A design method for DR-stabilized MESFET oscillators

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Abstract. In this paper a method is proposed for the design of DR-stabilized MESFET oscillators where several different design objectives, associated both to stability and output power, can simultaneously be taken into account. First, stability and self-starting capability of an oscillator are related to suitable performance indexes which are defined in the paper. Then an efficient procedure aimed at searching for an optimal compromise between opposing design requirements is described. The validity of the whole procedure and the performance indexes introduced has been investigated by considering the design of a microstrip parallel-feedback DR oscillator using a medium power GaAs MESFET.

1 INTRODUCTION

Dielectric Resonators (DR) are widely used as frequency stabilizing elements in oscillators and, in particular, in MIC's using GaAs MESFET transistors. The design of these circuits is quite a complex task not only for the intrinsic non-linearity of operation, but also because multiple design objectives must be simoultaneously satisfied. In fact, in addition to good frequency stability, also relatively large output power, reliable self-starting capability and convergence to stable steady-state oscillation must be guaranteed.

In order to search for a good compromise between these often conflicting performance requirements a suitable design procedure is needed; in particular, special care must be taken in the identification of suitable operating conditions both for the DR and the MESFET, with the aim of maximizing the frequency stabilization without sacrificing too much output power.

To this aim a suitable criterion for the evaluation and optimization of the frequency stabilizing effect introduced by the DR has been adopted in the design procedure described in this paper. In particular the design is carried out by means of a simplified yet sufficiently accurate procedure for the search for the optimal oscillator operating conditions. The design approach has been tested by considering different possible design examples; the results of accurate numerical simulations, including the analysis of sensitivity to circuit perturbations, have confirmed the validity of the design method proposed.

2 CIRCUIT STRUCTURE and STABILITY ANALYSIS of DRO's

Dielectric Resonators Oscillators are at present implemented according to different possible circuit configurations characterized by different topologies [1]; two of these, which are quite commonly used, are shown in fig.1 and can be classified as series-feedback (fig.1a) and parallel-feedback structures (fig.1b). In particular the first one is more commonly used with non-packaged MESFET transistors,

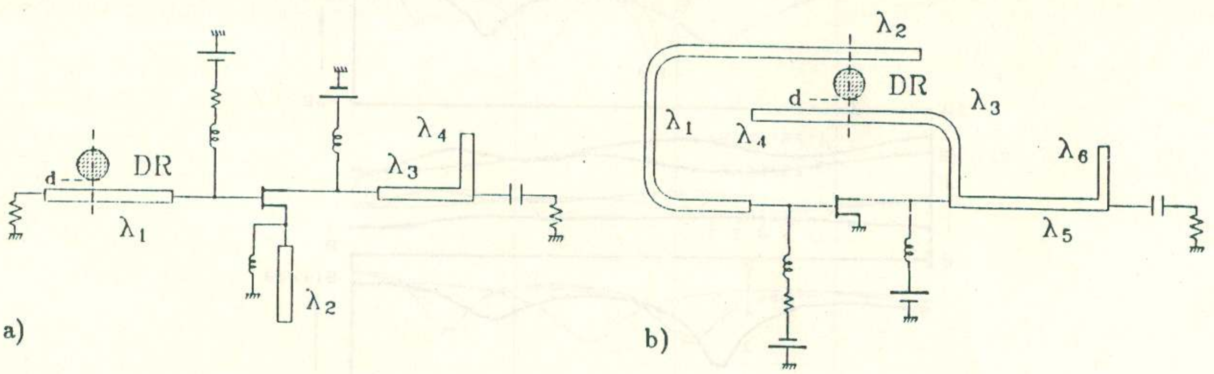

Fig.1: Microstrip implementations of two different DRO circuit configurations.

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while the latter one is more often used with packaged devices.

Since the electrical response of a Dielectric Resonator can be approximated by an equivalent lumped LC resonating circuit [1], any DR-stabilized oscillator (and in particular the DRO structures which are shown in fig.1), can be described by the schematic circuit diagram drawn in fig.2, where the ideal transformer represents the electromagnetic coupling of the DR with the microstrip lines [1]. In the same figure the four-port N describes the linear feedback-matching network which is assumed to be reciprocal and practically lossless, while G_L is the load conductance.

