Nonlinear Analysis of Microwave FET Oscillators Using Volterra Series

YONGCAI HU, JUAN JESUS OBREGON, AND JEAN-CLAUDE MOLLIER

Abstract —**In this paper, a novel approach to determine the amplitude and frequency of nonlinear FET oscillators is presented. The nonlinear elements of the active device are modeled by Volterra series. The frequency and amplitude of oscillation are then calculated by solving two algebraic equations. Experimental results obtained from a constructed oscillator confirm the validity of the theory, the discrepancy between measured and calculated frequency and amplitude values being less than 10 percent.**

I. Introduction

In order to determine the amplitude and frequency of a nearly sinusoidal oscillator, several methods have been used, among them the describing function method [1] and the harmonic balance method [2]. Recently, the use of Volterra series has been proposed to analyze the oscillations in nonlinear systems [3]. In this paper, we demonstrate the applicability of the Volterra series method to the design of microwave oscillators. The amplitude and frequency of oscillation may be obtained quickly to within any desired accuracy using a recursive algorithm.

II. Theoretical Background

The describing function formalism is very attractive for one-port active device oscillators. Based on this formalism, a method has been proposed by Bates [1] to calculate the amplitude and frequency of oscillation for an IMPATT diode oscillator.

For two-port actives devices such as FET's, the nonlinearities depend on two variables: the gate–source voltage $v_{gs}(t)$ and the drain–source voltage $v_{ds}(t)$. In this case, the Volterra series can serve in the analysis of a nearly sinusoidal oscillator.

Let us consider a single-loop nonlinear feedback system and its associated open loop, as illustrated in Fig. 1(a) and (b), respectively. This system is assumed to have a convergent Volterra series representation [4]:

$$y(t)=\sum_{n=1}^{\infty}\int_{-\infty}^{+\infty}\cdots\int_{-\infty}^{+\infty}\mathscr{H}_n(\omega_1,\omega_2,\cdots,\omega_n)\cdot\prod_{i=1}^{n}u(\omega_i)\exp(j\omega_i t_i)\,d\omega_i \quad (1)$$

Manuscript received March 30, 1987; revised June 13, 1989.
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IEEE Log Number 8930650.

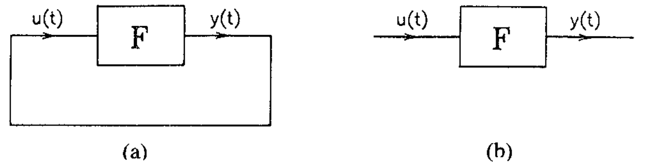

Fig. 1. (a) A closed-loop nonlinear feedback system. (b) Associated open-loop nonlinear system.

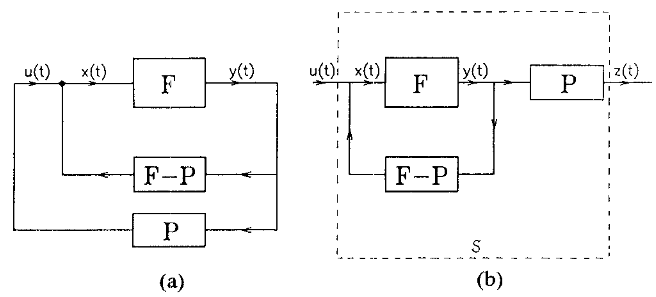

Fig. 2. (a) Equivalent representation of single-loop feedback system in Fig. 1(a). (b) Associated open-loop system.

where $u(t)$ and $y(t)$ are the input and the output signals, respectively, and ω_i is a radian frequency. We have to calculate the determining equation of this system, from which the amplitude and frequency of oscillation can be determined.

If the system in Fig. 1 has a periodic solution of radian frequency ω, then in general all harmonics $k\omega$ of this fundamental pulsation can be found in the Fourier spectrum of $u(t)$ and $y(t)$. The fundamental frequency component can be extracted from $y(t)$ using an ideal low-pass filter P and the remaining components by an ideal high-pass filter $F-P$. Then, transformation of the single-loop feedback system into the system shown in Fig. 2(a) immediately follows.

From the definition of P, we can write

$$u(t)=P[y(t)]=\frac{A}{2}e^{j\omega t}+\frac{A^*}{2}e^{-j\omega t} \quad (2)$$

where A is the complex amplitude of the signal at the pulsation ω.

