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LINEAR AND NONLINEAR STABILITY ANALYSIS OF MMICs

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Abstract:

Stability analysis of MMICs becomes a preoccupation of engineers for the design of power amplifiers as well as of oscillators. In this paper it is shown that the classical approach of linear stability analysis fails in the case of unstable unloaded circuits. Nyquist based methods are proposed to circumvent this problem. A method relying on the general Harmonic Balance formulation is proposed, which enables the linear and non linear stability analysis. Examples of global stability analysis based on the resolution of the bifurcation equations are given. Finally an open loop method is proposed which allows to design analog frequency dividers with the aid of standard CAD packages.

1- Introduction

Recent developments of integrated circuits for RF and microwave applications require powerful simulation methods as any change in the manufactured circuits is no longer possible. Among the critical features of integrated circuits, the stability properties of those circuits are essential. Indeed electrical instabilities can occur in various types of circuits such as power amplifiers, mixers and free running or synchronized oscillators; the latter constituting a class of potentially unstable circuits. Instabilities in electronic circuits can appear as oscillatory behaviour for power amplifiers under constant bias conditions or as frequency or amplitude jumps under the application of a microwave signal. The former instabilities are referred as linear instabilities while the latter are referred as nonlinear instabilities. Both types lead to the failure of the design of MMICs.

Up to now instabilities have been studied by elementary methods which do not take into account all the effects involved in the behavior of microwave circuits and which can be very complex. For linear stability analysis the concept of stability factor (Linvil Factor) [1] is generally used. It has been shown [2] that this concept is unsufficient for predicting the linear stability of some multi devices circuits. In the case of nonlinear stability analysis there exist a very few methods which are generally applied to oscillators and which rely on the description function approach.

In this paper we will present the limitations of the classical approach. Then we will present a global stability analysis method, based on the Harmonic Balance algorithm, which allows to predict the stability of microwave circuits under linear and nonlinear conditions. This method has been applied to power amplifiers as well as to oscillators and frequency dividers.

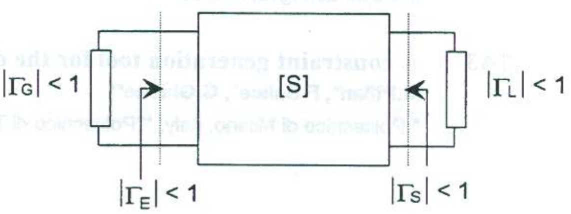
2- Linear stability analysis

2-1 Limitations of the K factor analyisis

Classically the linear stability analysis is performed by the calculation of a number of factors derived from the [S] parameters of the device. For the linear two ports of fig-1 the stability is determined by the values of the reflection coefficients seen at the input ant output of the netrwork. The two conditions stated in Fig-1 are fullfilled for any source and load reflection coefficients if the stability factor K>1 and an auxiliary condition are met. For linear two ports those conditions read:

$$K = \frac{\left(1 - |S_{11}|^2 - |S_{22}|^2 + |\Delta_s|^2\right)}{2 \cdot |S_{12}| \cdot |S_{21}|} > 1 \qquad (1)$$

With $|\Delta_s| = |S_{11} \cdot S_{22} - S_{12} \cdot S_{21}| < 1$



$$|\Gamma_E| < 1 \text{ with : } \Gamma_E = S_{11} + \frac{S_{12} \cdot S_{21} \cdot \Gamma_L}{1 - S_{22} \cdot \Gamma_L} \qquad |\Gamma_L| < 1$$

$$|\Gamma_S| < 1 \text{ with : } \Gamma_S = S_{22} + \frac{S_{12} \cdot S_{21} \cdot \Gamma_G}{1 - S_{11} \cdot \Gamma_G} \qquad |\Gamma_G| < 1$$

Fig-1 Configuration of a linear two ports for stability analysis.

Unfortunately those conditions are only valid if the circuit is stable when unloaded. Indeed the reduction of a multinode circuit to a two ports network can lead to the cancellation of poles of the transfer function which have a positive real part. This fact seriously limits the validity of the classical approach.

2-2 New approaches for linear stability analysis

Recently several approaches have been proposed in order to circumvent the limitations involved in the application of the K factor stability criterion [3], [4] They rely on the application of the Nyquist criterion to a well chosen network function. Following Bode approach [5] a network containing one active element and a feedback network can be represented as shown in Fig-2-a. The stability of the network is ensured if all the real parts of the zeroes of F(p) are negative. Generally the stability is analysed by applying the Nyquist criterion to the open loop function $\mu \cdot \beta(j\omega)$. **If P is the number of unstable poles of $\mu \cdot \beta(j\omega)$ the system will be stable if the number of counterclockwise encirclements of the critical point (1,0) is equal to P.**

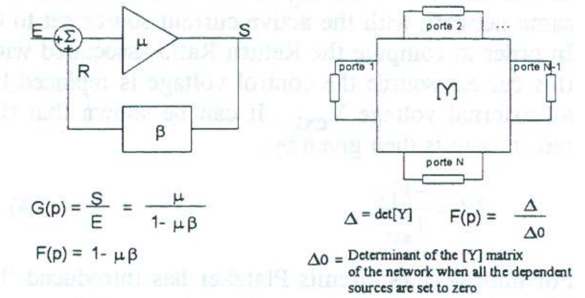


Fig-2 Representation of a closed loop system (a) and the generalisation to a multinode electrical network (b)

