Signal Space and Its Implementation for Microwave Nonlinear-Network Analysis

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***Abstract*— Today, harmonic balance is the most useful approach for microwave nonlinear-network analysis. Fast Fourier transform (FFT) is usually used to convert the nonlinear-elements' time waveforms into the frequency domain as part of the harmonic-balance process. This approach is straightforward for a single-frequency excitation, but is quite complicated and time consuming for the multifrequency-excitation case. In this paper, we propose a mathematical model termed "signal space," which enables (for a given nonlinearity) a direct calculation of the currents spectrum (given the voltages spectrum) and is suited for implementation for harmonic balance. Under describing-function criteria, we use the signal-space approach to get expressions for oscillator parameters such as oscillation frequency, stability condition, and injection-locking bandwidth. There is a good agreement between our results and Kurakawa's expressions.**

I. Introduction

ANALYSIS of nonlinear microwave networks is a very useful tool for most practical microwave components, and several approaches have been developed in both time and frequency domains. Undoubtedly, the frequency-domain harmonic-balance approach is today the most useful in microwave computer-aided design (CAD) software. Rhyne and Steer [1] proposed the generalized power series, which enables, for a given nonlinearity, a direct calculation of the currents spectrum given the voltages spectrum. In [2], we proposed a new mathematical method termed "signal space" for analyzing nonlinear networks. The signal space, which has the same basis as Steer's suppositions, is a polynomial's phasor vector space, in which each harmonic current can be described by a linear combination of base polynomials, and then enables, for a given nonlinear device, a direct calculation of the currents spectrum given the voltages' spectrum. The expressions for the harmonic currents, which appear in Section I, are simple and suited to implementation for harmonic balance.

Several researchers have implemented nonlinear analysis of free-running oscillators using describing-function techniques [6]–[8]. In these techniques, the system equations and oscillation criteria are combined to yield a set of algebraic equations. In this paper, we applied describing-function criteria to signal space for design of a parallel-resonator oscillator based on negative conductance. Under these conditions, we get expressions for the frequency of oscillation, stability condition, injection-locking bandwidth, and other oscillator parameters. Our results are compatible with Kurokawa's expressions and a Gunn oscillator design.

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II. The Signal Space

We demonstrate the method by considering a nonlinear two-port network. The relation between the time-domain currents and voltages for a memoryless two-port network is represented generally as

$$I_1 = f_1(V_1, V_2) \quad I_2 = f_2(V_1, V_2) \tag{1}$$

where f_1 and f_2 are functions representing the nonlinear device, and V_1 and V_2 are the voltages of the nonlinear device where I_1 and I_2 are the currents of the nonlinear device. We assume that both voltages are composed of a signal at frequency ω and its harmonics, namely, a periodic function

$$\begin{aligned} V_1(t) &= \sum_{k=-n}^{k=n} V_{1k} \exp(jk\omega t) \\ &= \sum_{k=0}^{k=n} V_{1Pk} \exp(jk\omega t) + V_{1Nk} \exp(-jk\omega t) \\ V_2(t) &= \sum_{k=-n}^{k=n} V_{2k} \exp(jk\omega t) \\ &= \sum_{k=0}^{k=n} V_{2Pk} \exp(jk\omega t) + V_{2Nk} \exp(-jk\omega t) \end{aligned} \tag{2}$$

where

$$V_{Pk} = \frac{\mathrm{Re}(V_k) - j\,\mathrm{Im}(V_k)}{2}$$
$$V_{Nk} = \frac{\mathrm{Re}(V_k) + j\,\mathrm{Im}(V_k)}{2}$$

where P and N indexes indicate a positive and negative contributions for the harmonic number of the current, respectively. The indexes 1 and 2 indicate ports 1 and 2, respectively, k indicates harmonic number, and n is the number of harmonic considered. Assume that f_1 and f_2 can be expressed as a Taylor series, namely,

$$f(V_1, V_2) = \sum_{i=0}^{\infty} \sum_{j=0}^{\infty} \frac{1}{i!j!} * \frac{\partial f^{i+j} f(V_1, V_2)}{\partial V_1^i \partial V_2^j} * V_1^i V_2^j. \tag{3}$$

Substitute the voltages, $V_1(t)$ and $V_2(t)$ from (2) to (3), using the multinominal formula, and arranging the result to groups of polynomials with the same harmonic number F.