Now cut the loop in Fig. 2(a) and redraw the resulting system in Fig. 2(b). The output z of the associated open-loop system S is written as follows:

$$z(t) = \frac{A_z}{2} e^{j\omega t} + \frac{A_z^*}{2} e^{-j\omega t} \quad (3)$$

because of the "ideal filter P."

The necessary and sufficient condition for the system in Fig. 1(a) to have a periodic solution of pulsation is that $A_z = A$. According to this condition, the determining equations can be calculated.

With the help of Volterra series theory, the Nth-order determining equation is obtained as (see the Appendix)

$$d_N(A,\omega) = H_1(j\omega) + \Omega_1(j\omega)|A|^2 + \Omega_2(j\omega)|A|^4 + \cdots + \Omega_N(j\omega)|A|^{2N} - 1 = 0 \quad (4)$$

so that its solution gives the output amplitude A and pulsation of the nearly sinusoidal oscillation to any desired accuracy with increasing N. In our applications, only weak nonlinearity is considered and an acceptable accuracy can be obtained (about 10 percent) in solving the first-order determining equation. So for $N=1$, eq. (4) becomes

$$d_1(j\omega) = H_1(j\omega) + \Omega_1(j\omega)A^2 - 1 = 0 \quad (5)$$

where

$$\Omega_1(j\omega) = \frac{1}{4}\{\mathscr{H}_3(j\omega, j\omega, -j\omega) + \mathscr{H}_3(j\omega, -j\omega, j\omega) + \mathscr{H}_3(-j\omega, j\omega, j\omega)\} \quad (6)$$

with

$$\mathscr{H}_3(j\omega, j\omega, -j\omega) = H_2(j2\omega, -j\omega)\frac{H_2(j\omega, j\omega)}{1 - H_1(j2\omega)} + H_2(j\omega, 0)\frac{H_2(j\omega, -j\omega)}{1 - H_1(0)} + H_3(j\omega, j\omega, -j\omega) \quad (7a)$$

$$\mathscr{H}_3(-j\omega, j\omega, j\omega) = H_2(0, j\omega)\frac{H_2(-j\omega, j\omega)}{1 - H_1(0)} + H_2(-j\omega, j2\omega)\frac{H_2(j\omega, j\omega)}{1 - H_1(j2\omega)} + H_3(-j\omega, j\omega, j\omega) \quad (7b)$$

and

$$\mathscr{H}_3(j\omega, -j\omega, j\omega) = H_2(0, j\omega)\frac{H_2(j\omega, -j\omega)}{1 - H_1(0)} + H_2(j\omega, 0)\frac{H_2(-j\omega, j\omega)}{1 - H_1(0)} + H_3(j\omega, -j\omega, j\omega). \quad (7c)$$

The second- and third-order transfer functions H_2 and H_3 are defined in the Appendix.

III. Application to Microwave FET Oscillators

Let us consider a microwave FET oscillator with a parallel feedback configuration as schematized in Fig. 3. C_F and L_F are feedback components which may be considered as electrically equivalent to a length of transmission line.

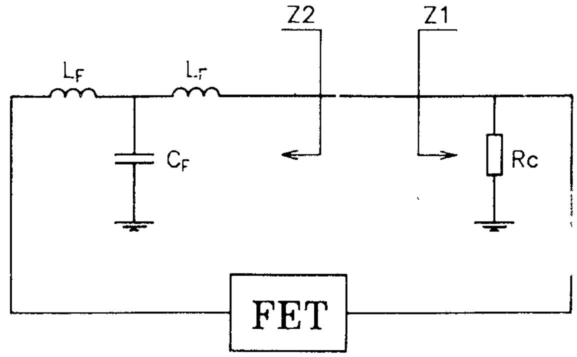

Fig. 3. A microwave FET oscillator with parallel feedback and resistive load R_c.

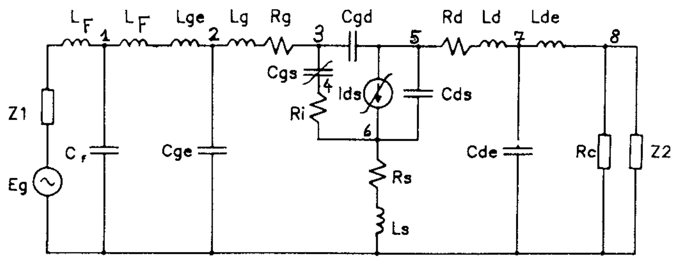

Fig. 4. Equivalent circuit of the open-loop FET oscillator for calculating the transfer functions:

$L_{ge} = 0.136$ nH	$C_{ge} = 0.25$ pF	$R_g = 2.75\ \Omega$
$L_g = 0.33$ nH	$C_{gd} = 0.046$ pF	$R_i = 3.8\ \Omega$
$L_{de} = 0.09$ nH	$C_{ds} = 0.20$ pF	$R_s = 1.2\ \Omega$
$L_d = 0.42$ nH	$C_{de} = 0.29$ pF	$R_d = 1.9\ \Omega$
$L_s = 0.04$ nH		

The transfer functions H_1, H_2, and H_3 of its associated open-loop circuit (Fig. 4) can be calculated by a recursive algorithm (see the Appendix) where three nonlinearities have been considered: the transconductance (g_m) and the drain conductance (g_d) (both represented by a current source I), and the Schottky-barrier junction capacitance C_{gs}. Z_1 and Z_2 represent the load impedances when the loop is opened.

The drain-to-source and gate-to-source currents I_{ds} and I_{gs} can be approximated by a power series as

$$I_{ds}(t) = \sum_{k=1}^{3}\left[g_{mk}v_{gs}^k(t) + g_{dk}v_{ds}^k(t)\right] \quad (8)$$

and

$$I_{gs}(t) = \frac{d}{dt}\sum_{k=1}^{3} C_{gs_k}v_{gs}^k(t) \quad (9)$$

where v_{ds} is the drain–source voltage and v_{gs} is the voltage across the capacitance C_{gs}. The g_{mk}, g_{dk}, and C_{gs_k} coefficients are then derived with the following steps:

i) Measure the dc current–voltage characteristics of the FET.
ii) Fit the experimental curves with Tajima's equation [5], $I_{DS} = fv_{gs}$, v_{ds}, and with the abrupt junction capacitance equation, $C_{gs} = f(v_{gs})$.

TABLE I
VALUES OF TRANSCONDUCTANCE g_m, DRAIN CONDUCTANCE g_d, AND SCHOTTKY-BARRIER JUNCTION CAPACITANCE C_{gs} FOR THE BIAS $V_{gs0} = -1.0$ V AND $V_{ds0} = 4.0$ V

g_{m1}	1.053E-1	g_{d1}	8.999E-3	C_{gs1}	1.180E-12
g_{m2}	3.082E-2	g_{d2}	- 2.914E-4	C_{gs2}	3.278E-13
g_{m3}	- 2.525E-5	g_{d3}	2.755E-5	C_{gs3}	2.731E-13

Units are mS and pF.

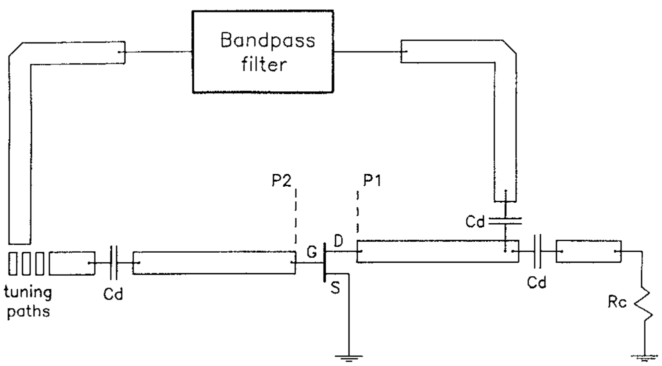

Fig. 5. Realization of the microwave FET oscillator. Bias circuits are omitted for clarity.

iii) Derive numerical values of the g_{mk}, g_{dk}, and C_{gs_k} coefficients.

Table I gives the values for a medium-power FET (Mitsubishi MGF-1802). With these values, the transfer functions H_1, H_2, and H_3 can be calculated, and finally the amplitude and frequency of oscillation can be obtained using the first-order determining equation (5).

IV. EXPERIMENTAL RESULTS

In order to verify the theoretical predictions, a microwave FET oscillator has been realized (Fig. 5). The usual FET bias circuit with two independent voltages V_{gs} and V_{ds} (not drawn in the figure) allows connection of the source electrode directly to ground, giving enhanced output power. A length of microstrip line has been used as a parallel feedback component. The transfer function of the passive circuit (between planes $P1$ and $P2$) has been measured for three different electrical lengths, in order to check the lumped-element values C_F, L_F of the equivalent T configuration.

Table II shows the theoretical and experimental values of the oscillation frequency and output power for those three lengths of microstrip. The discrepancy between numerical data and measured values is always smaller than 10 percent.