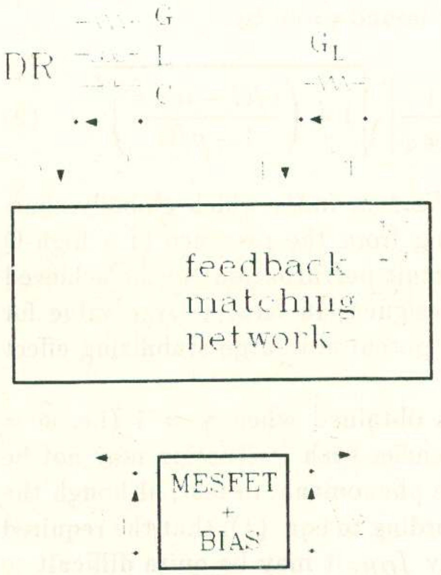

Fig.2: Schematic equivalent circuit of a DRO.

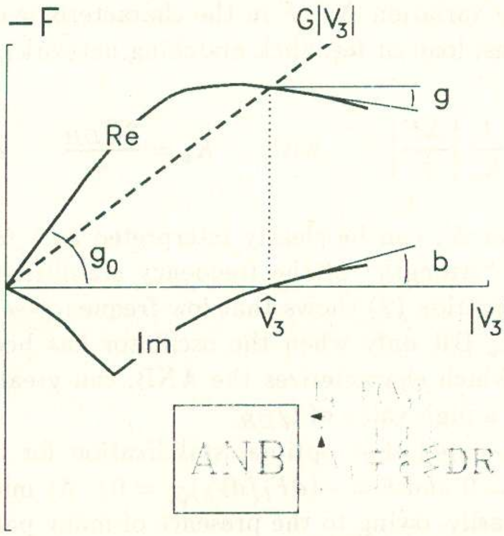

Fig.3: Plot of the real and imaginary parts of the characteristic $I_3 = F[|V_3|]$ of the Active Non-linear Bipole. The intersection point $\widehat{V_3}$ with the straight line associated to the DR admittance at the resonance frequency identifies the amplitude of oscillation.

The main stability properties of the circuit in fig.2 can be conveniently studied at the connection port (see figs.2 and 3) between the equivalent resonant circuit and the Active Non-Linear Bipole (ANB) composed by all the remaining circuit elements (i.e. N, G_L, the MESFET and its biasing network). The stability analysis becomes quite simple if (with no relevant loss of accuracy owing to the necessarily high quality factor of the DR) the voltage V_3 is assumed to be practically sinusoidal. In such conditions the amplitude $|\widehat{V_3}|$ and the frequency $\widehat{f}$ of oscillation can be computed by solving the non-linear phasor equation:

$$G\left[1 + jQ_{DR}\left(\frac{f}{f_{DR}} - \frac{f_{DR}}{f}\right)\right]|V_3| + F[|V_3|] = 0 \tag{1}$$

where Q_{DR} and f_{DR} are, respectively, the unloaded quality factor and resonance frequency of the DR while the equivalent conductance G accounts for the losses in the DR; $I_3 = F[|V_3|] = F_R[|V_3|] + jF_I[|V_3|] = |F|e^{j\phi}$ is a non-linear complex function (see fig.3) which defines the phasor I_3 of the first order harmonic[1] of the current $i_3(t)$ in the ANB as a function of the amplitude $|V_3|$ (zero phase can be assumed) of the almost sinusoidal voltage $v_3(t)$.

The stability properties of the DRO can be studied [3] in terms of the differential parameters $g \doteq -(dF_R/dV_3)_{\widehat{V_3}}$, $b \doteq -(dF_I/dV_3)_{\widehat{V_3}}$ and $g_0 \doteq -(dF_R/dV_3)_{V_3=0}$ which are obtained, respectively, by linearizing the non-linear characteristic $F[|V_3|]$ at the steady-state solution $|\widehat{V_3}|$ and at the initial solution $V_3 = 0$ (see fig.3). In fact, according to the Mikailov's criterion [2] and in the hypothesis of a

[1] It should be noted that considering only the first order harmonic of the current $i_3(t)$ does not involve relevant loss of accuracy since the highly frequency-selective impedance of the DR is such as to guarantee that the voltage $v_3(t)$ is almost independent from the higher-order harmonics of $i_3(t)$. Moreover, for the same reason, also the frequency-dependence of the non-linear characteristic F in the neighbourhood of the resonance frequency f_{DR} is practically negligible.

relatively large quality factor Q_{DR}, it can be shown that the self-starting capability of the oscillator is guaranteed when the *small-signal instability factor* $K_i = g_0/G$ is greater than one. Analougosly, the steady-state solution is stable provided that the large-signal *amplitude stability factor* $K_a = g/G$ is smaller than one.