2-2 Return Ratio and Normalized Determinant Function

One of the key point of Bode analysis is the identification of the main loop of the circuit. In some cases it can be difficult to perform this task so that Bode proposes the concepts of Return Difference and Return Ratio. In a single loop system the Return Difference RD represents the feeback coefficient of the system $1-\mu \cdot \beta(j\omega)$. The Return Ratio RR is defined as minus the open loop gain of the system i.e: $-\mu \cdot \beta(j\omega)$, so that the relation between the return difference and the return ratio reads:

$$RR = 1 + RD \qquad (2)$$

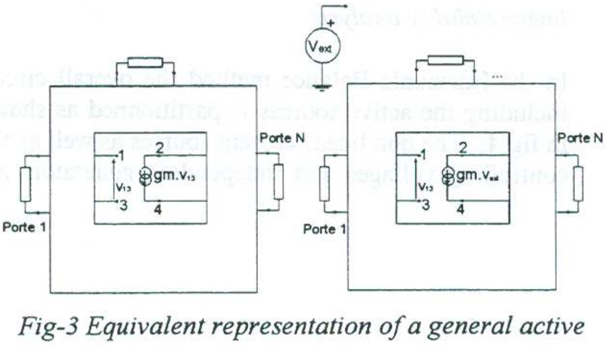


Fig-3 Equivalent representation of a general active

For a multi loop system the active elements are isolated in order to define Return Ratio relatively at

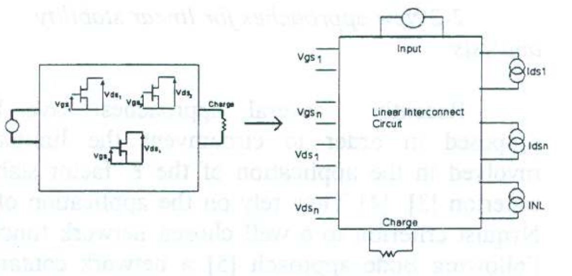


Fig-4 General representation of an active circuit for HB analysis

each non linear element. Consider the circuit of the fig-3 where we have extracted one particular voltage controlled current source. The return difference for this current source is given by:

$$RD = \frac{\Delta}{\Delta_0} \qquad (3)$$

Where Δ is the determinant of the [Y] matrix of the network while Δ_0 is the determinant of the same network with the active current source set to 0. In order to compute the Return Ratio associated with this current source the control voltage is replaced by an external voltage V_{ext} . It can be shown that the return ratio is then given by:

$$RR = \frac{-V_{13}}{V_{ext}} \qquad (4)$$

For multidevices circuits Platzker has introduced the Normalized Determinant Function which is a generalisation of the Return Difference first defined by Bode. This NDF can be expressed as a product of elementary Return differences. The NDF can then be expressed as shown in eq (5).

$$NDF = \frac{\Delta}{\Delta_0} = (RR_1 + 1)\cdot(RR_2 + 1)\cdots(RR_N + 1) \quad (5)$$

Where RR_1 is the return ration calculated with the first source set to 0, RR_2 is the return ratio with the first and second sources set to 0 ,...RR_k the return ratio with the first k sources set to 0 and so one until the ciruit is completely passive. As RR_k can be readily calculated using standard CAD softwares it is possible to obtain the NDF of an MMIC and then a Nyquist analysis of the NDF provides useful informations about the stability of the circuit. This technique allows to circumvent the problems encountered by the application of the K factor method.

2-3 Harmonic balance Formulation of the linear stability analysis

In the Harmonic Balance method the overall circuit including the active sources is partitionned as shown in fig-4. The non linear current sources as well as the controlling voltages and independent generators are fact seriously limits the validity of the classical approach.

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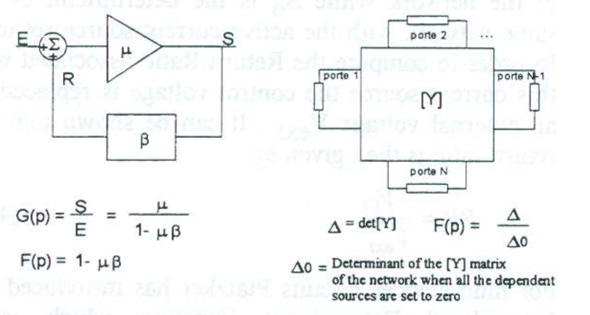

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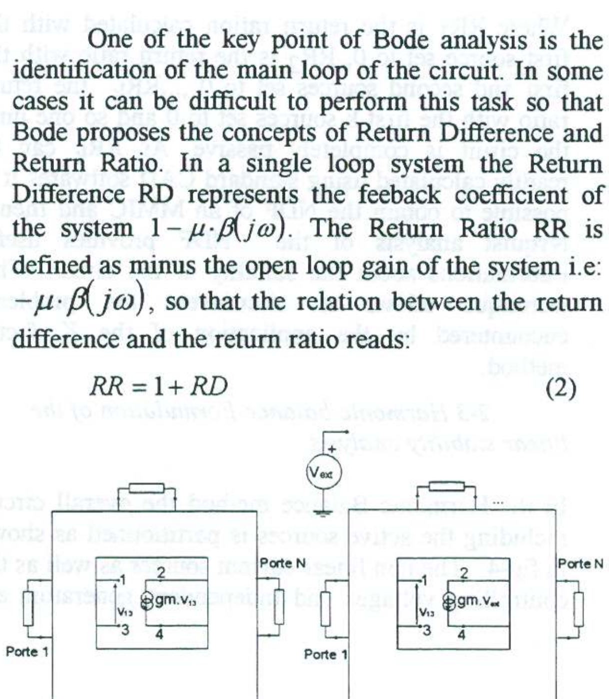

