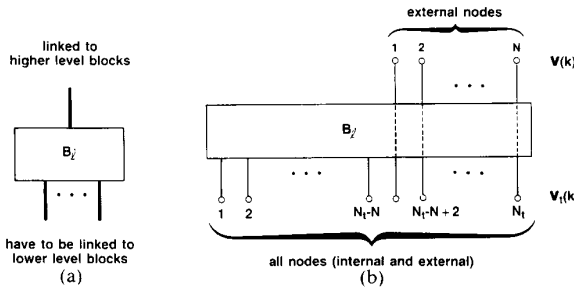


Fig. 2. A typical subcircuit connection block: (a) as seen from Fig. 1; (b) detailed representation of all the nodes of the subnetwork. Nodes at the top (bottom) of the rectangular box are the external (external and internal) nodes of the subnetwork.

highest level of the hierarchy using

$$\left\{ \begin{bmatrix} \mathbf{1} & \mathbf{0} \\ \mathbf{0} & \mathbf{Y}_s(k) \end{bmatrix} + \begin{bmatrix} \mathbf{Z}_s(k) & \mathbf{0} \\ \mathbf{0} & \mathbf{1} \end{bmatrix} \mathbf{Y}(k) \right\} \mathbf{V}(k) = \begin{bmatrix} \mathbf{V}_s(k) \\ \mathbf{I}_s(k) \end{bmatrix} \quad (1)$$

where the overall quantity in the curly bracket is an  $N$  by  $N$  matrix linking the port voltages  $\mathbf{V}(k)$  with the external sources for the terminated circuit. As defined in Table I,  $\mathbf{Y}_s(k)$  and  $\mathbf{Z}_s(k)$  are diagonal matrices containing terminating admittances and impedances, respectively, of the circuit shown in Fig. 3.  $\mathbf{Y}(k)$  is the admittance matrix of the unterminated circuit.  $\mathbf{V}_s(k)$  and  $\mathbf{I}_s(k)$  denote the voltage and current excitations of the circuit, respectively. The solution vector  $\mathbf{V}(k)$  contains external voltages of the circuit block under consideration. Then, all (both internal and external) nodal voltages  $V_i(k)$  of this subnetwork can be obtained from the equation

$$\mathbf{A}(k) \begin{bmatrix} \mathbf{V}_i(k) \\ \mathbf{I}(k) \end{bmatrix} = \begin{bmatrix} \mathbf{0} \\ \mathbf{V}(k) \end{bmatrix} \quad (2)$$

where  $\mathbf{A}(k)$  is the modified nodal admittance matrix of the subnetwork, as defined in Table I.  $\mathbf{I}(k)$  represents currents into the subcircuit through its external ports.

The solution of (2), i.e.,  $\mathbf{V}_i(k)$ , provides external voltages of all the subnetworks at the next level down the hierarchy. Therefore, (2) is used iteratively for the first, second,

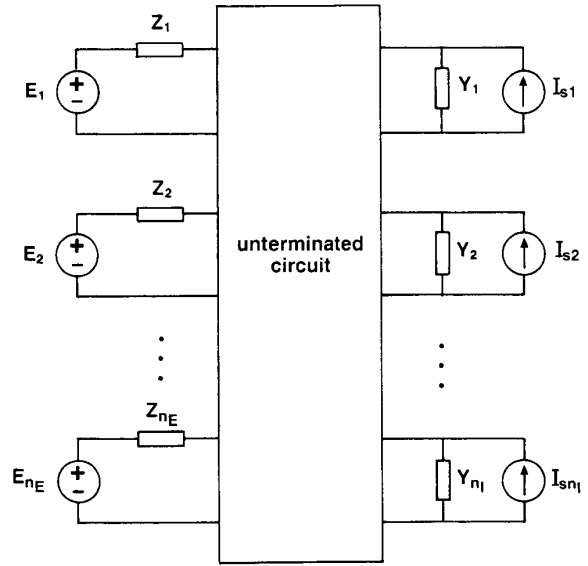


Fig. 3. A representation of a terminated subnetwork. Both current and voltage sources can be accommodated. The overall port sequence is such that ports  $1, 2, \dots, n_E$  correspond to voltage sources and ports  $n_E + 1, n_E + 2, \dots, n_E + n_I$  correspond to current sources. The total number of ports is  $N$ , i.e.,  $N = n_E + n_I$ .

... levels of the hierarchy until all desired nodal voltages are found.

Our formulas can directly accommodate both open- and short-circuit terminations. For example, a short-circuit termination at port 1 simply means  $Z_1 = 0$  in the matrix  $\mathbf{Z}_s$  in (1). An open-circuit termination at port  $n_E + 2$  simply means  $Y_2 = 0$  in the matrix  $\mathbf{Y}_s$  in (1).

