Nonlinear/electromagnetic co-design of microwave transceivers including integrated antennas (invited paper)

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Abstract: The paper reports on the coupled nonlinear/electromagnetic design of active LTCC transmitters including integrated antennas, based on neural models of the entire passive circuit layout. EM tools [1], [2] are used to train multiple neural network models of both the linear subnetwork and the antenna. A separate artificial neural network (harmonic neural network HNN) is used in the neighbourhood of each harmonic of interest in order to overcome the bandwidth problem. Special ANN are used to model the radiated field, allowing the design goals to be formulated in terms of far-field specifications, and suitable figures of merit encompassing radiation and nonlinear performance are introduced. The transmitter analysis and optimization is then carried out by connecting the nonlinear devices to the EM-derived multiport network.

Keywords: Nonlinear/EM co-design, LTCC technology, Neural network models, EM tools.

1. Introduction and outline of the method

LTCC (low-temperature co-fired ceramics) technology provides an effective approach to the realisation of microwave integrated circuits coupling small size and low cost with high functional complexity (up to the transceiver level), which may be especially interesting for volume applications such as RF-ID systems. These circuits consist of several thin ceramic layers (typically a few tens of micrometers thick) supporting metallisations of arbitrary shapes at each interface plane, with different metallisation layers connected by via holes. Additional layers supporting active devices and integrated antennas may be added on top of the stack to build entire transceivers. Due to this structural complexity, electromagnetic couplings among different parts of an LTCC system are ubiquitous, especially if antennas are present. This makes EM simulation virtually indispensable for an accurate characterisation. Circuit design by numerical optimisation then becomes a formidable task from a numerical viewpoint. A possible solution is to use EM analysis before starting the optimisation to generate an accurate and computationally fast model of the linear subnetwork (LS), e.g., by means of artificial neural networks (ANN). The model is then directly handled by the optimisation algorithm in much the same way as an ordinary circuit model. This method has been successfully applied by several authors (e.g., [3]). The method presented here extends this standard procedure in two ways. As a first aspect, owing to the nonlinearity of the system of interest, the linear subnetwork must be characterised at all harmonics. Thus in principle the neural model should accurately reproduce the dependence of the S- or Y-parameters on all the designable layout parameters across a very broad frequency band, which would be very hard to achieve (if possible at all) in most practical cases. In order to overcome this difficulty, we make use of a different ANN (called a harmonic neural network, HNN) to describe the linear subnetwork in a comparatively narrow band located in the neighbourhood of each harmonic of interest. The second extension is the development of an ANN model of the radiated far-field, to be used in conjunction with the optimisation of radiating components. This model is built by training a set of auxiliary HNN on the basis of the far-field information produced by each EM analysis of the main training set. In order to explain this idea, we first note that the LS is linear, exactly as the radiation phenomenon it originates. The far field is thus a linear homogeneous function of its sources, which for the present case may be identified as the voltages at the LS ports. If $\mathbf{V_D}$ is the vector of such voltages, we may write

$$\mathbf{E}(r, \theta, \phi) = \frac{\exp(-j\beta r)}{r} \mathbf{A}(\theta, \phi) \bullet \mathbf{V_D} \tag{1}$$

where r, θ, φ are spherical coordinates, **E** is the radiated far-field that the EM simulator automatically computes after each analysis. **A** is a matrix of complex coefficients that may be determined by cheap linear operations from the EM analysis results. In this way the training and validation sets for the auxiliary HNN may be generated at a low additional cost. The inputs to these networks are once again the designable layout parameters, while the outputs are the entries of **A**. At each iteration, the LS admittance parameters are computed by the main HNN and are used to carry out a nonlinear analysis by the harmonic-balance (HB) technique. This provides the vector $\mathbf{V_D}$, which in turn is used to compute both the ordinary network functions (by conventional circuit analysis), and the far field by (1). Far-field quantities such as the radiation intensity or the cross-polarisation suppression may thus be directly included among the design goals. Note that the design variables normally include the numbers of dielectric layers interposed between the different metallisation layers.

2. Transmitter optimization

In fig. 1 the EM-based representation of a front-end-antenna assembly is shown [4]. The transmitter is directly designed by numerical optimization with respect to a set P of physical and/or layout parameters by simultaneously accounting for all EM coupling and radiation effects in a unified and most general way. Since the antenna is an integral part of the transmitter assembly, which is treated as a unique EM system, a front-end output port and an antenna input port are no longer available. It thus becomes impossible to define in a meaningful way, and thus to individually compute, some familiar concepts such as antenna impedance and antenna and front-end gain that are normally used in link design. We are thus facing the open problem of suitably defining a transmitter effective conversion gain G_{EC} to be used as a network function in the optimization process. G_{EC} should implicitly encompass the ordinary front-end and antenna gains without being explicitly based on any of such quantities.