The expression for the current at harmonic F consists of polynomials with a general form

$$I_{1F} = \sum_D \sum_E^{T_{\rm MAX}} G_{DE} \prod_{k=0}^n V_{1Pk}^{B_k} \prod_{k=0}^n V_{2Pk}^{Q_k} \prod_{k=0}^n V_{1Nk}^{T_k} \cdot \prod_{k=0}^n V_{2Nk}^{S_k} \exp(jF\omega t) \quad (4)$$

where

$$\sum_k B_k + \sum_k T_k = D$$
$$\sum_k Q_k + \sum_k S_k = E. \quad (5)$$

The sum $T_{\rm MAX}$ is the maximum order of the truncated tailor series used in practice, and the summation in (4) is done on all the combinations which appear in (5) for given D and E. n is the number of harmonics considered and F is the harmonic number of the considered term

$$F = \left(\sum_{k=0}^n k * B_k + \sum_{k=0}^n k * Q_k - \sum_{k=0}^n k * T_k - \sum_{k=0}^n k * S_k\right). \quad (6)$$

Polynomials of the above form represent a vector space that we term "signal space." All the polynomials with a specific harmonic number F form a subspace of the signal space. Thus, every one of the harmonic currents of the nonlinear device can be expressed as a vector in the signal space. The coefficients G_{DE} are calculated by the general formula

$$G_{DE} = \frac{\partial^{(D+E)} f(0, 0)}{\partial V_1^D \partial V_2^E} \cdot \frac{1}{\prod_{k=0}^n B_k! \prod_{k=0}^n T_k! \prod_{k=0}^n Q_k! \prod_{k=0}^n S_k!}. \quad (7)$$

The same idea can be implemented for a one-port device. The expression for the current at harmonic F consists of polynomials of a general form

$$I_{\rm GF} = \sum_D^{T_{\rm MAX}} G_D \prod_{k=0}^n V_{Pk}^{B_k} \prod_{k=0}^n V_{Nk}^{T_k} \exp(jF\omega t) \quad (8)$$

where

$$\sum_{k=0}^n B_k + \sum_{k=0}^n T_k = T_{\rm MAX} \quad (9)$$

and

$$F = \omega\left(\sum_{k=0}^n k \cdot B_k - \sum_{h=0}^n k \cdot T_k\right). \quad (10)$$

The coefficients G_D are

$$G_D = \frac{\partial^D f(0)}{\partial V^D} \frac{1}{\prod_{k=0}^n B_k! \prod_{k=0}^n T_k!}. \quad (11)$$

Thus, for the nonlinear device, it is possible to analytically calculate the spectral content of the device current given the spectral content of the device voltage.

The above idea is readily applicable for a reactive nonlinear device. Consider, for example, a nonlinear capacitor for which the current–voltage relation is represented by

$$I_C = dQ/dt = dQ/dV * dV/dt = C(V)\, dV/dt.$$

Following the analysis above, and assuming a periodic voltage waveform, one can derive the expressions for the current spectral content expressed as vectors in the signal space. The expression for the capacitance $C(V)$ in a Taylor series is

$$C(V) = \frac{\partial Q}{\partial V} = \sum_{T=0}^{T_{\rm MAX}} C_T \cdot V^T \quad (12)$$

and its expression in the signal space

$$C(V) = \sum_{F=0}^n \sum_{T=0}^{T_{\rm MAX}} C_{\rm TPF} \prod_{k=-n}^n V_k^{P_{Fk}} \cdot \exp(jF\omega t) + \sum_{F=0}^n \sum_{T=0}^{T_{\rm MAX}} C_{\rm TNF} \prod_{k=-n}^n V_k^{N_{Fk}} \cdot \exp(-jF\omega t) \quad (13)$$