C. Simulation of Nonlinear Circuits

The frequency-domain simulation of a nonlinear circuit is done effectively by the harmonic balance technique [1]–[16]. The problem is to find a  $\bar{\mathbf{V}}$  such that

$$\bar{\mathbf{F}}(\bar{\mathbf{V}}) \triangleq \bar{\mathbf{I}}_{NL}(\bar{\mathbf{V}}) + \bar{\mathbf{I}}_L(\bar{\mathbf{V}}) = \mathbf{0} \quad (3)$$

where the vectors  $\bar{\mathbf{I}}_L$  and  $\bar{\mathbf{I}}_{NL}$  are defined as the currents into the linear and nonlinear parts at the ports of their connection.  $\bar{\mathbf{V}}$  contains the split real and imaginary parts of voltages, as defined in Table I. The Newton update for solving (3) is

$$\bar{\mathbf{V}}_{\text{new}} = \bar{\mathbf{V}}_{\text{old}} - \bar{\mathbf{J}}^{-1} \bar{\mathbf{F}}(\bar{\mathbf{V}}_{\text{old}}) \quad (4)$$

where  $\bar{\mathbf{J}}$  is the Jacobian matrix defined by

$$\bar{\mathbf{J}} \triangleq (\partial \bar{\mathbf{F}}^T / \partial \bar{\mathbf{V}})^T. \quad (5)$$

The  $(i, j)$ th entry of the Jacobian matrix  $\bar{\mathbf{J}}$  is the derivative of the  $i$ th entry of  $\bar{\mathbf{F}}$  with respect to the  $j$ th entry of  $\bar{\mathbf{V}}$ .

In the context of the overall hierarchical structure, the solution of (3) provides the external voltages  $\mathbf{V}(k)$ ,  $k =$

#### IV. ADJOINT SYSTEM SIMULATION

Efficient and exact sensitivity analysis can be achieved by solving an adjoint system. In this section, a new and unified formulation of adjoint systems for hierarchically structured linear/nonlinear circuits is presented.

##### A. Adjoint System for Linear Networks

At the highest level of the hierarchy, the adjoint system is excited by a unit source at the output port. Suppose the output voltage  $V_{\text{out}}$  can be selected from  $V(k)$  by an  $N$ -vector  $e$  as

$$V_{\text{out}} = e^T V(k). \quad (6)$$

For example, if  $V_{\text{out}}$  is chosen as the voltage at the first port, then the vector  $e$  contains 1 as the first entry and zeros everywhere else. By solving

$$\left\{ \begin{bmatrix} \mathbf{1} & \mathbf{0} \\ \mathbf{0} & Y_s(k) \end{bmatrix} + \begin{bmatrix} Z_s(k) & \mathbf{0} \\ \mathbf{0} & \mathbf{1} \end{bmatrix} Y^T(k) \right\} \hat{V}(k) = e \quad (7)$$

we obtain adjoint voltages  $\hat{V}(k)$  at external ports at the highest level of the hierarchy.  $Y_s(k)$ ,  $Z_s(k)$ , and  $Y(k)$  are the same matrices as used in (1). In order to obtain adjoint voltages  $\hat{V}_i(k)$  at all (both internal and external) nodes of the circuit block, we solve the equation

$$A^T(k) \begin{bmatrix} \hat{V}_i(k) \\ -\hat{I}(k) \end{bmatrix} = \begin{bmatrix} \mathbf{0} \\ -\hat{V}(k) \end{bmatrix} \quad (8)$$

where  $A^T(k)$  is the transpose of the modified nodal admittance matrix of the subnetwork used in (2). The solution vector  $\hat{V}_i(k)$  provides external adjoint voltages for all subnetworks at the next level down the hierarchy. Therefore, (8) can be used iteratively for the first, second,  $\dots$  levels of the hierarchy until all desired adjoint voltages are found.

Notice that (8) is a convenient formulation of the adjoint system since the LU factors of  $A(k)$  can already be available from solving (2).

##### B. Adjoint System for Nonlinear Networks

Suppose  $\bar{V}_{\text{out}}$  is the real or imaginary part of output voltage  $V_{\text{out}}$  and can be selected from the voltage vector  $\bar{V}$  by a vector  $\bar{e}$  as

$$\bar{V}_{\text{out}} = \bar{e}^T \bar{V}. \quad (9)$$

The adjoint system is the linear equation

$$\bar{J}^T \hat{\bar{V}} = \bar{e} \quad (10)$$

where  $\bar{J}$  is the Jacobian at the solution of (3). Notice that  $\bar{V}$  and  $\hat{\bar{V}}$  are both 2HN-vectors containing the split real and imaginary parts of voltages at the connection ports of the linear and nonlinear subcircuits. According to our notation  $\bar{V}$  is defined for the original network and  $\hat{\bar{V}}$  is defined for the adjoint network. Also notice that the LU factors of  $\bar{J}$  can be available from the last iteration of (4). Therefore, to obtain  $\hat{\bar{V}}$  from (10), we need only the forward and backward substitutions.