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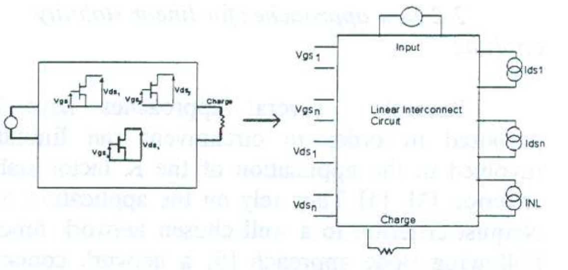

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2-3 Harmonic balance Formulation of the linear stability analysis

In the Harmonic Balance method the overall circuit including the active sources is partitionned as shown in fig-4. The non linear current sources as well as the controlling voltages and independent generators are extracted from the embedding network. The non linear network equation links the non linearities and the controlling voltages in a matrix form as given in eq (6)

$$
\begin{bmatrix} V_{GS1} \\ \cdot \\ \cdot \\ V_{GSn} \\ V_{DS1} \\ \cdot \\ V_{DSn} \end{bmatrix}
-
\left[\begin{array}{cccc|c} Z_{G11} & \cdot & \cdot & Z_{G1n} & Z_{GNL1} \\ \cdot & \cdot & \cdot & \cdot & \cdot \\ \cdot & \cdot & \cdot & \cdot & \cdot \\ Z_{Gn1} & \cdot & \cdot & Z_{Gnn} & Z_{GNLn} \\ \hline Z_{D11} & \cdot & \cdot & Z_{D1n} & Z_{DNL1} \\ \cdot & \cdot & \cdot & \cdot & \\ Z_{Dn1} & \cdot & \cdot & Z_{Dnn} & Z_{DNL1} \end{array}\right]
\times
\begin{bmatrix} I_{DS1} \\ \cdot \\ \cdot \\ I_{DSn} \\ I_{NL} \end{bmatrix}
-
\begin{bmatrix} E_{G1} \\ \cdot \\ \cdot \\ E_{Gn} \\ E_{D1} \\ \cdot \\ E_{Dn} \end{bmatrix}
= 0
\tag{6}
$$

In a non linear resolution process the previous equation is given for each frequency involved in the non linear steady state regime. For linear stability analysis it is sufficient to consider the bias conditions together with a perturbation signal of frequency ω. Linearisation of eq (6) around the bias condition gives the matrix equation (7)

$$
\begin{bmatrix} dV_{GS1} \\ \cdot \\ \cdot \\ dV_{GSn} \\ dV_{DS1} \\ \cdot \\ dV_{DSn} \end{bmatrix}
-
\left[\begin{array}{cccc|c} Z_{G11} & \cdot & \cdot & Z_{G1n} & Z_{GNL1} \\ \cdot & \cdot & \cdot & \cdot & \cdot \\ \cdot & \cdot & \cdot & \cdot & \cdot \\ Z_{Gn1} & \cdot & \cdot & Z_{Gnn} & Z_{GNLn} \\ \hline Z_{D11} & \cdot & \cdot & Z_{D1n} & Z_{DNL1} \\ \cdot & \cdot & \cdot & \cdot & \\ Z_{Dn1} & \cdot & \cdot & Z_{Dnn} & Z_{DNLn} \end{array}\right]
\times
\left[\begin{array}{ccc|ccc} G_{m1} & \cdot & \cdot & G_{d1} & \cdot & \cdot \\ \cdot & \cdot & \cdot & \cdot & \cdot & \cdot \\ \cdot & \cdot & G_{mn} & \cdot & \cdot & G_{dn} \\ \hline G_{nlg1} & \cdot & G_{nlgn} & G_{nld1} & \cdot & G_{nldn} \end{array}\right]
\begin{bmatrix} dV_{GS1} \\ \cdot \\ \cdot \\ dV_{GSn} \\ dV_{DS1} \\ \cdot \\ dV_{DSn} \end{bmatrix}
= 0
\tag{7}
$$

Previous equation can also be written in the compact form:

$$
(\mathbf{I_d} - \mathbf{Z} \cdot \mathbf{G}) \cdot \mathbf{v} = \mathbf{0} \tag{8}
$$

Where $\mathbf{I_d}$ is the identity matrix **Z, G** are the matrices of eq(7) and where **v** is the vector of perturbation voltages.

The matrix system given in eq (8) has a structure which is formally similar to that of the single loop circuit as shown in Fig-5.

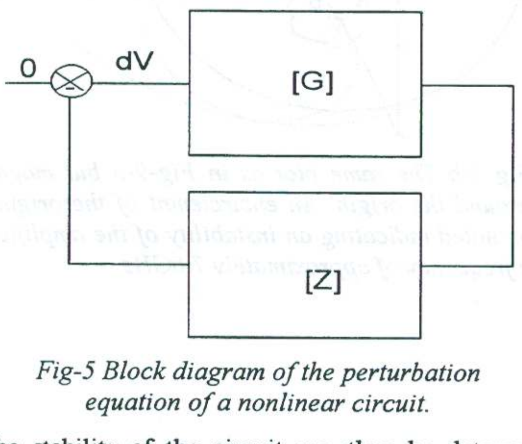


Fig-5 Block diagram of the perturbation equation of a nonlinear circuit.