where

$$C_{\rm TPF} = \frac{C_T \cdot T!}{\prod_{k=-n}^n P_{Fk}!}$$
$$\sum P_{Fk} = T$$
$$C_{\rm TNF} = \frac{C_T \cdot T!}{\prod_{k=-n}^n N_{Fk}!}$$
$$\sum N_{Fk} = T, \qquad T = 1, 2, 3, \cdots, T_{\rm MAX}.$$

The expression for the F harmonic-capacitor current is the result derived by multiplication of $C(V)$ in (13) by the expression of the voltage time derivative and is

$$\begin{aligned} I_{CF} &= C(V)\frac{\partial V}{\partial t} \\ &\cong \Bigg[-j\omega\left(\sum_{T=1}^{T_{\rm MAX}} C_{TP(F+1)} \prod_{k=-n}^n V_k^{P_{(F+1)k}}\right)V_{N1} \\ &\quad + j\omega\left(\sum_{T=1}^{T_{\rm MAX}} C_{TN1} \prod_{k=-n}^n V_k^{N_{1k}}\right)(F+1)V_{P(F+1)} \\ &\quad + j\omega C_0 \cdot F \cdot V_{PF}\Bigg] \exp(jF\omega t) \end{aligned} \quad (14)$$

where

$$C_0 = \frac{C(0)}{2}.$$

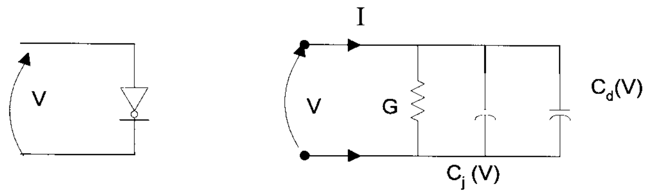

Fig. 1. A diode and its equivalent circuit.

A nonlinear device composed of a time delay τ has harmonic current of the general form

$$\begin{aligned} I &= f[V(t-\tau)] \\ &= \sum \left[C \cdot \prod_{k=0}^{n} V_{Pk}^{B_k} \exp(-jk\omega\tau) \cdot \prod_{k=0}^{n} V_{Nk}^{T_k} \exp(-jk\omega\tau) \right] \\ &\quad \cdot \exp(jF\omega t). \end{aligned} \tag{15}$$

III. Diode Representation in Signal Space

Using the expressions described above, we derive the representation of a diode in the signal space.

A simple diode and its equivalent circuit are shown in Fig. 1. The equivalent circuit consists of a conductance G, diffusion capacitance $C_d(V)$, and depletion capacitance $C_j(V)$ [11]. The expression for the conductance current is

$$I_G = I_{SS} \exp(qV/KT). \tag{16}$$

The expression for the F harmonic of its current is described in (8). Where the coefficients G_D are

$$G_D = I_{SS} \frac{(q/KT)^{\sum_{k=0}^{n}(B_k+T_k)}}{\prod_{k=0}^{n} B_k! \prod_{k=0}^{n} T_k!}. \tag{17}$$

The expression for the diffusion capacitance is

$$\frac{dQ}{dV} = C_d(V) = A \cdot \exp(qV/KT) = C_{ss}. \exp(\beta V). \tag{18}$$

Its F harmonic current is described in (13) and (14), where the coefficients C_T are

$$C_T = \frac{\beta^T}{T!}. \tag{19}$$

The expression for the depletion capacitance is

$$C_j(V) = \frac{C_{00}}{(1+bV)^g}. \tag{20}$$

This function can be approximated by curve tracing to the tailor series, as appears in (12).

Its F harmonic current is described in (13) and (14).