The adjoint voltages can be computed even if the output port is suppressed from the harmonic equation (3). Al-

though the theoretical derivation for this case is rather involved, as given in the Appendix, we found a very logical and easy-to-implement method to handle this situation. First, we compute the adjoint voltages at the external ports of the linear subnetwork. This can be done by disconnecting the nonlinear part and then solving the linear part for individual harmonics separately. The resulting vector, denoted by  $\hat{\bar{V}}_L$ , is then transformed to the actual adjoint excitations of the overall circuit (including both linear and nonlinear parts) to be incorporated to (10) in place of  $\bar{e}$ . The final equation takes the form

$$\bar{J}^T \hat{\bar{V}} = \bar{Y}^T \hat{\bar{V}}_L. \quad (11)$$

In (11),  $\hat{\bar{V}}$  and  $\hat{\bar{V}}_L$  have exactly the same dimensions and both represent the split real and imaginary parts of adjoint voltages at the connection ports of the linear and nonlinear subcircuits. The former is computed from the overall circuit and the latter is computed from the linear subcircuit only.

Equations (10) or (11) provide adjoint voltages at external ports at the highest level of the hierarchy. We then use (8) iteratively for the first, second,  $\dots$  levels of the hierarchy to obtain adjoint voltages at both internal and external nodes of all subnetworks.

#### V. SENSITIVITY ANALYSIS

##### A. Adjoint System Approach to Sensitivity Evaluation

Let  $x$  be a design variable of the nonlinear circuit. Differentiating (3) with respect to  $x$  gives

$$(\partial \bar{F}^T / \partial \bar{V})^T (\partial \bar{V} / \partial x) + (\partial \bar{F} / \partial x) = \mathbf{0} \quad (12)$$

or

$$\partial \bar{V} / \partial x = -\bar{J}^{-1} (\partial \bar{F} / \partial x) \quad (13)$$

where  $\bar{J}$  has been defined in (5). Premultiplying (13) by  $\bar{e}^T$  results in

$$\begin{aligned} \partial \bar{V}_{\text{out}} / \partial x &= -\bar{e}^T \bar{J}^{-1} (\partial \bar{F} / \partial x) \\ &= -\hat{\bar{V}}^T (\partial \bar{F} / \partial x). \end{aligned} \quad (14)$$

This expression is further simplified by considering the locations of  $x$  in  $\bar{F}$ . Notice that each entry of vector  $\bar{F}$  corresponds to a port and to a harmonic of the circuit. Take, for instance, a nonlinear resistor described by  $i(t) = i(v(t), x)$  and connected across the  $j$ th port. The variable  $x$  enters  $\bar{F}$  at the positions relating to port  $j$  and harmonic  $k$ ,  $k = 0, 1, \dots, H-1$ , by the Fourier transform of  $i(v(t), x)$ . In this case, (14) is simplified to

$$\partial \bar{V}_{\text{out}} / \partial x = -\sum_k \text{Real} [\hat{V}_j(k) G^*(k)] \quad (15)$$

where  $\hat{V}_j(k)$  is the adjoint voltage at the  $j$ th port,  $G(k)$  is the  $k$ th Fourier coefficient of  $\partial i / \partial x$  and superscript  $*$  denotes the complex conjugate.

TABLE II  
SENSITIVITY EXPRESSIONS AT THE ELEMENT LEVEL

| Type of Element*                            | Expression for $G_b(k)$  | Applicable Equation |
|---|--|---------------------|
| linear G                                    | 1  | (16a)               |
| linear R                                    | $-1/R^2$   | (16a)               |
| linear C                                    | $j\omega_k$  | (16a)               |
| linear L                                    | $-1/(j\omega_k L^2)$   | (16a)               |
| nonlinear VCCS or resistor $i = i(v(t), x)$ | [kth Fourier coefficient of $\partial i/\partial x$ ]            | (16b)               |
| nonlinear capacitor $q = q(v(t), x)$        | $\omega_k$ [kth Fourier coefficient of $\partial q/\partial x$ ] | (16c)               |
| current driving source                      | 1  | (16b) or (16c)*     |
| voltage driving source                      | $\frac{1}{\text{source impedance}}$                              | (16b) or (16c)*     |

\* element is in branch  $b$  and contains  $x$

\* (16b) for the real part and (16c) for the imaginary part of the driving source

$\omega_k$  is the  $k$ th harmonic angular frequency

$0, 1, \dots, H-1$ , at the highest level of the linear part. The desired internal and external voltages at all levels of the hierarchy can be solved by using (2) iteratively.