The stability of the circuit can then be determined rigorously by applying the Nyquist criterion to the characteristic locus which are constituted of the locus of eigen values of eq (8) defined on a Riemann surface [6]. However finding the eigenvalues of the characteristic system is a difficult task so that, provided there is no cancellation of poles and zeros, the stability analysis can be performed by applying the Nyquist criterion to the determinant of the characteristic system.[7]

2-4 Examples

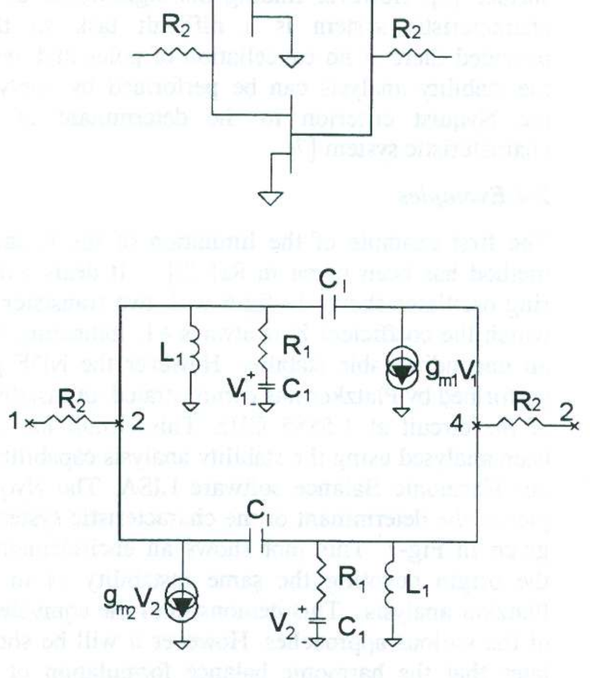
The first example of the limitation of the K factor method has been given in Ref [2] . It deals with a ring oscillator shown in fig-6 with two transistor for which the coefficient K is always >1, indicating thus an unconditionable stability. However the NDF plot performed by Platzker has demonstrated an instability of the circuit at 1.5885 GHz. This circuit has also been analysed using the stability analysis capability of our Harmonic Balance software LISA. The Nyquist plot of the determinant of the characteristic system is given in Fig-7. This plot shows an encirclement of the origin denoting the same instability as in the Platzker analysis. This demonstrates the equivalence of the various approaches. However it will be shown later that the harmonic balance formulation of the stability analysis can be extended to nonlinear cases thus providing us with a powerful tool for local and global nonlinear stability analysis of MMIC.

The second example concerns a two stage MMIC power amplifier realized with 0.7 μm gate width MESFETS. The structure of the amplifier is given in fig-8. The NDF plot shown in fig-9 denotes an instability of the amplifier for the bias point (V_d = 3Volts and V_g = -2.4V).

3- Nonlinear Stability analysis

Another class of instabilities which occur under RF signal excitation has to be analysed in order to be able to predict the whole nonlinear behaviour. This is true for power amplifiers where spurious tones can appear in the dynamic regime which can be synchronized at half the frequency of the input signal; the power amplifier behaving then as an analog frequency divider. This is also true for potentially unstable circuits such as free running or synchronized oscillators and frequency dividers.

The nonlinear stability analysis rely on a particular formulation of the Harmonic Balance equation which allows the determination of bifurcation points corresponding to qualitative changes in the behaviour of the circuit. Frequency and amplitude jumps, frequency division, frequency synchronization and apparition of spurious tones are various manifestations of those bifurcations. However it is still possible to analyze the behavior of potentially unstable circuits with standard CAD packages by applying the open loop concept.



R1=10 Ω ; R2=50 Ω ; L1=560pH ; C1=16pF
Cl=0.1pF ; gm1= 500mS ; gm2= 400mS;

Fig-6 Ring oscillator chosen for linear stability analysis.

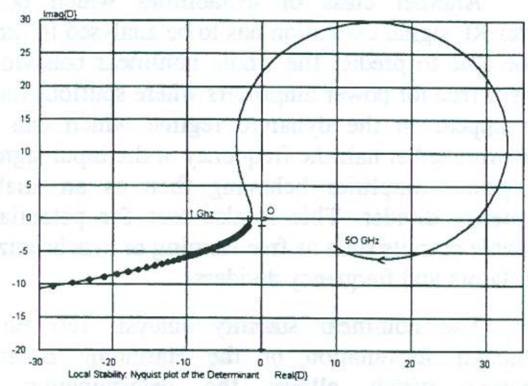


Fig-7 Nyquist plot of the determinant of the characteristic system of the ring oscillator

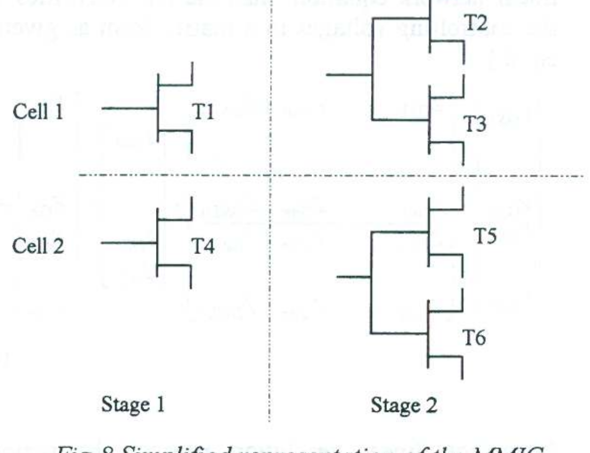


Fig-8 Simplified representation of the MMIC analysed .