Moreover, in order to achieve good frequency stability, the frequency of oscillation should be almost independent of the characteristic of ANB which includes the MESFET and, consequently, is quite strongly dependent on temperature, bias voltage, ageing phenomena etc.. By means of a linear perturbation analysis it can be shown [3] that the relative variation $\Delta f/\widehat{f}$ of the oscillation frequency, caused by any variation $\Delta F/\widehat{F}$ in the characteristic of ANB (due to any possible perturbation in the MESFET, bias, load or feedabck matching network), has an upper bound given by:

$$\left|\frac{\Delta f}{\widehat{f}}\right| \leq \frac{1}{K_s}\left|\frac{\Delta F}{\widehat{F}}\right| \qquad \text{with} \qquad K_s = \frac{2Q_{DR}}{\gamma} \qquad \text{and} \qquad \gamma = \left|\frac{1}{\cos\widehat{\phi}}\right|\sqrt{1+\left(\frac{b/G-\tan\widehat{\phi}}{1-g/G}\right)^2} \tag{2}$$

The parameter K_s can be clearly interpreted as a *frequency stabilization index* which globally characterizes the "strength" of the frequency stabilizing effect deriving from the presence of a high-Q resonator. Equation (2) shows that low frequency-sensitivity to circuit perturbation can be achieved with a high-Q DR only when the oscillator has been carefully designed; in fact, a large value for the term γ, which characterizes the ANB, can greatly reduce the potentially large stabilizing effect associated to a high value of Q_{DR}.

It can be noted that optimal stabilization for a given DR is obtained when $\gamma = 1$ (i.e. $\widehat{\phi} = \arg\{F[|\widehat{V_3}|]\} = 0$ and $b = -(dF_I/dV_3)_{\widehat{V_3}} = 0$). At microwave frequencies such a situation may not be obtained so easily, owing to the presence of many parasitic reactive phenomena; in fact, although the condition $\widehat{\phi} = 0$ (i.e. $F_I[|\widehat{V_3}|] = 0$) can be satisfied by imposing, according to eqn. (1), that the required oscillating frequency $\widehat{f}$ coincides[2] with the DR resonance frequency f_{DR}, it may be quite difficult to have $b = 0$ at the same time without sacrificing other performance requirements. In this respect, the problem of maximizing the frequency stabilization index K_s should be carefully taken into account in DRO design.

According to the above considerations DRO performance is characterized by the stability parameters K_i, K_a, K_s, in addition to the output power $P_{OUT} = |V_4|^2 G_L/2$ which is delivered to the load. A suitable design procedure is needed in order to compute the values of the circuit parameters (e.g. the dimensions of the microstrip lines in the circuit structures in fig.1) which provide optimal performance trade-off.

3 DRO DESIGN

The proposed DRO design procedure is based on the hypothesis of moderately non-linear almost sinusoidal operating conditions for the MESFET; in this condition only the fundamental harmonic components of the electrical variables can be taken into account. The substantial validity of this approximation is confirmed not only by other authors [4..7] but also by the results of the validation procedure described in sect. 5.

In such conditions the design of a DR-stabilized oscillator can be carried out by determining the "optimal" values of the elements of the admittance matrix $Y = [jB_{mn}]$ which describes the response of the reciprocal lossless four-port N (see fig.2) at the given frequency of oscillation. In fact, once the values of the designable parameters B_{mn} have been suitably chosen, the dimensions of the microstrip lines which implement the required structure (see fig.1) can easily be computed by following any linear synthesis procedure.

The element B_{mn} must be determined with the aim of optimizing the stability parameters K_i, K_a, K_s compatibly with the given output power requirements. This task can be quite complex since the computation of the "optimal" values of B_{mn} involves solving the non-linear equations which describe the large-signal operation of the oscillator. The solution might be obtained by means of very "expensive" non-linear analysis and numerical optimization tecniques; however, a more simpler and

[2]Obviously, in practice, a tuning device for the DR may be needed.

quite accurate design procedure can be used. In fact, if a suitable criterion is adopted for the choice of the voltage phasors V_1 and V_2 (i.e. the MESFET large-signal operating condition), and by applying the substitution theorem at port 1 and 2 in fig.2, the non-linear problem of determining B_{mn} can be trasformed into a simpler one which does not involve any longer the non-linear MESFET equations.