TABLE II
THEORETICAL AND EXPERIMENTAL RESULTS OF OSCILLATION FREQUENCY AND OUTPUT POWER FOR THE BIAS $V_{gs0} = -1.0$ V AND $V_{ds0} = 4.0$ V

Frequency calculated	Frequency measured	Output power calculated	Output power measured
2.90 (GHz)	2.75 (GHz)	18.6 (dBm)	18.2 (dBm)
2.80 (GHz)	2 68 (GHz)	18.7 (dBm)	18.3 (dBm)
2.70 (dBm)	2.61 (GHz)	18.9 (dBm)	18.7 (dBm)

V. CONCLUSION

A novel approach to determining the amplitude and frequency of nonlinear microwave FET oscillators has been presented. The nonlinearities of the transistor are modeled from S parameters and dc characteristic measurements. Then, nonlinear transfer functions $H_1, H_2, \cdots$ are calculated with the Volterra series formalism. Finally, the output power and the oscillation frequency are obtained by solving a determining equation. Comparison with measured data gives a discrepancy less than 10 percent. This good agreement between the theoretical and the experimental results shows that Volterra series provide an interesting tool for the analysis of a nearly sinusoidal oscillator. Moreover, using these nonlinear transfer functions, the intermodulation between noise sources and the signal at the oscillation frequency can be calculated and the oscillator output spectrum derived.

APPENDIX
DERIVATION OF TRANSFER FUNCTIONS H_1, H_2 AND H_3

For a single-loop nonlinear feedback system (Fig. 1(a)) the output $y(t)$ of its associated open loop (Fig. 1(b)) can be expressed as a Volterra series of the input $u(t)$ as

$$y(t) = \sum_{n=1}^{\infty} \int_{-\infty}^{\infty} \cdots \int h_n(\tau_1, \tau_2, \cdots, \tau_n) \prod_{i=1}^{n} u_i(t - \tau_i)\, d\tau_i. \tag{A1}$$

As a particular case, let the input to the system F be

$$u(t) = \sum_{i=1}^{M} A_i e^{j\omega_i t}. \tag{A2}$$

Substituting (A2) into the Volterra series (A1) gives the corresponding output of F:

$$\begin{aligned} y(t) &= \sum_{n=1}^{\infty} \int_{-\infty}^{+\infty} \int h_n(\tau_1, \tau_2, \cdots, \tau_n) \left(A_i e^{j\omega_i(t-\tau_1)}\right) \\ &\quad \cdot \left(\sum_{i=1}^{M} A_i e^{j\omega_i(t-\tau_2)} \right) \cdots d\tau_1 d\tau_2 \cdots d\tau_n \\ &= \sum_{n=1}^{\infty} \left(\sum_{i_1, i_2, \cdots, i_n = 1}^{M} \mathscr{H}_n(j\omega_{i_1}, j\omega_{i_2}, \cdots, j\omega_{i_n}) \right. \\ &\quad \left. \cdot A_{i_1} A_{i_2} \cdots A_{i_n} e^{j(\omega_{i_1} + \omega_{i_2} + \cdots + i_n)t} \right) \end{aligned} \tag{A3}$$

where $\mathscr{H}_n(j\omega_1, j\omega_2, \cdots, j\omega_n)$ is the nth-order transfer function of the system F and is also called the n-dimensional Fourier transform of $h_n(\tau_1, \tau_2, \cdots, \tau_n)$.

Applying formula (A3) to the system S (Fig. 2(b)) with input $u(t) = \frac{A}{2}e^{j\omega t} + \frac{A^*}{2}e^{-j\omega t}$, the output can be written as follows:

$$z(t) = \frac{A_z}{2}e^{j\omega t} + \frac{A_z^*}{2}e^{-j\omega t}$$
$$= \sum_{n=1}^{\infty}\left(\sum_{\iota_1, \iota_2 \cdots \iota_n = 1}^{2} n(j\omega_{\iota_1}, j\omega_{\iota_2} \cdots j\omega_{\iota_n}) A_{\iota_1} A_{\iota_2} \cdots A_{\iota_n}\right.$$
$$e^{j(\omega_{\iota_1} + \omega_{\iota_2} + \cdots + \omega_{\iota_n})t} \quad \text{(A4)}$$

where $\mathscr{H}_n$ is nth-order transfer function of the system S,

$$A_{\iota_k} = \frac{A}{2} \quad \text{or} \quad \frac{A^*}{2}$$

and $\omega_{\iota_k} = \pm\,\omega$ $k = 1, \cdots, n$, with n an odd integer.