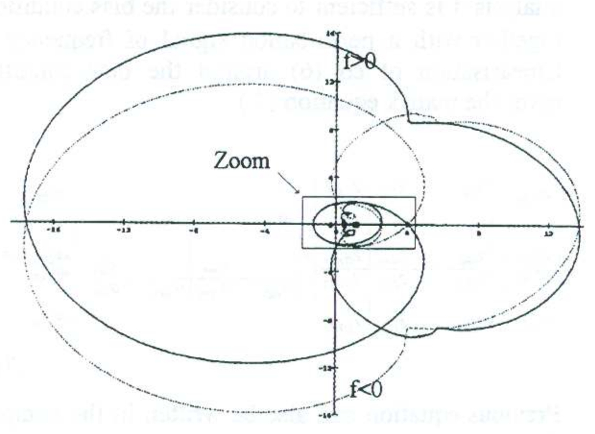


Fig 9-a: Nyquist plot of the NDF for the MMIC of Fig -8 for a frequency range of [-50GHz: 50GHz].

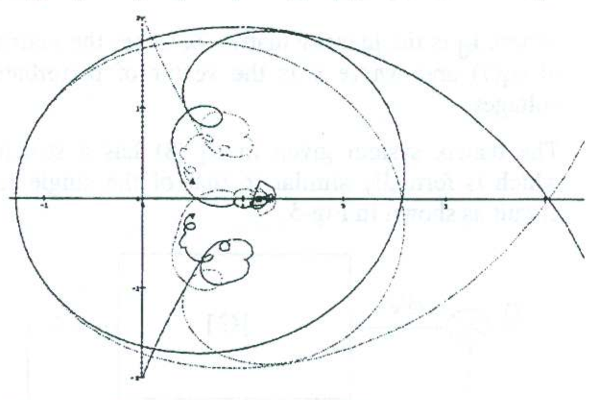


Fig 9-b The same plot as in Fig-9-a but magnified around the origin. An encirclemnt of the origin can be noted indicating an instability of the amplifier for a frequency of approximately 7.0GHz

3-1 Nonlinear stability analysis

The spectral balance equation is a generalization of the matrix equation given in eq (6). Consider the vector of independent controlling variables which can be voltages or currents. The spectral Balance equation can be expressed as:

$$\mathbf{H}(\mathbf{X}) = \mathbf{X}(f) - \mathbf{A_y}(f)\cdot\mathbf{Y}(\mathbf{X}(f)) - \mathbf{E}(f) = \mathbf{0} \quad (9)$$

where $\mathbf{A_y}(f)$ is the matrix linking the commands with the nonlinear sources $\mathbf{Y}(\mathbf{X}(f))$. $\mathbf{E}(f)$ is the vector of Thevenin (Norton) equivalent sources.at the controlling ports. Projecting this equation on the basis of fundamental frequencies involved i.e :

$$\begin{aligned} &e^{j2\pi\cdot f_k t} \quad \text{with} \quad f_k = \mathbf{k}\cdot\mathbf{f}_B^T \\ &\mathbf{f}_B^T = \{f_1, f_2, \cdots\} \end{aligned} \quad (10)$$

gives the nonlinear system to be solved:

$$\begin{aligned} &\mathbf{H}_k(\mathbf{X}) = \mathbf{X}_k - \mathbf{A_y}(f_k)\cdot\mathbf{Y}_k(\mathbf{X}) - \mathbf{E}_k \\ &k \in \{-N, \cdots, 0, \cdots, N\} \end{aligned} \quad (11)$$

By introducing a small perturbation of complex frequency $\omega - j\sigma$, a perturbation system is obtained which reads:

$$\begin{aligned} &(\mathbf{I}_d - \mathbf{A_y}(\omega - j\sigma)\cdot\mathbf{U})\cdot\Delta\mathbf{X} = \mathbf{0} \\ &\mathbf{U} = (U_{k-l})_{-N\le k\le N; -N\le l\le N} \\ &U_{k-l} = \frac{\partial Y_k}{\partial X_l} \end{aligned} \quad (12)$$

The characteristic matrix can then be written as

$$\Gamma(\sigma + j\omega) = (\mathbf{I}_d - \mathbf{A_y}(\omega - j\sigma)\cdot\mathbf{U}) \quad (13)$$

The stability of the circuit is ensured if there is no eigenvalue with positive real part. However computing the eigenvalues of a complex system is a difficult task so that the Nyquist criterion is applied to the determinant of the characteristic matrix, evaluated for real perturbation frequency ω.

$$\Delta(j\omega) = \det[\Gamma(j\omega)] \quad (14)$$

In the periodic regime of period $\frac{2\cdot\pi}{\Omega_0}$, $\Delta(j\omega)$ is almost periodic of period Ω_0. so that the Nyquist locus will be traced between 0 and Ω_0. This provides a criterion for the local stability of the steady state regime.