The MESFET large-signal operating condition V_1, V_2 can be suitably chosen with the aim of maximizing the power $P_T = Re\{V_2 I_2^* + V_1 I_1^*\}/2$ delivered by the transistor (which almost coincides with P_{OUT} if the losses in the DR are relatively small) and introducing sufficient gain compression, which is needed in order to stabilize the steady-state oscillation. The search for the optimal values of V_1 and V_2 can be carried out in different ways, either experimentally (by using a "load-pull" large-signal measurement set-up [6,7]) or analytically or, possibly, through numerical optimization when using harmonic-balance analysis and conventional large-signal transistor models [11]. As an alternative, the above search can be quite easily carried out by means of a previously proposed "black-box" MESFET modelling approach [8].

Once the transistor large-signal operating condition has been suitably chosen, by taking into account the MESFET model equations the four electrical variables V_1, V_2, I_1, I_2, are completely determined; this implies that 4 of the n designable parameters[3] B_{mn} associated to the matrix Y must be such as to satisfy the current balances at the ports 1 and 2 of the network N. Thus, the remaining $n-4$ "free" designable parameters can be suitably chosen with the aim of optimizing the performance parameters K_i, K_a, K_s, while keeping the losses in the DR at a reasonably low level. Then, the knowledge of the Y matrix of N enables to determine, by using any linear synthesis procedure, the dimensions of the microstrip lines which implement the linear feedback-matching network (see fig.1).

4 AN EXAMPLE of APPLICATION

The proposed DRO design procedure has been applied [9] for the design of a parallel-feedback DRO of the type shown in fig.1b. The structure considered is characterized by a relatively large number of design parameters B_{mn} which enable a number of different design objectives to be simultaneously satisfied. In particular, it can be shown that the purely imaginary elements of the Y matrix of the network N which are actually avalaible as designable parameters are: jB_{11}, jB_{22}, jB_{33}, jB_{44}, $jB_{13} = jB_{31}$ and $jB_{23} = jB_{32}$. In fact the topological properties of the structure shown in fig.1b impose the constraints: $jB_{12} = jB_{21} = jB_{14} = jB_{41} = jB_{34} = jB_{43} = 0$ and $jB_{24} = jB_{42} = -jB_{44}$.

The search for the optimal values of V_1, V_2 which, according to the considerations made in section 3, maximize the power P_T delivered by the MESFET compatibly with a sufficient gain compression, has been carried out by means of a simplified "black-box" MESFET modelling approach, which has been previously proposed and validated [8..10] by the authors. According to this approach the large-signal transistor response can be approximated, in terms of wave variables a and b, by the simplified phasor equations [8]:

$$b_1 = S_{11}a_1 + S_{12}a_2 + \mathcal{F}_{11}(|a_1|)a_1 \quad ; \qquad b_2 = S_{21}a_1 + S_{22}a_2 + \mathcal{F}_{21}(|a_1|)a_1 \tag{3}$$

where S_{ij} are the small-signal S-parameters, $\mathcal{F}_{11}$ and $\mathcal{F}_{21}$ non-linear complex functions which account for the deviations from linear response [4]; the mathematical model (3) can easily be characterized through conventional small- and large-signal S-parameter measurements [8,10].

The search for the optimal values of $|a_1|$, a_2, or equivalently $|V_1|$, V_2 [5], can easily be carried out by means of the above model. In fact eqns. (3) allow for the straightforward computation, for any value of $|a_1|$, of the value of a_2 which maximizes the power P_T. Thus the search for the optimum value of $|a_1|$ (which provides a good compromise between the requirements on output power and gain

[3] Taking into account that the feedback-matching network N is lossless and reciprocal, the maximum possible number n of independent designable parameters is equal to 10; however, the microstrip topologies which are practically used, intrinsically introduce additional constraints on the possible values of the elements of Y so that the actual number of available design parameters is often smaller than 10.