The current at harmonic F is the sum of the three terms

$$I_F = I_{\mathrm{GF}} + I_{\mathrm{CdF}} + I_{\mathrm{CjF}} \tag{21}$$

and in the time domain

$$I_F(t) = 2\,\mathrm{Re}[I_F \cdot \exp(jF\omega t)].$$

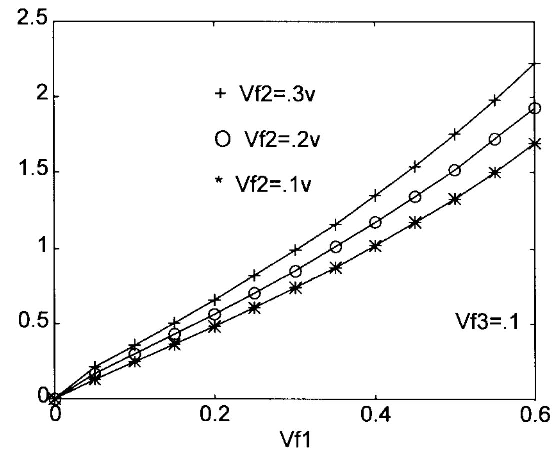

Fig. 2. I_{F1} current as a function of V_{F1} in the FFT method and signal space while varying parameter V_{F2}.

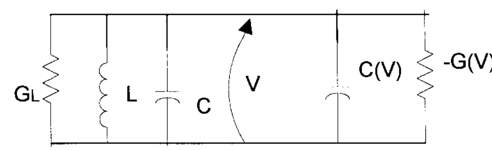

Fig. 3. Parallel resonant oscillator.

IV. Signal-Space Accuracy

To investigate the accuracy of the signal-space approach, we calculated the harmonic of the current of a simple diode using the signal-space process and compared the results to the fast Fourier transform (FFT) process. The expression of the diode current is $I = I_s * \exp(\beta V)$. Suppose $\beta = 2$ and $I_s = 10^{-12}\ A$, one could approximate this expression to the fifth power of a Taylor series. Suppose also that there are three harmonics of voltages V_{F1}, V_{F2}, V_{F3}. Using (4)–(7), we calculated the first current harmonic I_{F1} as a function of V_{F1} in the signal space and the FFT method, while V_{F3} is a constant equal to 0.1 V and V_{F2} as a varying parameter between 0.1–0.3 V. From the results in Fig. 2, there is a good agreement between the two methods. The signal-space method was tested using other nonlinear elements and was found compatible to the FFT method. We get the same accuracy for extreme nonlinear elements using the signal space compared to the standard FFT process. However, as the element is more nonlinear, its presentation in the signal space consists of more polynomials, and the computing time is longer.

V. Oscillator Analysis and Design

The signal-space approach can be used for analytic investigation of microwave oscillator networks. We have derived a method to analyze a negative-resistance oscillator, which yields design guide rules similar to the well-known analysis approach of Kurakawa. Consider the network in Fig. 3. The nonlinear device is represented by a nonlinear negative resistance and a nonlinear capacitance $(-G(V),\ C(V))$. The device is resonated by the resonant circuit L, C and G_L is the load conductance.

To analyze this circuit, we make the commonly used assumption that the voltage V contains only the fundamental component at frequency ω. The other voltage harmonics are assumed negligible due to the filtering action of the resonant circuit. Under these conditions, we use the signal-space approach to get expressions for the current spectrum components. In this case, the expression for the first harmonic current of the negative resistor is

$$I_{G1} = \sum_{k=0}^{T_{\rm MAX}} a_{2k+1} \frac{(2k+1)!}{k!(k+1)!} \cdot V_{p1}^{k+1} \cdot V_{N1}^{k}. \qquad (22)$$

It is obvious that the voltage phase can be arbitrarily chosen, thus, for a phase equal to zero, $V_{p1} = V_{N1} = V_R/2$, and (22) reduces to

$$I_{G1} = \frac{1}{2} \sum_{k=0}^{T_{\rm MAX}} a_{2k+1} \frac{(2k+1)!}{k!(k+1)!} \cdot V_R^{2k+1}. \qquad (23)$$

The coefficients a_{2k+1} have to be calculated using the expressions of [2] and [3] and the Gunn model.