### B. Sensitivity Expressions

Suppose a variable  $x$  belongs to branch  $b$ . We have derived the following general formula for computing the exact sensitivity of  $V_{\text{out}}$  with respect to  $x$ :

$$\frac{\partial \bar{V}_{\text{out}}}{\partial x} = \begin{cases} - \sum_k \text{Real} [\hat{V}_b(k) V_b^*(k) G_b^*(k)] & \text{if } x \in \text{linear subnetwork} & (16a) \\ - \sum_k \text{Real} [\hat{V}_b(k) G_b^*(k)] & \text{if } x \in \text{nonlinear VCCS or nonlinear resistor or real part of a complex driving source} & (16b) \\ - \sum_k \text{Imag} [\hat{V}_b(k) G_b^*(k)] & \text{if } x \in \text{nonlinear capacitor or imaginary part of a complex driving source.} & (16c) \end{cases}$$

Complex quantities  $V_b(k)$  and  $\hat{V}_b(k)$  are the voltages of branch  $b$  at harmonic  $k$  and are obtained from vectors  $V_i(k)$  and  $\hat{V}_i(k)$ , respectively.  $G_b(k)$  denotes the sensitivity expression of the element containing variable  $x$ . For example, if  $x$  is the conductance of a linear resistor,  $G_b(k) = 1$ . If  $x$  belongs to a nonlinear resistor represented by  $i = i(v(t), x)$ ,  $G_b(k)$  is the  $k$ th Fourier coefficient of  $\partial i/\partial x$ . A list of various cases of  $G_b(k)$  is given in Table II.

Our sensitivity formula (16) has no restrictions on the selection of harmonic frequencies or the time samples. In a multitone case, the index  $k$  in (16) corresponds to all the harmonics used in the harmonic equation (3). When the multidimensional Fourier transform is used, we simply place a multidimensional summation in (16).

TABLE III  
GRADIENTS OF MIXER CONVERSION GAIN

| Variable $x$   | Gradient Expression  |
|--|--|
| RF power   | $c \text{ Real}((\partial V_{\text{out}}/\partial x)/V_{\text{out}}) - 1$  |
| $R_g(f_{\text{RF}})$   | $c \text{ Real}((\partial V_{\text{out}}/\partial x)/V_{\text{out}}) + c/(2R_g(f_{\text{RF}}))$  |
| $R_d(f_{\text{IF}})$   | $c \text{ Real}((\partial V_{\text{out}}/\partial x)/V_{\text{out}} - 1/(R_d(f_{\text{IF}}) + jX_d(f_{\text{IF}}))) + c/(2R_d(f_{\text{IF}}))$ |
| $X_d(f_{\text{IF}})$   | $c \text{ Real}((\partial V_{\text{out}}/\partial x)/V_{\text{out}} - j/(R_d(f_{\text{IF}}) + jX_d(f_{\text{IF}})))$                           |
| any other parameter  | $c \text{ Real}((\partial V_{\text{out}}/\partial x)/V_{\text{out}})$  |
| $c = 20/\ln 10$  |  |
| R and X represent the real and the imaginary parts of the impedance terminations, respectively. Subscripts g and d represent the gate and the drain terminations, respectively.  |  |
| complex quantity $\partial V_{\text{out}}/\partial x$ is obtained by solving (9), (10) and (16) twice, once for the real part and the other for the imaginary part. The LU factors of J and the Fourier transforms of element sensitivities are common between the two operations. |  |

Notice that our sensitivity formulas permit variable  $x$  to appear in any subcircuit at any level of the hierarchy since all required voltages can be calculated as needed.

### C. Comparison with the Perturbation Method

To approximate the sensitivities using the traditional perturbation method, one needs a circuit simulation for each variable. The best possible situation for this method is that all simulations finish in one iteration. For our exact adjoint sensitivity analysis, the major computation, i.e., solving the adjoint equations, is done only once for all

variables. A detailed comparison reveals that the worst case for our approach takes less computation than the best situation of the perturbation method. In our experiment, we used only 1.6 percent of the CPU time required by the perturbation method to obtain all sensitivities.

### D. Gradient Vector for Optimization

The novel formula (16) can be used as a key to formulate the gradient vectors for design optimization and yield maximization of nonlinear circuits. Table III lists the gradients of a FET mixer conversion gain with respect to various variables, expressed as simple functions of  $\partial V_{\text{out}}/\partial x$ .