[4] Equations (3) are valid under the constraint that the required gain compression is introduced only by the non-linear effects (i.e. $\mathcal{F}_{11}$, $\mathcal{F}_{21}$) controlled by the amplitude of the input incident wave a_1; this choice has the advantage of reducing output harmonic distortion and simplifying the design procedure [9].

[5] Since the circuit is autonomous the phase of V_1 can be arbitrarily assumed to be zero.

compression) can be directly carried out on computer-generated plots of the type shown in fig.4, where P_T and the transmission gain parameter $|S_{21} + \mathcal{F}_{21}|$ are given as functions of $|a_1|$.

The designable parameters B_{mn} of the network N must be such as to satisfy the current-balance equations at ports[6] 1, 2 and 3:

$$I_1 = j[B_{11}V_1 + B_{13}V_3] \quad , \qquad I_2 = [jB_{22} + B_{24}^2/(G_L - jB_{24})]V_2 + jB_{23}V_3 \qquad (4)$$
$$-GV_3 = j[B_{13}V_1 + B_{23}V_2 + B_{33}V_3]$$

These have been obtained by imposing, according to the considerations made in section 2, that the oscillation frequency $\hat{f}$ coincides with the resonance frequency f_{DR} of the DR.

The currents I_1, I_2 in system (4) are uniquely defined through the non-linear equations of the MESFET model since its large-signal operating condition (V_1, V_2) has been chosen. Thus, all the designable parameters B_{mn} can easily be computed[7], by solving eqns. (4), as functions of the still unknown voltage V_3. Moreover, the differential parameters g, b, g_0 and, consequently, also K_i, K_a, K_s can be easily determined through a simple linear analysis in terms of the corresponding differential parameters of the transistor model. In such conditions the performance parameters can be directly computed in terms of V_3 and, thus, DRO design basically consists of choosing a suitable value for the amplitude and phase of this voltage. In particular, the stability parameters can be plotted (see fig.5) as functions of $\angle V_3$ and its amplitude $|V_3|$ (or, equivalently, $\sqrt{P_{DR}/P_T}$). These plots provide a design map where a good compromise between the often conflicting requirements of good frequency stability (i.e. large K_s), reliable self-starting capability (i.e. K_i sufficiently greater than one) and low losses in the DR can easily be sought for.

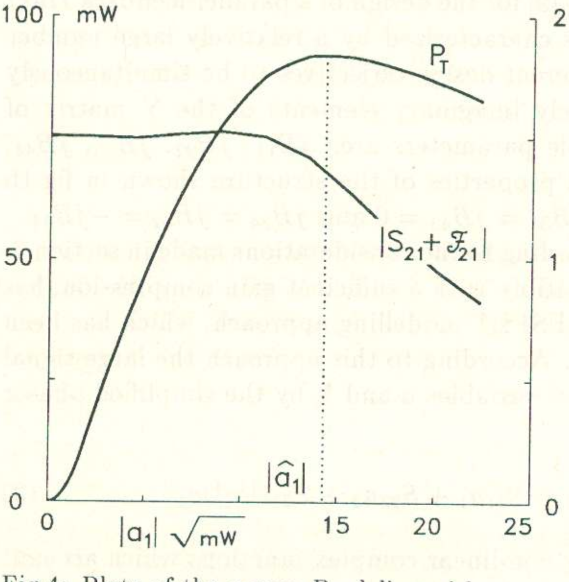

Fig.4: Plots of the power P_T delivered by the MESFET and the transmission gain parameter $|S_{21} + \mathcal{F}_{21}|$ vs $|a_1|$.

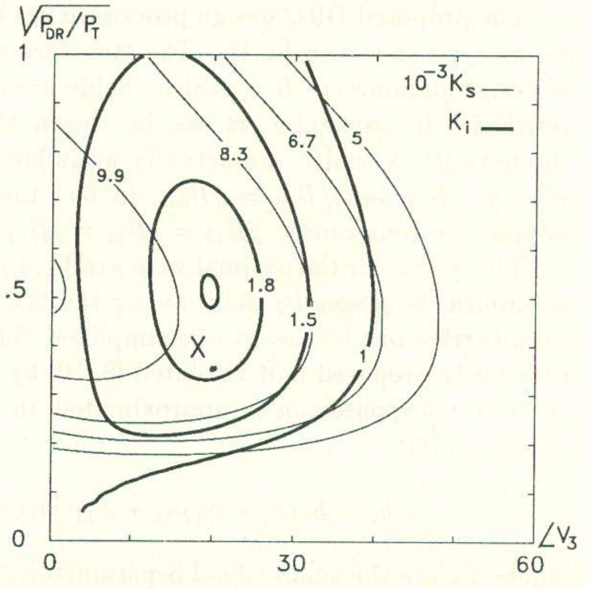

Fig.5: Design map for the choice of optimal DRO performance trade-off. Point X marks the parameter values chosen.

