Coupled Electromagnetic/Nonlinear Optimization of Self-Oscillating Microstrip Antennas with Far-Field Performance Specifications

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Abstract — **The broadband design of self-oscillating microstrip antennas by direct numerical optimization based on electromagnetic (EM) simulation coupled with harmonic-balance (HB) analysis is demonstrated for the first time. The design goals are formulated in terms of far-field performance such as radiation intensity and cross-polarization suppression. The optimization problem is transformed into a system-solving problem, and the solution is found by a Newton iteration globalized by a trust-region method. This results in an order-of-magnitude reduction in the number of expensive EM analyses that are required to achieve convergence.**

I. Introduction

Self-oscillating antennas are active subsystems including in a single circuit an oscillator and a radiating element, usually integrated on a same substrate. These devices have interesting applications in modern telecommunications systems, since they allow phased arrays to be realized with no need for phase shifters nor for corporate feed networks. In addition, these arrays can be electronically steered by applying DC control signals to a few boundary elements only [1] - [3]. In a self-oscillating antenna, the oscillator circuit and the radiating element strongly interact, and in some cases the antenna also acts as the resonant cavity whereby the frequency of oscillation is determined [4]. It is thus obvious that an accurate design of these subsystems requires the simultaneous use of rigorous nonlinear circuit techniques and of electromagnetic (EM) analysis methods for the passive subnetwork description. Indeed, this is the only way of fully accounting for the antenna influence on the oscillator performance (including near-field couplings), and for the effects of the oscillator circuit layout on the antenna radiation pattern. To the authors' knowledge, a systematic approach to this complex design problem taking full advantage of modern CAD techniques, has never been reported in the literature. In this paper we try to bridge this gap by a novel design approach that effectively combines the harmonic-balance technique for autonomous nonlinear circuits [5] with the method of moments to allow a direct numerical optimization of integrated self-oscillating antennas, both free-running and injection locked. An interesting aspect of this method is that the design goals may include direct specifications on the far-field antenna pattern. As an example, the optimization can determine the oscillator layout and the connection to the antenna in such a way as to produce a prescribed radiation intensity and cross-polarization suppression in the broadside direction, and the like. The basic concept underlying the proposed design method consists in the formulation of the design problem as the solution of a nonlinear system encompassing both functional and harmonic-balance specifications. The system is solved by a norm-reducing Newton iteration scheme whereby each Newton update computation is followed by a multidimensional minimization step based on the trust region concept [6]. This approach will be conventionally referred to as direct-Newton optimization (DNO). With DNO the number of expensive EM analyses that are required to achieve convergence is drastically reduced with respect to a conventional optimization, which normally brings the design effort well within the reach of ordinary workstations or even modern PCs. In addition, degenerate (nonoscillatory) solutions of the nonlinear HB system are automatically suppressed in this way, thus removing an important shortcoming of HB techniques for autonomous circuits [5]. Note that DNO is inherently a broadband design method, as we discuss in section II. While it is true that self-oscillating antennas are often intended for single-frequency use, the device performance across a finite band in the neighborhood of the nominal frequency should always be kept under control in order to guarantee a successful outcome of the design process. Indeed, this allows the operating point to be kept away from critical conditions (such as bifurcations) that could cripple the system performance especially under modulated RF drive.

II. Direct-Newton Optimization

A representative layout of a self-oscillating antenna is given in fig. 1. The nonlinear device is described by the advanced model discussed in [7], which is extracted from measured data. The model includes the parasitic effects introduced by the finger matalizations, while the gate, source, and drain pads are incorporated in the passive circuit layout. This procedure avoids the need for including the fine details of the FET electrode layout in the EM analysis of the passive subnetwork, which allows the grid size to be substantially increased with no appreciable loss of accuracy. At the same time, the main coupling effects between the passive circuit and the device electrodes are taken into account in this way. The passive subnetwork is described as a two-port including the antenna and the oscillator circuit layout. It is analyzed as a whole by an EM simulator based on the method of moments, which directly produces its scattering or admittance parameters. Note that in many cases (as in fig. 1) the active antenna works as an injection-locked amplifier, so that the passive subnetwork has a third port for injection of the locking signal. However, this port is short-circuited during the design process, so that only the two-port case will be considered in the following. Let us assume that the circuit operates in a periodic steady-state regime of fundamental frequency $\omega_0^{(s)}$. The number of positive harmonics to be taken into account in the HB analysis will be denoted by n_H. Let $\mathbf{P}$ be an n_P-vector of designable electrical and layout parameters, and $\mathbf{Y}^{(s)}(\mathbf{P})$ a real vector containing the real and imaginary parts of the linear subnetwork admittance parameters at all the harmonics of interest. The HB solving system for the nonlinear circuit may be written in the form