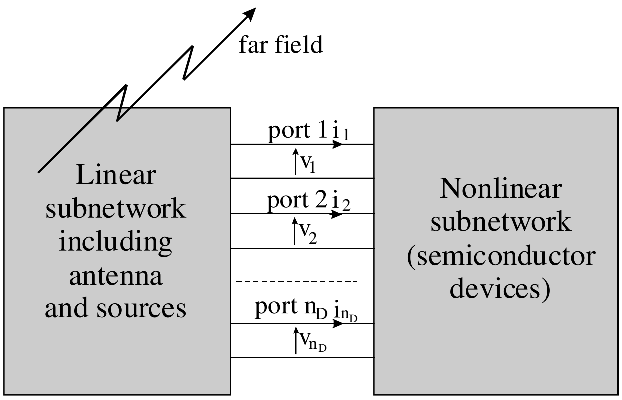


Fig. 1. EM-based representation of a transmitter front-end including the antenna.

In order to develop an expression for G_{EC}, we first note that the quantity really of interest for transmitter design purposes is the far-field power density incident upon the receiving antenna, which from (1) may be cast in the form

$$\frac{\left\|\mathbf{E}(r_L, \theta_L, \phi_L; \omega_{RF})\right\|^2}{2\eta} = \frac{\left\|\mathbf{A}(\theta_L, \phi_L; \omega_{RF}) \bullet \mathbf{V}_{\mathbf{k_R}}\right\|^2}{2\eta r_L^2} \tag{2}$$

where r_L is the link length and η is the free-space wave impedance and $\mathbf{V}_{\mathbf{kR}}$ is a subvector of $\mathbf{V_D}$ regarding ω_{RF}. Indeed, (2) directly determines the power that the receiving antenna can extract from the incident field and deliver to the receiver front end, so that it is quite reasonable to use (2) as the system output as a replacement for the ordinary output power. In order to meaningfully define a gain, we now need a reference quantity homogeneous with (2) and associated with a unit gain situation, which means, a unit gain front end feeding a unit gain antenna. Let us assume that the IF source is a voltage generator of impressed voltage U_G and internal impedance RG, so that the available input power at IF is

$$P_{IF} = \frac{|U_G|^2}{8R_G} \tag{3}$$

In our reference situation, the power (3) is up-converted to RF and is totally delivered to a matched unit-gain antenna, i.e., an isotropic radiator. In such conditions, the far-field power density radiated at a distance r_L in any direction of space would be

$$\frac{\left\|\mathbf{E_I}(r_L; \omega_{RF})\right\|^2}{2\eta} = \frac{P_{IF}}{4\pi r_L^2} = \frac{|U_G|^2}{32\pi R_G r_L^2} \tag{4}$$

The effective conversion gain is thus given by

$$G_{EC} = \frac{\left\|\mathbf{E}(r_L, \theta_L, \phi_L; \omega_{RF})\right\|^2}{\left\|\mathbf{E_I}(r_L; \omega_{RF})\right\|^2} = 16\pi \frac{R_G}{\eta} \frac{\left\|\mathbf{A}(\theta_L, \phi_L; \omega_{RF}) \bullet \mathbf{V}_{\mathbf{k_R}}\right\|^2}{|U_G|^2} \tag{5}$$

In our optimization approach an HB analysis is carried out each time P is updated, and the above procedure is used together with ordinary circuit analysis to find the network functions and to compute a least-pth objective. The latter is minimized with respect to P by a quasi-Newton algorithm. A practical way of implementing this nested loop architecture is discussed in the next section.

3. **Harmonic neural networks**

In order to avoid direct EM simulation inside the optimization loop, EM analysis may be used before starting the optimization to generate accurate and computationally fast models of the entire linear subnetwork (LS) or of parts of it. These models may then be directly handled by the optimization algorithm in much the same way as ordinary circuit models. A practical way of finding accurate approximations to a number of nonlinear functions of several variables such as the linear subnetwork S-parameters is to use artificial neural networks (ANN) [3]. Due to the need for taking several harmonics into account, the LS neural model should accurately reproduce the dependence of the S-parameters on all the designable layout parameters across a very broad frequency band, which would be very hard to achieve in most practical cases. In order to overcome this difficulty, we make use of a different ANN (called a harmonic neural network, HNN) to describe the LS in a comparatively narrow band located in the neighbourhood of each harmonic of interest. In order to demonstrate the whole procedure, we discuss the design of a simple transmitter assembly consisting of an antenna fed by a doubly-balanced 4-FET mixer (Gilbert cell) [4]. It is assumed that the subsystem (including the antenna) has to be realized by an integrated multiple-layer technology such as LTCC. Assuming that the operating frequencies are f_{RF} = 2.4 GHz, f_{LO} = 2.47 GHz, f_{IF} = 70 MHz, and that the spectrum used in each HB analysis includes 4 LO harmonics plus one lower and one upper sideband per LO harmonic, we have a total of 13 positive frequencies plus DC. The overall band over which the performance of all circuit components has to be accurately controlled is thus [0 ÷ 9.95 GHz]. The RF transformer of the transmitter to be realized in the form of a distributed integrated balun is shown in fig. 2: the transmission line sections are realized by spiral metal strips and the capacitor electrodes by parallel metal plates. The lines are stacked in order to save substrate area, and metal vias are used for interconnections between layers. In order to build an efficient numerical model we use five EM-based HNN [1], one in the neighbourhood of each LO harmonic plus one in the neighbourhood of the IF frequency. Each HNN is a standard 3-layer perceptron whose number of hidden neurons is chosen in such a way as to minimize the training error. The training algorithm is based on the Levenberg-Marquardt method. The input parameters of the HNN are frequency, the number of dielectric layers (each of fixed thickness equal to 30 μm) separating each couple of metal layers, and the length of the capacitor metal plate. The output parameters are the real and imaginary parts of the balun S-parameters.