Identifying the proper terms, we can write

$$\frac{A_z}{2}e^{j\omega t} = \mathscr{H}_1(j\omega)\frac{A}{2}e^{j\omega t}$$
$$+ \mathscr{H}_3(j\omega, j\omega, -j\omega)\frac{A}{2}\frac{A}{2}\frac{A^*}{2}e^{j(\omega+\omega-\omega)t}$$
$$+ \mathscr{H}_3(j\omega, -j\omega, j\omega)\frac{A}{2}\frac{A^*}{2}\frac{A}{2}e^{j(\omega-\omega+\omega)t}$$
$$+ \mathscr{H}_3(-j\omega, j\omega, j\omega)\frac{A^*}{2}\frac{A}{2}\frac{A}{2}e^{j(-\omega+\omega+\omega)t}. \quad \text{(A5)}$$

If the time origin is chosen such that the condition $A_z = A$ is verified, the nth-order determining equation is written as

$$d_N(A, \omega) = \mathscr{H}_1(j\omega) + \Omega_1(j\omega)|A|^2 + \Omega_2(j\omega)|A|^4$$
$$+ \cdots + \Omega_N(j\Omega)|A|^{2N} - 1 = 0 \quad \text{(A6)}$$

where $\mathscr{H}_1(j\omega) = H_1(j\omega)$ the first-order transfer function and $\Omega_\iota(j\omega)$ is a linear combination of the higher order transfer functions $H_2, H_3, \cdots, H_{2\iota+1}$.

These nonlinear transfer functions are then determined with the following method. According to (8) and (9), the linear parts of the drain–source and gate–source currents are

$$I_{dsL}(t) = g_{m1}v_{gs}(t) + g_{d1}v_{ds}(t) \quad \text{(A7)}$$

$$I_{gsL}(t) = \frac{d}{dt}\left[C_{gs1}v_{gs}(t)\right] \quad \text{(A8)}$$

with the shunt nonlinear current source

$$I_{dsNL}(t) = \sum_{k=2}^{3} i_{dsk}(t) = \sum_{k=2}^{3} |g_{mk}v_{gs}^k(t) + g_{dk}v_{ds}^k(t)| \quad \text{(A9)}$$

and

$$I_{gsNL}(t) = \sum_{k=2}^{3} i_{gsk}(t) = \frac{d}{dt}\sum_{k=2}^{3}\left[C_{gsk}v_{gs}^k(t)\right] \quad \text{(A10)}$$

respectively.

Now let $V_{gs}(t)$ and $V_{ds}(t)$ have a series development in

$$v_{gs}(t) = \sum_{n=1}^{\infty} v_{gsn}(t) \quad \text{(A11)}$$

$$v_{ds}(t) = \sum_{n=1}^{\infty} v_{dsn}(t) \quad \text{(A12)}$$

where $V_{gsn}(t)$ and $V_{dsn}(t)$ can be expressed as Volterra series of the input $E_g(t)$ as

$$v_{gsn}(t) = \int_{-\infty}^{+\infty} \cdots \int_{-\infty}^{+\infty} h_{gsn}(\tau_1, \cdots, \tau_n)\prod_{\iota=1}^{n} E_g(t-\tau_1)\,d\tau_1 \quad \text{(A13)}$$

$$v_{dsn}(t) = \int_{-\infty}^{+\infty} \cdots \int_{-\infty}^{+\infty} h_{dsn}(\tau_1, \cdots, \tau_n)\prod_{i=1}^{n} E_g(t-\tau_1)\,d\tau_1. \quad \text{(A14)}$$

Substituting (A11) and (A12) into (A9) and (A10) and collecting the proper terms yields

$$i_{ds2}(t) = g_{m2}v_{gs1}^2(t) + g_{d2}v_{ds1(t)}^2 \quad \text{(A15)}$$

$$i_{ds3}(t) = g_{m3}v_{gs1}^3(t) + 2g_{m2}v_{gs1}(t)v_{gs2}(t)$$
$$+ g_{d3}v_{ds1}^3(t) + 2g_{d2}v_{ds1}(t)v_{ds2}(t) \quad \text{(A16)}$$
$$\vdots$$

and

$$i_{gs2}(t) = \frac{d}{dt}\left[C_{gs2}v_{gs1}^2(t)\right] \quad \text{(A17)}$$

$$i_{gs3}(t) = \frac{d}{dt}\left[C_{gs3}v_{gs1}^3(t) + 2C_{gs2}v_{gs1}(t)v_{gs2}(t)\right] \quad \text{(A18)}$$
$$\vdots$$