## VI. EXAMPLES

*Example 1: Hierarchical Circuit Description*

Many researchers, e.g., [3], and [7], have used FET mixer examples to test harmonic balance simulators. Here, we describe a mixer under the framework of hierarchical analysis. Such a description fits in with existing commercial software such as Super-Compact. The overall nonlinear circuit with its biasing and driving sources is described by a Super-Compact-like circuit file as follows.

```

* HIERARCHICAL ANALYSIS OF A MESFET MIXER
BLOCK
* INPUT MATCHING AND GATE BIAS SUBNETWORK
IND 3 4 L = 15NH
IND 2 3 L = .5NH
CAP 3 0 C = 2.2PF
CAP 1 2 C = 2.2PF
IND 2 5 L = .55NH
* DEFINE THE SUBCIRCUIT AS A 3-PORT
CKT1: 3PORT 1 4 5
END
BLOCK
* OUTPUT MATCHING AND DRAIN BIAS SUBNETWORK
IND 2 3 L = 15NH
IND 1 2 L = 1.1NH
CAP 2 0 C = 20PF
CAP 1 4 C = 20PF
* DEFINE THE SUBCIRCUIT AS A 3-PORT
CKT2: 3PORT 1 4 3
END
BLOCK
* THE HIGHEST HIERARCHY
CKT1 1 3 5
CKT2 7 2 4
CAP 6 C = 2PF
* A TRANSMISSION LINE BETWEEN PORT 6 0 AND PORT 7
MIC 6 7
* BIAS SOURCES
BIAS 3 V = -.9
BIAS 4 V = 3.
* NONLINEAR FET
* NODE NUMBERS REFER TO GATE, DRAIN AND SOURCE
NFET 5 6 0
END
FREQUENCIES
* DEFINE LO FREQUENCY
TONE 1
11GHZ
* DEFINE RF FREQUENCY
TONE 2
12GHZ
END
SOURCES
* DEFINE LO DRIVING SOURCE
TONE 1
POWER 1 0 P = 7DBM
* DEFINE RF DRIVING SOURCE
TONE 2
POWER 1 0 P = -15DBM
END

```

The LO and RF input matching and the gate bias circuits are analyzed separately in subnetwork CKT1. The IF output matching and drain bias circuits are analyzed in subnetwork CKT2. These subnetworks are then connected to a higher level of the hierarchy formulating an unterminated circuit block. This circuit block is then connected to nonlinear device ports. Using formulas developed in Sections III and IV, we are able to hierarchically simulate the original circuit as well as the adjoint circuit. This is a direct

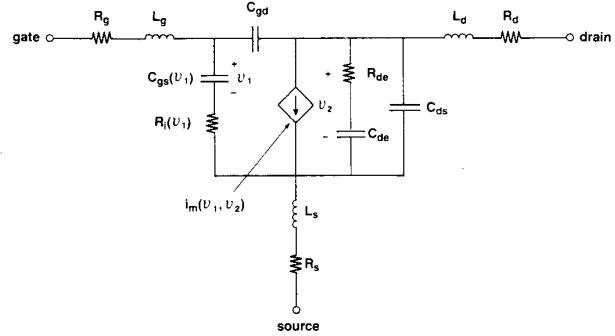


Fig. 4. A large-signal MESFET model. All parameter values are consistent with [7].

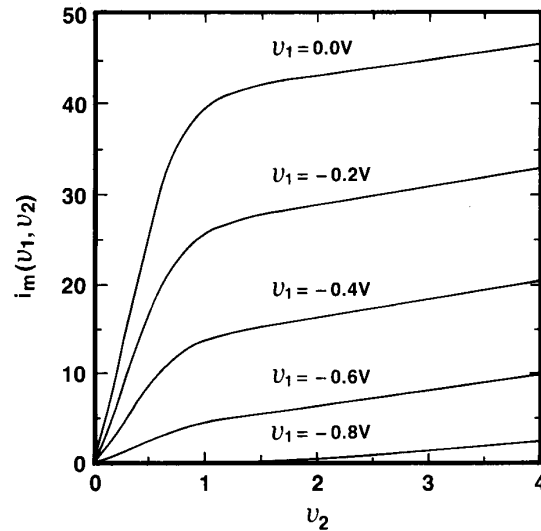


Fig. 5. The dc characteristics of the MESFET model.

realization of the syntax-oriented step-by-step topological description [20], permitting the sensitivity analysis of a large circuit to be performed by solving a set of small original and adjoint systems.