Under the same conditions, the expression for the first harmonic current of the nonlinear capacitor is

$$I_C = -j\omega_1 \sum_{k=0}^{T_{\rm MAX}} \bigcup(k) \frac{C_k \cdot k!}{\left(\frac{k+2}{2}\right)! \left(\frac{k-2}{2}\right)!} \cdot V_R^{k+1} \qquad (24)$$

where

$$U(k) = \begin{cases} 1, & \text{for } k = 0 \\ -1, & \text{for } k = 2p \\ 0, & \text{for } k = 2p+1. \end{cases}$$

The coefficients C_k have to be calculated using the expressions of [2] and [3] and the Gunn model.

The current into the resonator for the same voltage is

$$I_{\rm RES} = V_R \cdot G_L + j2QG_L \cdot V_R \cdot \frac{\Delta\omega}{\omega_o} \qquad (25)$$

where ω_o is the resonant frequency and $\Delta\omega = \omega_0 - \omega$.

Using Kirchoff's law, the summation of the device and load currents must equal zero as follows:

$$I_{\rm RES} + I_C + I_{G1} = 0. \qquad (26)$$

Using this condition and separating the real and imaginary parts, the oscillation frequency ω_1 and amplitude V_R are found to be (27) and (28), shown at the bottom of this page, where C_k represents the nonlinear-device capacitor coefficients in the signal space. As expected, the oscillation frequency equals ω_o for $C_k = 0$.

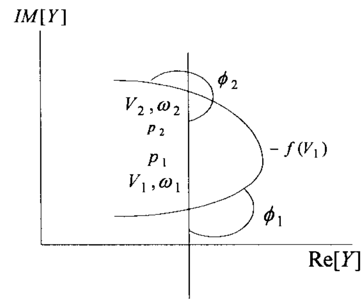

Fig. 4. The load $Y(j\omega)$ and device $f(V_1)$ curves in the complex plane.

Expressions (27) and (28) give the amplitude and frequency oscillation of a parallel resonant oscillator circuit, and consists of a nonlinear negative resistor and nonlinear capacitor loaded by a linear resonant circuit, as shown in Fig. 3. a_{2k+1} and C_k represent the nonlinear resistor and capacitor in the signal space.

A. Oscillator Stability Condition

Using the Kurakawa stability condition [9] or the nonlinear Nyquist [2], [10] condition for a parallel resonant circuit, the phase deference between the device and load curves [9] has to be 90°. The expression for the angle is

$$\tan(\phi) = \frac{\omega \sum_{k=o}^{T_{\rm MAX}} \frac{(k+1)!}{\left(\frac{k+2}{2}\right)! \left(\frac{k-2}{2}\right)!} \bigcup^1(k) \cdot C_k V_R^k}{\sum_{k=0}^{T} a_{2k+1} \frac{2k(2k+1)!}{k!(k+1)!} V_R^{2k}} \qquad (29)$$

where

$$\bigcup^1(k) = \begin{cases} 0, & k = 2p \\ -1, & k = 2p+1 \end{cases}$$
$$p = 0, 1, 2, \cdots.$$

Using (29), the phase deference between the device and load curves ϕ in an intersection point must be in the range $-(\pi/2) < \phi < \pi/2$.