$$\mathbf{E}^{(s)}\left[\mathbf{X}^{(s)};\ \mathbf{Y}^{(s)}(\mathbf{P})\right] = \mathbf{0} \qquad (1)$$

where $\mathbf{E}^{(s)}$ is the set of real and imaginary parts of the HB errors, and $\mathbf{X}^{(s)}$ is a modified state vector containing the real and imaginary parts of all the state-variable harmonics except for one harmonic chosen as reference, the magnitude of the reference harmonic, and a real tuning parameter whose task is to tune the free oscillation to the specified fundamental frequency of $\omega_0^{(s)}$ for any given choice of $\mathbf{P}$. The dimension of both $\mathbf{E}^{(s)}$ and $\mathbf{X}^{(s)}$ is $n_T = 2(2n_H + 1)$. The extension to a circuit containing a larger number of nonlinear devices would be straightforward. For a given spectrum, the design goals can be formulated by imposing that a number of real performance indexes or network functions (see section III) be greater than or equal to a set of specified values (design goals). This set will be denoted by $\mathbf{G}_{min}^{(s)}$, and the number of specifications (i.e., the dimension of $\mathbf{G}_{min}^{(s)}$) by n_G. Since each performance index is obviously a function of the circuit state and of the passive subnetwork admittance parameters, the design specifications may be synthetically formulated as

$$\mathbf{F}^{(s)}\left[\mathbf{X}^{(s)};\ \mathbf{Y}^{(s)}(\mathbf{P})\right] \geq \mathbf{G}_{min}^{(s)} \qquad (2)$$

where $\mathbf{F}^{(s)}$ in an n_G-vector of network functions. For a broadband optimization the electrical regime has to meet the design goals at a number (say S) of spectra (*design spectra*), that differ from one another exclusively in the value of the fundamental frequency $\omega_0^{(s)}$. We may thus state the broadband optimization problem in the following way: find $\mathbf{P}$ and $\mathbf{X}^{(s)}$ in such a way that both (1) and (2) be satisfied for $1 \leq s \leq S$. The key idea underlying the DNO technique is to formulate the design task as a system-solving problem rather than an optimization. The number of unknowns is $n_U = Sn_T + n_P$, while the number of constraints (including (1) and (2)) that define the problem is $n_C = S(n_T + n_G)$. A broadband design problem is normally overconstrained, in the sense that $Sn_G > n_P$, or $n_C > n_U$. We thus need to change the formulation of the constraints in such a way as to convert the given problem into an equivalent one for which $n_C = n_U$. One possible way of achieving this result is to suitably subdivide the Sn_G specifications (2) into n_P groups, and to combine the constraints belonging to each group into a single inequality to be formulated in terms of a suitable least-qth function, as detailed below. To do so, we first introduce the vectors $\mathbf{X}$, $\mathbf{Y}$ of the real and imaginary parts of the SV harmonics and of the admittance parameters, respectively, at all design spectra. Assuming then that the i-th specification of the n-th group ($1 \leq n \leq n_P$) corresponds to the j-th of (2) ($1 \leq j \leq n_G$) for some s, we define a set of error functions of the form

$$e_i^{(n)}[\mathbf{X};\ \mathbf{Y}(\mathbf{P})] = w_i^{(n)}\{G_{minj}^{(s)} - F_j^{(s)}[\mathbf{X}^{(s)};\ \mathbf{Y}^{(s)}(\mathbf{P})]\} \qquad (3)$$

where the w's are positive weights. Now, if $e_{max}^{(n)}$ is the largest of the errors $e_i^{(n)}$ (in the algebraic sense), we introduce the n_P-set of combined network functions

$$C^{(n)}[\mathbf{X};\ \mathbf{Y}(\mathbf{P})] = \begin{cases} \left[\sum_i^{+} \left\{e_i^{(n)}[\mathbf{X};\ \mathbf{Y}(\mathbf{P})]\right\}^q\right]^{1/q} & \text{if } e_{max}^{(n)} \geq 0 \\ -\left[\sum_i \left\{-e_i^{(n)}[\mathbf{X};\ \mathbf{Y}(\mathbf{P})]\right\}^{-q}\right]^{-1/\bar{q}} & \text{if } e_{max}^{(n)} < 0 \end{cases} \qquad (4)$$

where the first summation is extended to positive errors only, and $1 \leq n \leq n_P$. $C^{(n)}$ is differentiable if $q > 1$. In addition, $C^{(n)}$ is negative if so are all the errors $e_i^{(n)}$, that is, if all the specifications (2) belonging to the n-th group are satisfied. Thus (2) may be cast in the *equivalent* form

$$C^{(n)}\left[\mathbf{X};\ \mathbf{Y}(\mathbf{P})\right] \leq 0 \qquad (1 \leq n \leq n_P) \qquad (5)$$