*Example 2: Simulation and Sensitivity Analysis of a MESFET Mixer*

The MESFET mixer example reported in [7] was used to verify our theory. Figs. 4 and 5 show the large-signal MESFET model and the dc characteristics of the device. The frequencies are  $f_{LO} = 11$  GHz,  $f_{RF} = 12$  GHz, and  $f_{IF} = 1$  GHz. The dc bias voltages are  $V_{GS} = -0.9$  V and  $V_{DS} = 3.0$  V. With LO power  $P_{LO} = 7$  dBm and RF power  $P_{RF} = -15$  dBm, the conversion gain was 6.4 dB. Twenty-six variables were considered, including all parameters in the linear as well as the nonlinear parts, dc bias, LO power, RF power, and IF, LO, and RF terminations. Exact sensitivities of the conversion gain with respect to all the variables are computed using our novel theory. The results were in excellent agreement with those from the perturbation method, as shown in Table IV. The circuit was solved in 22 seconds on a VAX 8600. The CPU time

TABLE IV  
NUMERICAL VERIFICATION OF SENSITIVITIES OF THE MIXER

| Location of Variable     | Variable              | Exact Sensitivity | Numerical Sensitivity | Difference (%) |      |
|--------------------------|-----------------------|-------------------|-----------------------|----------------|------|
| linear subnetwork        | $C_{ds}$              | 2.23080           | 2.23042               | 0.02           |      |
|                          | $C_{gd}$              | -29.44595         | -29.44659             | 0.00           |      |
|                          | $C_{de}$              | 0.00000           | 0.00000               | 0.03           |      |
|                          | $R_g$                 | 3.17234           | 3.17214               | 0.01           |      |
|                          | $R_d$                 | 6.42682           | 6.42751               | 0.01           |      |
|                          | $R_s$                 | 11.50766          | 11.50805              | 0.00           |      |
|                          | $R_{ds}$              | -0.02396          | -0.02412              | 0.66           |      |
|                          | $L_g$                 | -0.50245          | -0.50346              | 0.20           |      |
|                          | $L_d$                 | -0.20664          | -0.20679              | 0.07           |      |
|                          | $L_s$                 | 1.15334           | 1.15333               | 0.00           |      |
|                          | nonlinear subnetwork* | $C_{gs0}$         | -6.17770              | -6.17786       | 0.00 |
|                          |                       | $\tau$            | 0.49428               | 0.49414        | 0.03 |
| $V_g$                    |                       | -20.85730         | -20.85758             | 0.00           |      |
| $V_{p0}$                 |                       | -26.48210         | -26.48041             | 0.01           |      |
| $V_{ds}$                 |                       | 0.01064           | 0.01028               | 3.33           |      |
| $I_{dsp}$                |                       | 9.93696           | 9.93680               | 0.00           |      |
| bias and driving sources |                       | $V_{GS}$          | -31.62080             | -31.62423      | 0.01 |
|                          | $V_{DS}$              | -2.17821          | -2.17823              | 0.00           |      |
|                          | $P_{LO}$              | 2.76412           | 2.76412               | 0.00           |      |
|                          | $P_{RF}$              | -0.05401          | -0.05392              | 0.16           |      |
| terminations             | $R_g(f_{LO})$         | 0.06671           | 0.06657               | 0.22           |      |
|                          | $X_g(f_{LO})$         | 0.37855           | 0.37854               | 0.00           |      |
|                          | $R_d(f_{RF})$         | 0.78812           | 0.78798               | 0.02           |      |
|                          | $X_d(f_{RF})$         | 0.45120           | 0.45119               | 0.00           |      |
|                          | $R_s(f_{IF})$         | 0.71451           | 0.71436               | 0.02           |      |
|                          | $X_s(f_{IF})$         | 0.10886           | 0.10871               | 0.14           |      |

\* Nonlinear elements are characterized by

$$C_{gs}(v_1) = C_{gs0} / \sqrt{1 - v_1 / V_g},$$

$$R_s(v_1)C_{gs}(v_1) = \tau$$

and the function for  $i_m(v_1, v_2)$  is shown in Fig. 5, whose mathematical expression is consistent with [7].  $V_g$ ,  $V_{p0}$ ,  $V_{ds}$  and  $I_{dsp}$  are parameters in the function  $i_m(v_1, v_2)$ .

for sensitivity analysis using our method and the perturbation method are 3.7 seconds and 240 seconds, respectively. The CPU time saving of our method is 90 percent for both simulation and sensitivity calculations, and 98 percent if only sensitivity analysis is compared.

The dangling node between the nonlinear elements  $C_{gs}$  and  $R_s$ , a case which could cause trouble in HB programs, is directly accommodated in our approach.

We have plotted selected sensitivities versus LO power in Fig. 6. For example, as LO power is increased, conversion gain becomes less sensitive to changes in gate bias  $V_{GS}$ .