Generally in nonlinear microwave circuits, some parameters are likely to vary so that a global stability of the circuit can be derived if the bifurcation condition is solved , i.e:

$$\Delta(\sigma + j\omega) = 0 \text{ with } \sigma(\eta_0) = 0 \text{ et } \left.\frac{d\sigma}{d\eta}\right|_{\eta=\eta_0} \neq 0 \quad (15)$$

Depending of the value of the perturbation frequency for which a bifurcation occurs, three cases can be distinguished which are:

- $\omega = 0 + k\cdot\Omega_0$ Direct bifurcation (D type)
- $\omega = \frac{(2\cdot k+1)\cdot\Omega_0}{2}$ Inverse Bifurcation (I type)
- $\omega \neq \frac{p}{q}\cdot\Omega_0$ Hopf Bifurcation (H Type)

3-2 Application to synchronized circuits [8]

For synchronized circuits such as oscillators or frequency dividers. The bifurcation analysis proposed gives a very powerful tool for a deep insight in the behaviour of the circuit. In that case the two parameters of interest are the input power and the input frequency : P_{in} *and* ω_{in}. The bifurcation equation then reads:

$$\begin{cases} \Delta(j\omega, \Omega_{in}, P_{in}) = 0 \\ \quad\text{with} \\ \mathbf{H}_b(\mathbf{X}, \Omega_{in}, P_{in}) = 0 \end{cases} \quad (16)$$

The non linear system (16) always contain a number of unknowns equal to the number of equations plus one. thus system(16) defines a bifurcation locus in the ω_{in} , P_{in} plane. Depending of the values taken by ω along the bifurcation locus several types of locus can be distinguished. Two main types are of interes : synchronous perturbation locus which corresponds to $\omega = k\cdot\frac{\Omega_0}{2}$ and asynchronous bifurcation locus which corresponds to the locus of Hopf type bifurcations.

The previous analysis has been applied to MMIC synchronized oscillators as well as to MMIC frequency dividers [8] with success. It should be noticed that the bifurcation analysis has been extended to the quasi-periodic case [9] thus providing a complete stability portrait of an MMIC frequency divider.

The analysis of the oscillator begins with the plot of the Nyquist locus in order to determine the possible oscillation frequencies . This plot is given in fig-10-a. Then the large signal analysis provides the output frequency and power variations versus the gate voltage as shown in fig-10-b. Finally the synchronized oscillator is analyzed by means of the method previously exposed in order to provide the global stability portrait shown in fig-11.

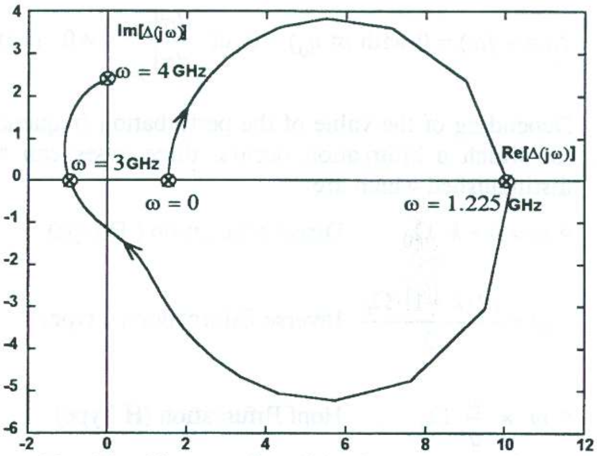


Fig-10-a-Nyquist plot of the determinant of the characteristic system of the 3GHz oscillator indicating an instability with an oscillation frequency of 3Ghz

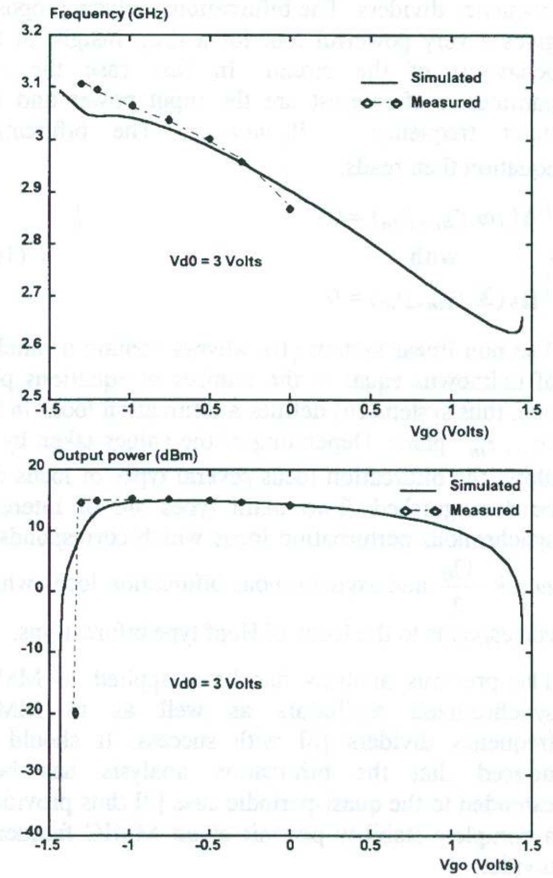


Fig-10-b Measured and computed output frequency and and out put power of the oscillator in the large signal regime.