Let us now introduce the nonlinear system of n_U equations in n_U unknowns

$$\begin{cases} \mathbf{E}^{(s)}\left[\mathbf{X}^{(s)};\ \mathbf{Y}^{(s)}(\mathbf{P})\right]=\mathbf{0} \\ (1 \le s \le S) \\ \\ C^{(n)}\left[\mathbf{X};\ \mathbf{Y}(\mathbf{P})\right]=-\varepsilon^{(n)} \\ (1 \le n \le n_P) \end{cases} \tag{6}$$

where $\varepsilon^{(n)}$ is a nonnegative margin. Any solution of (6) satisfies both the HB equations (1) and the design specifications (2) at all the design spectra, and thus defines an acceptable design. Since system solving is much more efficient than optimization, this procedure normally leads to a very significant reduction in the required number of iterations, and thus in the overall number of expensive EM simulations. Note that (6) is slightly more restrictive than (1), (2), in the sense that the margins $\varepsilon^{(n)}$ in (6) are a priori fixed, while they are unconstrained in the original problem. Note, however, that the margins are not imposed on specific network functions, but only on the combinations defined by (4), which practically means, on the design goal of each group that is most difficult to fulfill. In this way the fixed margins normally do not have a significant influence on the performance of the design algorithm. In addition, the margins are dynamically updated during the iteration in the following way. As already mentioned in the introduction, after the computation of each Newton update a multidimensional minimization is carried out by a trust-region technique [6]. At the end of this step, the combined network functions (4) are evaluated. If any of such functions is found to take on a negative value, say $C^{(n)} = -v^{(n)}$, so that all the associated specifications are met, the next iteration is run with $\varepsilon^{(n)} = v^{(n)}$. This substantially restores the degrees of freedom of the original formulation (2) of the design goals.

III. Specification of the Far-Field Performance

The passive subnetwork is linear, so that any electromagnetic quantity it supports may be expressed as a linear combination of the gate and drain voltages V_G, V_D applied to its ports (see fig. 1). Since the microstrip antenna is incorporated in the subnetwork, this conclusion is also true for the radiated field. Let us denote by x, y the directions of the microstrip patch edges, by E_x, E_y the far-field scalar components, and by z the broadside direction (see fig. 1). We may write

$$\begin{aligned} zE_x &= A_x V_G + B_x V_D \\ zE_y &= A_y V_G + B_y V_D \end{aligned} \tag{7}$$

where A_x, A_y, B_x, B_y, are positive complex coefficients. The key point is that such coefficients may be computed by an inexpensive post-processing of the data generated by an EM analysis. Far-field quantities such as the cross-polarization suppression $|E_y|/|E_x|$ or the radiation intensity $I_R = |E_y|^2/2\eta$ (where η is the free-space wave impedance) and the like, may thus be easily computed at each iteration and directly included among the design specifications.

IV. An Example of Application

Let us consider the circuit schematically illustrated in fig. 1. The circuit includes a two-port oscillator suitable for injection locking and a microstrip patch antenna integrated on a same substrate. The design variables are 7 layout parameters, as shown in fig. 1. The FET gate voltage is used as the tuning parameter. The design specifications include lower bounds on the radiation intensity (≥ 25 mW/sterad), on the cross polarization suppression (≥ 60 dB), on the spectral purity of the radiated field (total harmonic power ≤ -40 dBc), and on the total drain efficiency (≥ 30%), so that $n_G = 4$. All far-field specifications are referred to the broadside direction. The nominal fundamental frequency of operation is 10 GHz, and the system is required to fulfill the design goals at S = 7 design spectra whose fundamental frequencies are uniformly spaced across the range [9.85 ÷ 10.15 GHz], with a 3 dB margin at the band edges. 4 positive harmonics (including the fundamental) are taken into account in each HB analysis, so that $n_T = 18$. With this problem size, $Sn_G = 28$ specifications have to be subdivided into $n_P = 7$ groups to build the combined network functions (4). This is done by including in each $C^{(n)}$ the four specifications associated with each design spectrum.