In Fig. 4, there is an example of the load and device curve in the complex plane. There are two points at which the circuit can oscillate, but according to condition (29), only $P1$ is a stable point.

$$-2 \sum_{k=0}^{T_{\rm MAX}} a_{2k+1} \frac{(2k+1)!}{k!(k+1)!} V_R^{2k} = G_L \qquad (27)$$

$$\omega_1 = \frac{\omega_o}{1 + \left(\frac{\omega_o}{QG_L}\right) \cdot \sum_{k=0}^{T_{\rm MAX}} \bigcup(k) \frac{C_k \cdot k!}{\left(\frac{k+2}{2}\right)! \left(\frac{k-2}{2}\right)!} \cdot V_R^{k+1}} \qquad (28)$$

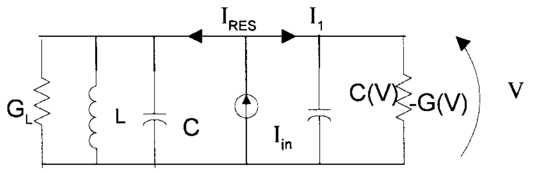

Fig. 5. The circuit diagram of injection-locked oscillator with I_{in} as the inject signal.

B. Injection Locking

Using the signal-space approach, we derived the expression for the injection-locking bandwidth. Referring to Fig. 5, the injected signal appears as I_{in}. By solving the current Kirchoff's law using the signal-space expressions for the different currents, one can express $\Delta\omega = \omega_{\mathrm{osc}} - \omega_0$ as a function of the other parameters. The expression for the injection-locking bandwidth is

$$\Delta\omega = \frac{\omega_o}{2Q}\,\frac{|I_{\mathrm{in}}|\,\sin\phi}{I_L} - \frac{\omega_o^2}{Q\cdot I_L} \cdot \sum \frac{\bigcup(k) C_k \cdot k!}{\left(\frac{k+2}{2}\right)!\left(\frac{k-2}{2}\right)!} V_R^{k+1} \qquad (30)$$

where I_L is the load current.

The first term is known as Adler's formula; thus, (30) is an extension of Adler's expression for a nonlinear device. The second term in this expression has more influence since the device is more nonlinear.

C. Gunn-Device Modeling and Results

The signal-space approach has been used to design an oscillator with a parallel resonator. The design was based on a Gunn device, which is coupled to a microstrip resonator. In the first step, we characterized the device in the signal space as a negative nonlinear resistor connected to a nonlinear capacitor. An M/A Gunn device MA49106 was inserted into a coaxial resonator, and its load was changed by an EH tuner to avoid oscillations. We measured the reflection coefficients of the loaded diode for different power levels (0, 5, and 10 dBm) and derived the Gunn-diode impedance for each input power. In the second step, we changed the EH tuner to allow oscillations at 9.8 GHz for different power outputs in the range of 12–16 dBm. The device impedance is the complex conjugate of the load impedance, which was measured for each power output. The diode impedance for each power was measured in the two steps used to translate the device to the signal space by calculating a_{2k+1} and C_k coefficients of the negative conductor and the capacitor accordingly, as shown in (23) and (24). Using (27)–(30) for the circuit in Fig. 3, the theoretical oscillation frequency and oscillation power were 9.8 GHz and 15.8 dBm, respectively. The experimental results of 9.8 GHz and 16.2 dBm we obtained, respectively. Our theoretical injection-locking bandwidth was 32 MHz, and we obtained an experimental value of 36 MHz.

VI. Conclusion

In this paper, we present a novel analytical approach for the analysis of nonlinear networks. The method presented belongs to the general category of a frequency-domain harmonic-balance algorithm. The main contribution of the new approach is an analytic and efficient calculation of the nonlinear-device current spectral components given the voltage spectral components and the characteristics of the nonlinear device. The new approach is based on a vector space termed the signal space.

The signal-space approach has been successfully used to derive analysis and design criteria for a negative-resistance oscillator, and we found a good agreement between theory and experiment.

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Dr. Madjar has served as the Israel IEEE S-AP/MTT chapter chairman for several years and, in that capacity, organized 13 symposia. From 1985 to 1989, he served as the secretary of the Israel Section of the IEEE. He served on the technical committees for MELECON (1981), 14th, 15th, 16th, 18th, and 19th conventions of the Electrical and Electronics Engineers (IEE), Israel.