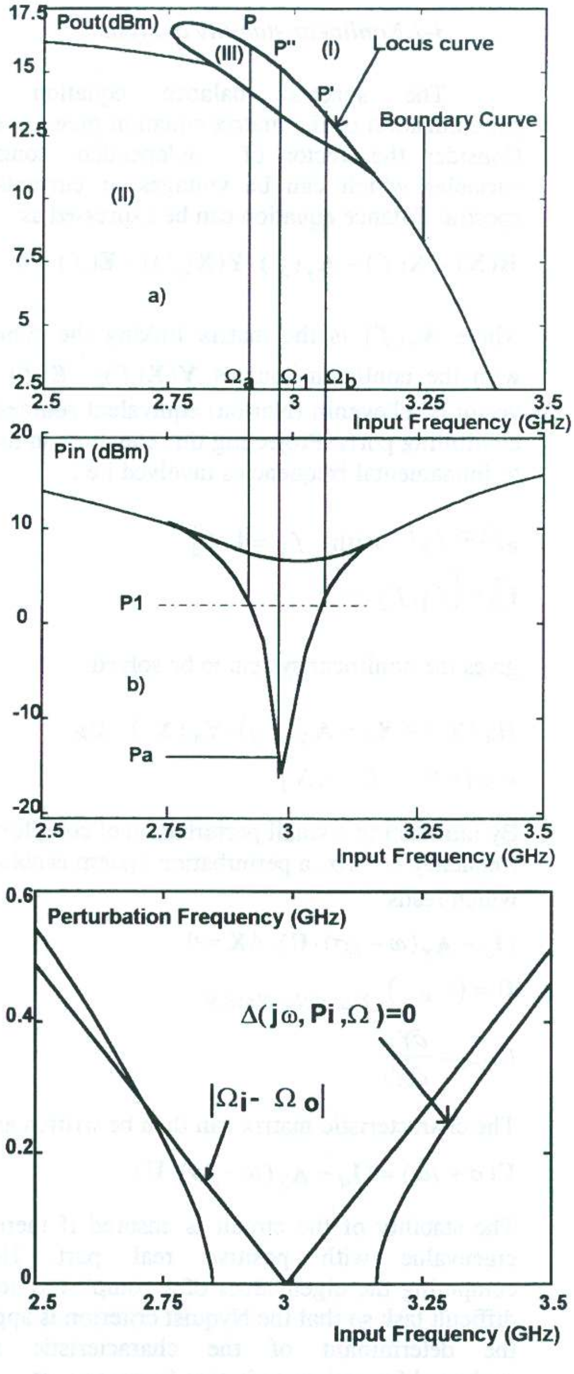


Fig-11 Global stability portrait of the synchronized oscillator. Synchronous bifurcation locus (locus curve and asynchronous bifurcation locus (boundary curve) in the output plane $P_{out}(F_{in})$ (a) and in the input plane $P_{in}(f_{in})$. Region I corresponds to the periodic regime where the oscillator behaves as a negative resistance amplifier. Region II corresponds to the synchronization of the oscillator and region III corresponds to the quasi periodic regime. From these figures the synchronization bandwidth is given by Ω_b-Ω_a for an input power P_{in}=P_1. The perturbation frequency along the boundary curve is given in fig 11-c

3-3 Open loop analysis of frequency dividers

Unfortunately the analysis method described in the previous section is not available on standard CAD platforms. Moreover simulation of synchronized circuits leads to convergence problems due to the extreme sensitivity of the convergence to the phase variation of unknowns. In order to overcome this problem an open loop method can be used.[10]. The method consists of analysing the frequency divider as a cascade of amplifiers. Each cell is constituted by opening the loop in the single transistor following the method shown in fig-12. The frequency divider is then analysed as an amplifier with a bias generator at the frequency f_{in} and an input generator at $f_{in}/2$. The oscillation condition is then given by:

$$V_{GS}(t) = V_{ext}(t) \quad V_{ext}(t) \text{ generator at } f_{in}/2 \qquad (17)$$

Practically the cell indicated in fig-12 is cascaded in order to make the amplitude and phase of the command voltage of the cell k V_{gsk} to converge towards the steady state value.

This technique has been applied to the design of a MMIC frequency divider with an input frequency of 60GHz . In order to calculate the phase of the Gate to Source voltage for each cell, the phase of the synchoization generator was set to 0 and the overall circuit was fed with an input signal at frequency $f_{in}/2$ with a variable phase. The phase of the command voltages are shown in Fig-13 where it can be seen that for the last cell the phase converges towards two values distant of 180°, respectively +100° and -80° This fact is characteristic of frequency dividers by two. It can be noticed that the value corresponding to point B corresponds to an unstable regime where the $F_{in}/2$ solution dissapears. This fact is confirmed by the time domain plots of the gate to source voltages of each cell shown in Fig-14.

The layout of the circuit manufactured by Thomson TCS foundry is shown on Fig-15. This circuit has demonstrated a synchronization bandwidth of 15%. as can be seen from the measured results shown in Fig-16.

4 - Conclusion

The prediction of the stability of MMICs become an important featur of the CAD process. This is specially true for multidevice power amplifiers where the classical analysis using the K factor may fail in cases of internal instabilities. This is also true for the global stability analysis of synchronized devices through the bifurcation analysis. Methods have been given which enable the circuit designer to deal with almost all the stability problems. However there is still a lot of work in order to introduce the formalism presented in commercial CAD packages.