When a suitable value for V_3 has been chosen, the parameters B_{mn} are uniquely determined by solving eqns. (4). Then, the design of the microstrip structure which implements the equivalent circuit in fig.2 becomes an easy task since simple linear synthesis techniques can be used. For instance, the electrical lengths λ_i of the microstrip structure in fig.1b can be directly determined, for a given distance

[6] It is not necessary to consider explicitly the electrical variables at port 4. In fact the power delivered to the load is the only interesting parameter at port 4 and is uniquely determined, through the power balance at the other three ports, since the feedback-matching network N is lossless.

[7] System (4) is a second order one; its solution, however, is extremely simple. The choice of one between two possible determinations of B_{24} is not relevant for the oscillator performance.

d associated to the DR coupling[8], by imposing its congruence with the already determined Y matrix.

5 VALIDATION OF THE DESIGN PROCEDURE

In the above-described DRO design method only the fundamental harmonic components of electrical variables have been considered; moreover, the large-signal transistor response has been approximated by a simplified "black-box" mathematical model. In order to evaluate the loss of accuracy arising from such approximations (in the absence of the other difficult-to-estimate errors which are always present in an experimental prototype), a suitable numerical validation procedure has been used. To this aim the quite accurate and widely-used Curtice model [11] has been adopted for the large-signal modelling of a medium power GaAs MESFET; the large-signal performance predictions were carried out by using both the non-linear simulation program SPICE and the Harmonic Balance technique [12..15].

The HB tool has been used in the simulation of the large-signal S-parameter measurements necessary for the characterization of the black-box transistor model at the required oscillating frequency of $3GHz$. On this basis, the "optimal" large-signal operating condition of the MESFET was chosen on computer-generated plots (see point $|\widehat{a_1}|$ in fig.4, corresponding to a power delivered by the transistor $P_T = 87.7mW$). The associated design map, which is shown in fig.5, suggested the choice of $\angle V_3 = 20°$ and $P_{DR}/P_T = 0.12$ (point X in the map) which provides good stability of the steady-state oscillation ($K_a = -0.19, K_s = 9.5 \cdot 10^3$ with $Q_{DR} = 5000$), a sufficiently large initial instability coefficient ($K_i > 1.8$) and a relatively low DR loss ($P_{DR} = 10.5mW$). Then the parameters B_{mn} were computed and the corresponding lengths λ_i of the microstrip lines (see fig.1b) determined according to the procedure described in section 4.

The SPICE time-domain simulation of the DRO circuit showed spontaneous starting of oscillation and natural convergence to stable almost-sinusoidal periodic steady-state. The circuit response computed by HB was found to be in good agreement (see Tab.1) with the performance prediction provided by the design procedure. The low harmonic distortion of the v_1, v_2, v_3 voltage waveforms justifies the quite good accuracy of the design procedure based only on fundamental harmonic components.

Table 1: Comparison of voltages in DRO predicted by the design procedure and simulated by an highly accurate HB numerical simulation.

	Design procedure		HB simulation		
	$\|V\|$	$\angle V$	$\|V\|$	$\angle V$	Dist.%
V_1	3.13	0.0°	3.07	0.0°	5.8
V_2	2.51	140.2°	2.36	138.7°	4.4
V_3	6.55	20.0°	6.18	19.5°	.05
P_{OUT}	$77mW$		$68mW$		

In order to verify the validity of the frequency stabilization index K_s, also other different possible designs of the same circuit structure (i.e. the same topology, MESFET and resonator, with different values of the stub lengths λ) were considered. All these designs are characterized (see columns 2 to 5 in Tab.2) by very similar values for the basic performance parameters P_{OUT}, K_i, K_a but quite different values of K_s. An accurate circuit analysis was then carried out by means of the HB technique; this provided the normalized parametric frequency sensitivities $s_p \doteq \frac{\Delta f}{f}/\frac{\Delta p}{p}$ with respect to a number of circuit parameters p (i.e. transmission line lengths λ, load reflection coefficient ρ_L, characteristic parameters g_m and C_{GS} of the MESFET model). The results summarized in Tab.2 show a relevant correlation between decreasing values of K_s and generally increasing values of s_p. These considerations show that the proposed frequency stabilization index K_s provides relevant information on the level of frequency stabilization introduced by a high-Q DR in an oscillator.