In order to produce a reasonable starting point for the EM design, the subsystem is first optimized in a conventional way making use of simple circuit models for both the oscillator and the antenna [1]. The results are then used to generate an initial layout, which is then optimized by the DNO algorithm. A few representative results are reported in figs. 2 - 5. Fig. 2 shows the E-plane far-field radiation pattern at the starting point (as resulting from the circuit-level design) and after the EM optimization. Fig. 3 shows the initial and final radiation pattern of the cross-polarized field, normalized to the radiation intensity in the broadside direction. Fig. 4 shows the total radiation pattern at the harmonics, once again normalized to the radiation intensity in the broadside direction. Finally, in fig. 5 the total drain efficiency is plotted against the number of DNO iterations. Note that for the present purposes the total drain efficiency is defined as

$$\eta = \frac{\text{total radiated power}}{\text{DC power delivered by the bias source}} \quad (8)$$

In all cases, the relatively poor starting-point performance is effectively corrected by the EM optimization, and the final design meets the specifications at all the design spectra. The radiation intensity ripple across a 100 MHz band is less than 0.6 dB (see fig. 2). The design requires a total of 16 EM analyses and about 5 hours of CPU time on an 800 MHz PC.

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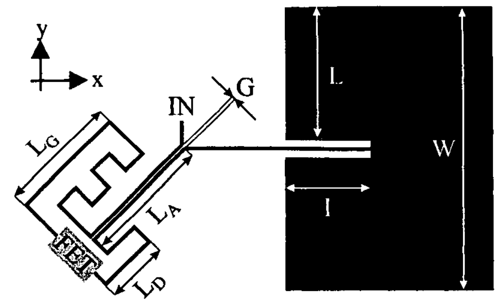

Fig. 1 Schematic layout of a self-oscillating antenna

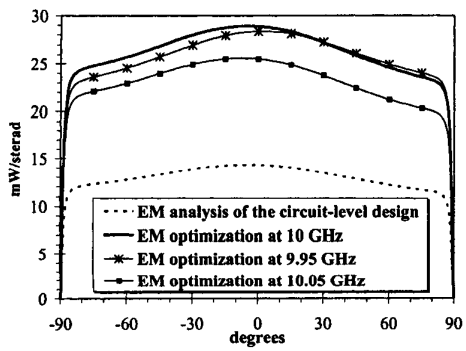

Fig. 2 E-plane far-field radiation pattern

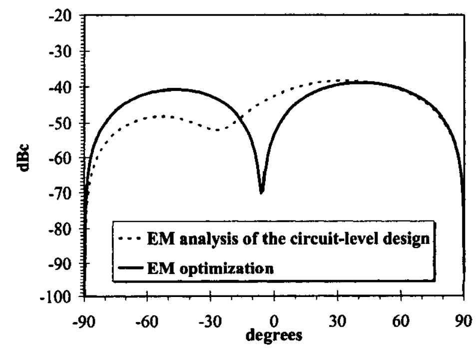

Fig. 3 Radiation pattern of the cross-polarized field

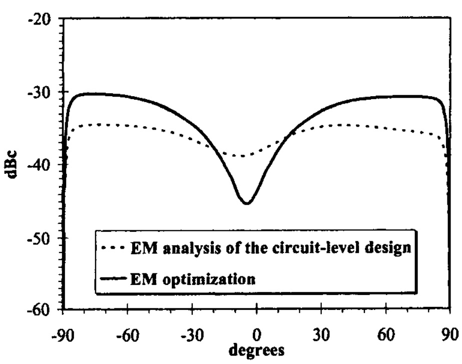

Fig. 4 Total radiation pattern at the harmonics

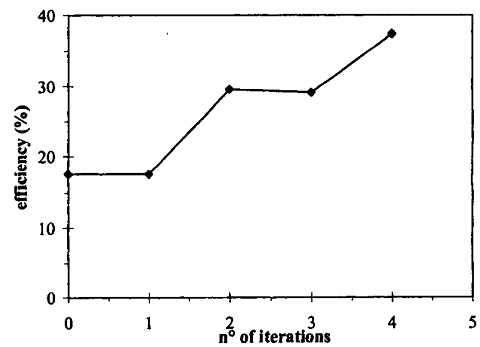

Fig. 5 Total drain efficiency vs. n° of iterations