References

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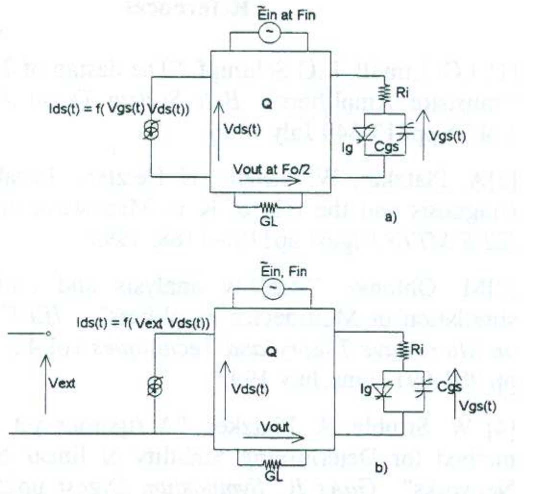

Fig-12 Open loop realization of a FET based circuit.

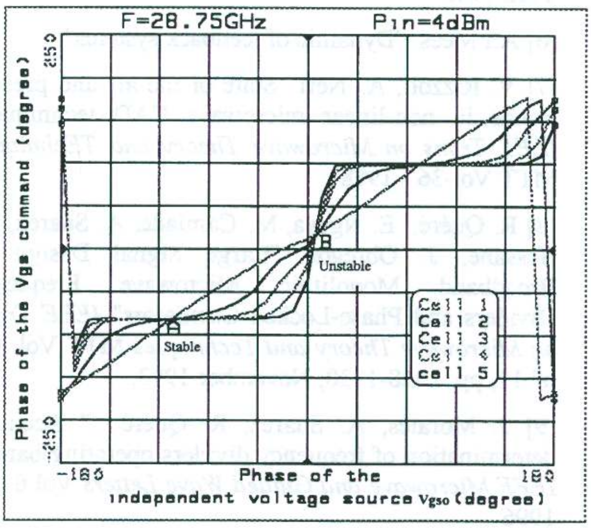

Fig-13- Representation of the phase of the command voltage for each cell considered in the analysis of the MMIC 60GHz frequency divider.

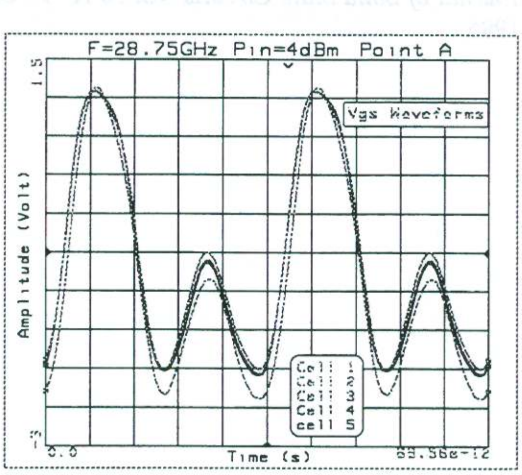

Fig-14-a

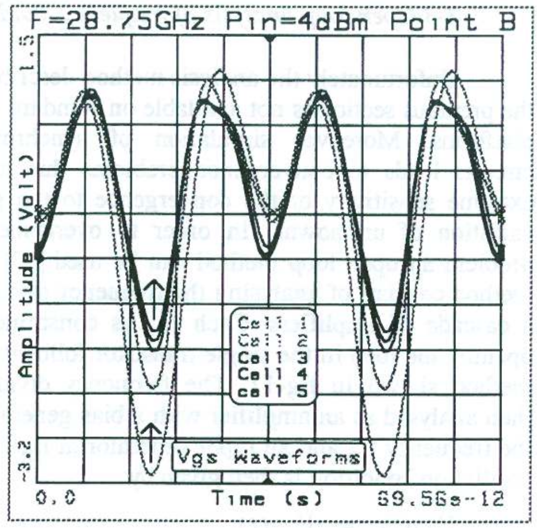

Fig 14-b)

Fig-14 Voltage waveforms of command voltages of each cell in the stable case a) and in the unstable case b)

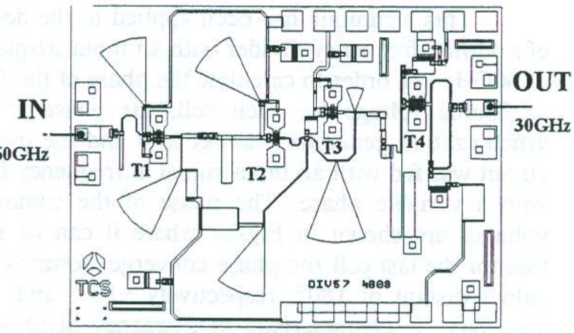

Fig-15 Layout of the frequency divider

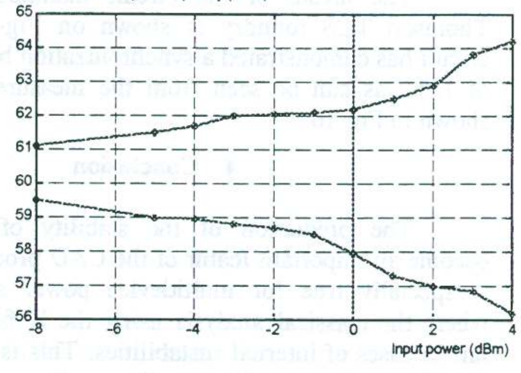

Fig-16 Measurement results in test fixture