When the excitation is specifically a sum of K distinct exponentials,

$$E_g(t) = \sum_{l=1}^{K} \exp(j2\pi F_l t)$$

the expressions in the frequency domain are given by the Fourier transform [6]:

$$I_{ds2}(F_1, F_2) = g_{m2}H_{gs1}(F_1)H_{gs1}(F_2)$$
$$+ g_{d2}H_{ds1}(F_1)H_{ds1}(F_2) \quad \text{(A19)}$$

$$I_{ds3}(F_1, F_2, F_3) = g_{m3}H_{gs1}(F_1)H_{gs1}(F_2)H_{gs1}(F_3)$$
$$+ 2g_{m2}S\left[H_{gs1}(F_1)H_{gs2}(F_2, F_3)\right]$$
$$+ g_{d3}H_{ds1}(F_1)H_{ds1}(F_2)H_{ds1}(F_3)$$
$$+ 2g_{d2}S\left[H_{ds1}(F_1)H_{ds2}(F_2, F_3)\right] \quad \text{(A20)}$$

and

$$I_{gs2}(F_1, F_2) = j2\pi(F_1 + F_2)C_{gs2}H_{gs1}(F_1)H_{gs1}(F_2)$$

$$I_{gs3}(F_1, F_2, F_3) = j2\pi(F_1 + F_2 + F_3)$$
$$\cdot\left\{C_{gs3}H_{gs1}(F_1)H_{gs1}(F_2)H_{gs1}(F_3)\right. \quad \text{(A21)}$$
$$\left.+ 2C_{gs2}S\left[H_{gs1}(F_1)H_{gs2}(F_2, F_3)\right]\right\} \quad \text{(A22)}$$

where

$$S[H_1(F_1)H_2(F_2,F_3)] = \frac{1}{3}[H_1(F_1)H_2(F_2,F_3) + H_1(F_2)H_2(F_3,F_1) + H_1(F_3)H_2(F_1,F_2)] \quad \text{(A23)}$$

with $H_1 = H_{gs1}$ or H_{ds1} and $H_2 = H_{gs2}$ or H_{ds2} respectively.

The node-pair method applied to the network in Fig. 4 gives the following matrix description between the voltage and current vectors V and I:

$$Y \times V = I \quad \text{(A24)}$$

where Y is an 8×8 admittance matrix (the nodes being referred to by numbers 1 to 8 in Fig. 4).

When the excitation is $E_g(t) = \exp(j2\pi Ft)$, the linear transfer functions $H_{gs1}(F)$, $H_{ds1}(F)$, and $H_1(F)$ in the frequency domain can be calculated by

$$\begin{bmatrix} H_{gs1}(F) \\ H_{ds1}(F) \\ H_1(F) \end{bmatrix} = [Y(F)]^{-1} \times \begin{bmatrix} Y_G E_G \\ 0 \\ 0 \end{bmatrix} \quad \text{(A25)}$$

Substituting $H_{gs1}(F)$ and $H_{ds1}(F)$ into (A19) and (A21), we can obtain the second-order transfer functions $H_{gs2}(F_1, F_2)$, $H_{ds2}(F_1, F_2)$, and $H_2(F_1, F_2)$ by the following equation:

$$\begin{bmatrix} H_{gs2}(F_1,F_2) \\ H_{ds2}(F_1,F_2) \\ H_2(F_1,F_2) \end{bmatrix} = [Y(F_1+F_2)]^{-1} \begin{bmatrix} 0 \\ -I_{gs2}(F_1,F_2) \\ -I_{ds2}(F_1,F_2) \end{bmatrix} \quad \text{(A26)}$$

Higher order transfer functions can then be calculated from (A20) and (A22) using a recursive method of the type described in [7] and [8].

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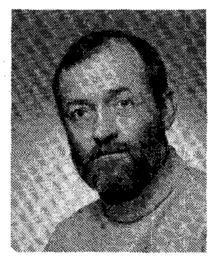

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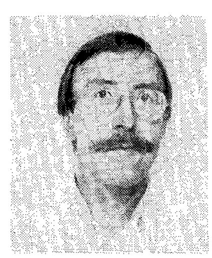

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