## VII. CONCLUSIONS

This paper presents a unified theory for frequency-domain simulation and sensitivity analysis of linear and nonlinear circuits. Our formula (16) encompasses the adjoint network approach previously used in linear [17], [18] and nonlinear dc circuits [18], [19] as special cases. Since the simulation of nonlinear circuits is expensive, gradient approximations for nonlinear circuits using repeated simu-

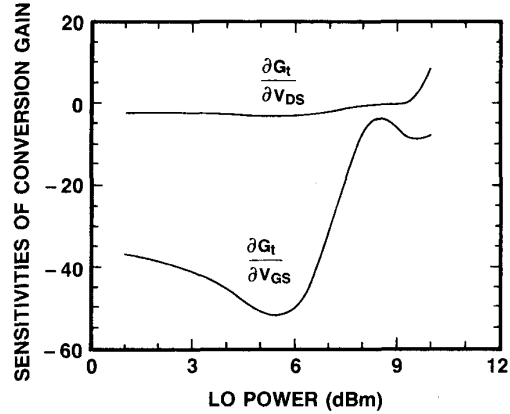


Fig. 6. Sensitivities of conversion gain with respect to bias voltages as functions of LO power.

lation are very costly. Consequently, the adjoint sensitivity analysis becomes far more significant for nonlinear circuits than for linear ones.

The hierarchical approach widely used for circuit simulation is generalized for sensitivity analysis and for computing responses in any subnetwork at any level of the hierarchy. Therefore, important aspects of frequency-domain circuit CAD such as simulation and sensitivity analysis, linear and nonlinear circuits, hierarchical and nonhierarchical approaches, voltage and current excitations, and open- and short-circuit terminations are unified in this general framework.

An immediate application of our theory would be the parameter extraction of nonlinear devices under RF large-signal excitations. The optimization criterion is to match computed responses with the measured ones at dc and at fundamental and higher harmonics. A powerful gradient optimizer should be used. Gradient information would be obtained using the adjoint network approach.

Our theory provides a key for the coming generation of microwave CAD software. It can take advantage of many existing and mature techniques such as the syntax-oriented hierarchical analysis and optimization and yield driven design to handle nonlinear as well as linear circuits.

Our novel sensitivity analysis approach has been verified by a MESFET mixer example. Compared with the perturbation method, the CPU time saving of our method is 90 percent for both simulation and sensitivity calculations, and 98 percent for sensitivity analysis only.

## APPENDIX A DERIVATION OF EQUATION 11

Suppose

$$\bar{V}_{out} = \bar{e}_t^T \bar{V}_t. \quad (A1)$$

The harmonic balance equations can be formulated with respect to all nodes of the circuit, i.e., without suppressing the internal nodes in a single level description of the circuit. In such a case the Jacobian matrix  $\bar{J}_t$  can be

defined similarly to (5), and

$$\bar{J}_i = \bar{Y}_i + QD^T P^T \quad (A2)$$

where  $D$  is a  $2HN \times 2HN$  matrix representing the contribution to  $\bar{J}$  from nonlinear components, i.e.,

$$\bar{J} = \bar{Y} + D^T. \quad (A3)$$

Matrices  $P$  and  $Q$  are  $2HN_i \times 2HN$  incidence matrices containing 0's and  $\pm 1$ 's.

Let

$$T = \bar{Y}_i^T. \quad (A4)$$

As with (9) and (10), based on (A1) the adjoint voltages at both internal and external nodes can be computed as

$$\hat{V}_i \triangleq (\bar{J}_i^T)^{-1} \bar{e}_i = (T + PDQ^T)^{-1} \bar{e}_i. \quad (A5)$$

Applying the Householder formula [21] to (A5) we have

$$\hat{V}_i = T^{-1} \bar{e}_i - T^{-1} P (D^{-1} + Q^T T^{-1} P)^{-1} Q^T T^{-1} \bar{e}_i. \quad (A6)$$

Notice that

$$(\bar{Y}^T)^{-1} = Q^T T^{-1} P. \quad (A7)$$

Let

$$X = \bar{Y}^T \quad (A8)$$

$$\hat{V}_L = Q^T T^{-1} \bar{e}_i. \quad (A9)$$

Premultiplying (A6) by  $Q^T$  gives

$$\hat{V} \triangleq Q^T \hat{V}_i = \hat{V}_L - X^{-1} (D^{-1} + X^{-1})^{-1} \hat{V}_L. \quad (A10)$$