In conclusion, the results of accurate numerical simulations and circuit sensitivity analyses have confirmed the validity of the proposed DRO design method. The main advantages of the design procedure described are related to the possibility of providing near-optimal performance trade-off

[8] It is not strictly necessary to directly use the distance d, which characterizes the DR coupling, as a design parameter in the synthesis of the given matrix Y; in fact, even when the distance d is assigned "a priori" (by choosing, for instance, a convenient value for which an experimental characterization is avalaible) the electrical lengths λ_1 to λ_6 are a set of design parameters which is sufficient in order to satisfy the given constraints on Y.

between several different design objectives (i.e. frequency stability, output power etc...) without using quite-complex non-linear optimization techniques, which may involve convergence problems.

A frequency sensitivity index K_s has been also defined which can be efficiently used when optimization of frequency stability is one of the main objectives in DRO design.

Design #		1	2	3	4	5
Basic performance parameters	P_{out}	77.2 mW	79.8 mW	84.3 mW	83.4 mW	85.2 mW
	K_i	1.8	1.8	1.5	1.5	1.4
	K_a	$-1.9\cdot10^{-1}$	$-1.5\cdot10^{-1}$	$-8.6\cdot10^{-2}$	$-1.1\cdot10^{-1}$	$-4.7\cdot10^{-2}$
Frequency stabilization index	K_s	$9.5\cdot10^{3}$	$2.9\cdot10^{3}$	$1.5\cdot10^{3}$	$8.3\cdot10^{2}$	$5.3\cdot10^{2}$
Parametric frequency sensitivities	s_{λ_1}	$6.8\cdot10^{-2}$	$2.7\cdot10^{-1}$	$5.4\cdot10^{-1}$	$6.6\cdot10^{-1}$	$7.8\cdot10^{-1}$
	s_{λ_2}	$2.0\cdot10^{-2}$	$1.3\cdot10^{-1}$	$8.0\cdot10^{-2}$	$2.2\cdot10^{-1}$	$1.0\cdot10^{-1}$
	s_{λ_3}	$2.1\cdot10^{-2}$	$2.5\cdot10^{-1}$	$1.7\cdot10^{-1}$	$1.6\cdot10^{-1}$	2.0
	s_{λ_4}	0.0	$6.6\cdot10^{-3}$	$1.4\cdot10^{-2}$	3.7	8.9
	s_{λ_5}	$2.5\cdot10^{-4}$	$1.2\cdot10^{-3}$	$1.3\cdot10^{-3}$	$7.1\cdot10^{-2}$	$2.5\cdot10^{-1}$
	s_{λ_6}	$1.5\cdot10^{-2}$	$6.4\cdot10^{-2}$	$3.6\cdot10^{-1}$	$9.1\cdot10^{-2}$	2.3
	$s_{Re(\rho_L)}$	$5.2\cdot10^{-4}$	$8.3\cdot10^{-3}$	$1.8\cdot10^{-2}$	$1.7\cdot10^{-2}$	$2.2\cdot10^{-2}$
	$s_{Im(\rho_L)}$	$1.3\cdot10^{-3}$	$3.5\cdot10^{-4}$	$6.4\cdot10^{-3}$	$2.9\cdot10^{-2}$	$3.3\cdot10^{-2}$
	$s_{C_{GS}}$	$4.1\cdot10^{-3}$	$6.9\cdot10^{-3}$	$1.2\cdot10^{-2}$	$3.2\cdot10^{-3}$	$1.0\cdot10^{-3}$
	s_{g_m}	$2.0\cdot10^{-2}$	$4.6\cdot10^{-2}$	$2.1\cdot10^{-1}$	$4.0\cdot10^{-1}$	$4.7\cdot10^{-1}$

Table 2: Basic performance and parametric sensitivity analysis for different designs of a parallel-feedback DRO; the same circuit topology and dielectric resonator (Q=5000) were used in all cases.

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