Again, using the Householder formula [21],

$$(D^{-1} + X^{-1})^{-1} = X - X(D + X)^{-1} X \quad (A11)$$

and substituting (A3) and (A8) into (A10) we get

$$\hat{V} = (\bar{J}^T)^{-1} \bar{Y}^T \hat{V}_L \quad (A12)$$

or

$$\bar{J}^T \hat{V} = \bar{Y}^T \hat{V}_L. \quad (A13)$$

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#### REFERENCES

- [1] V. Rizzoli, A. Lipparini and E. Marazzi, "A general-purpose program for nonlinear microwave circuit design," *IEEE Trans. Microwave Theory Tech.*, vol. MTT-31, pp. 762-769, 1983.
- [2] W. R. Curtice and M. Eittenberg, "A nonlinear GaAs FET model for use in the design of output circuits for power amplifiers," *IEEE Trans. Microwave Theory Tech.*, vol. MTT-33, pp. 1383-1394, 1985.
- [3] W. R. Curtice, "Nonlinear analysis of GaAs MESFET amplifiers, mixers, and distributed amplifiers using the harmonic balance technique," *IEEE Trans. Microwave Theory Tech.*, vol. MTT-35, pp. 441-447, 1987.
- [4] W. R. Curtice, "GaAs MESFET modeling and nonlinear CAD," *IEEE Trans. Microwave Theory Tech.*, vol. MTT-36, pp. 220-230,

- 1988.
- [5] R. J. Gilmore and F. J. Rosenbaum, "Modelling of nonlinear distortion in GaAs MESFET's," in *IEEE Int. Microwave Symp. Dig.* (San Francisco, CA), 1984, pp. 430-431.
- [6] R. Gilmore, "Nonlinear circuit design using the modified harmonic balance algorithm," *IEEE Trans. Microwave Theory Tech.*, vol. MTT-34, pp. 1294-1307, 1986.
- [7] C. Camacho-Penalosa and C. S. Aitchison, "Analysis and design of MESFET gate mixers," *IEEE Trans. Microwave Theory Tech.*, vol. MTT-35, pp. 643-652, 1987.
- [8] K. S. Kundert and A. Sangiovanni-Vincentelli, "Simulation of nonlinear circuits in the frequency domain," *IEEE Trans. Computer-Aided Design*, vol. CAD-5, pp. 521-535, 1986.
- [9] S. Egami, "Nonlinear, linear analysis and computer-aided design of resistive mixers," *IEEE Trans. Microwave Theory Tech.*, vol. MTT-22, pp. 270-275, 1974.
- [10] F. Filicori, V. A. Monaco, and C. Naldi, "Simulation and design of microwave class-C amplifiers through harmonic analysis," *IEEE Trans. Microwave Theory Tech.*, vol. MTT-27, pp. 1043-1051, 1979.
- [11] R. G. Hicks and P. J. Khan, "Numerical analysis of subharmonic mixers using accurate and approximate models," *IEEE Trans. Microwave Theory Tech.*, vol. MTT-30, pp. 2113-2120, 1982.
- [12] C. Camacho-Penalosa, "Numerical steady-state analysis of nonlinear microwave circuits with periodic excitation," *IEEE Trans. Microwave Theory Tech.*, vol. MTT-31, pp. 724-730, 1983.
- [13] M. S. Nakhla and J. Vlach, "A piecewise harmonic balance technique for determination of periodic response of nonlinear systems," *IEEE Trans. Circuits Syst.*, vol. CAS-23, pp. 85-91, 1976.
- [14] A. Ushida and L. O. Chua, "Frequency-domain analysis of nonlinear circuits driven by multi-tone signals," *IEEE Trans. Circuits Syst.*, vol. CAS-31, pp. 766-779, 1984.
- [15] K. S. Kundert, G. B. Sorkin, and A. Sangiovanni-Vincentelli, "Applying harmonic balance to almost-periodic circuits," *IEEE Trans. Microwave Theory Tech.*, vol. MTT-36, pp. 366-378, 1988.
- [16] V. Rizzoli and A. Neri, "State of the art and present trends in nonlinear microwave CAD techniques," *IEEE Trans. Microwave Theory Tech.*, vol. MTT-36, pp. 343-365, 1988.
- [17] S. W. Director and R. A. Rohrer, "The generalized adjoint network and network sensitivities," *IEEE Trans. Circuit Theory*, vol. CT-16, pp. 318-323, 1969.
- [18] L. O. Chua and P. M. Lin, *Computer-Aided Analysis of Electronic Circuits*. Englewood Cliffs, NJ: Prentice-Hall, 1975.
- [19] D. A. Calahan, *Computer-Aided Network Design*, 2nd ed. New York, NY: McGraw-Hill, 1972.
- [20] R. H. Jensen, "Computer-aided design of hybrid and monolithic microwave integrated circuits—State of the art, problems and trends," in *Proc. European Microwave Conf.* (Nurnberg, West Germany), 1983, pp. 67-78.
- [21] A. S. Householder, "A survey of some closed methods for inverting matrices," *SIAM J.*, vol. 5, pp. 155-169, 1957.

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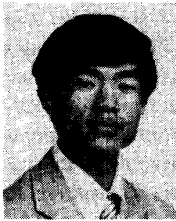


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