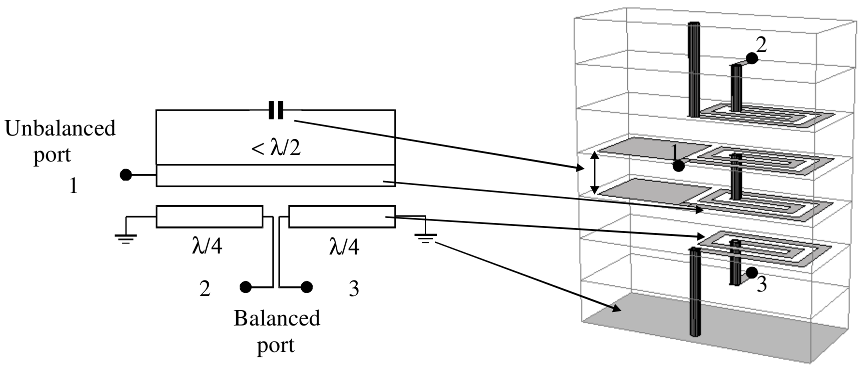
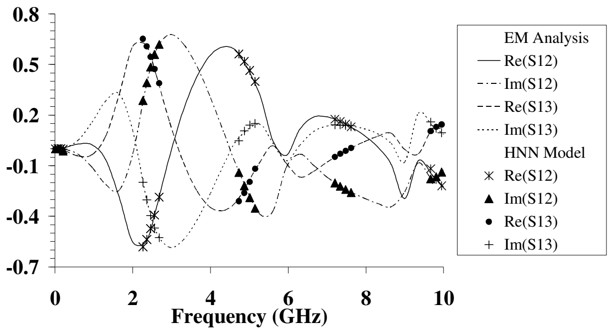


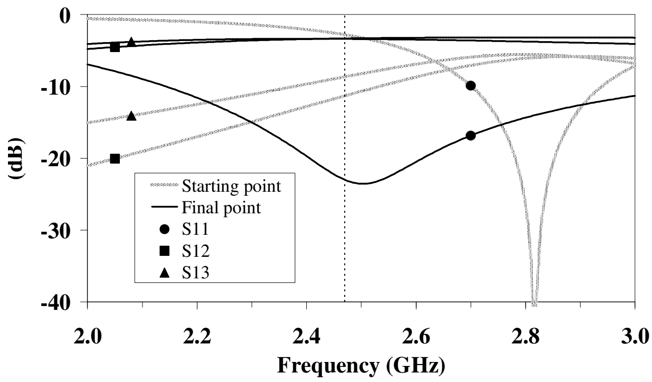
Fig. 2. Circuit schematic of the LO/RF balun and its LTCC implementation.

An automated mechanism has been implemented to drive the required EM analyses so that the EM setup is defined only once before starting the data generation. A number of points are randomly selected in the layout parameter space in order to build the training and validation sets, and the corresponding parameter values are automatically input to the EM simulator. The training set is gradually refined until each HNN is able to reproduce the EM-computed S-parameters with a prescribed accuracy at all points of the validation set. In fig. 3 (a) the S-parameters obtained by direct EM analysis and by the HNN model are compared for a randomly selected layout not included in the training set nor in the validation set. An RMS error of about 0.05% across each frequency band of interest is measured between the EM-simulated [1] and HNN-generated S-parameters. Excellent agreement is obtained at all the frequency points belonging to the selected spectrum.

By means of the modeling approach described in the previous section, the ideal RF and LO transformers are replaced by integrated baluns .And a complete view of the integrated transmitter including the antenna is given in fig. 4 (b). The transmitter is then designed as a whole and the goals include an effective conversion G_{EC} > 6 dB and a 16 dB return loss (with respect to 50 Ω) at the LO port. Note that in this way we need not require the antenna radiation impedance to be 50 Ω (or any other arbitrarily preselected value), which adds considerably to the design flexibility. For the HB analysis of the entire assembly, HNN models are derived for all linear components in the way discussed in the previous section. On the other hand, only the one HNN describing each component in the neighbourhood of its nominal operating frequency is used in each individual design. This is illustrated in fig. 3 (b) for the RF balun. The optimization is carried out with respect to all the designable parameters, with starting values directly derived from the circuit schematic. Only the HNN describing the circuit performance in the neighbourhood of the LO frequency is used in this design. The initial performance reported in fig. 3 (b) is rather poor, showing the need for EM analysis to accurately control the component performance.



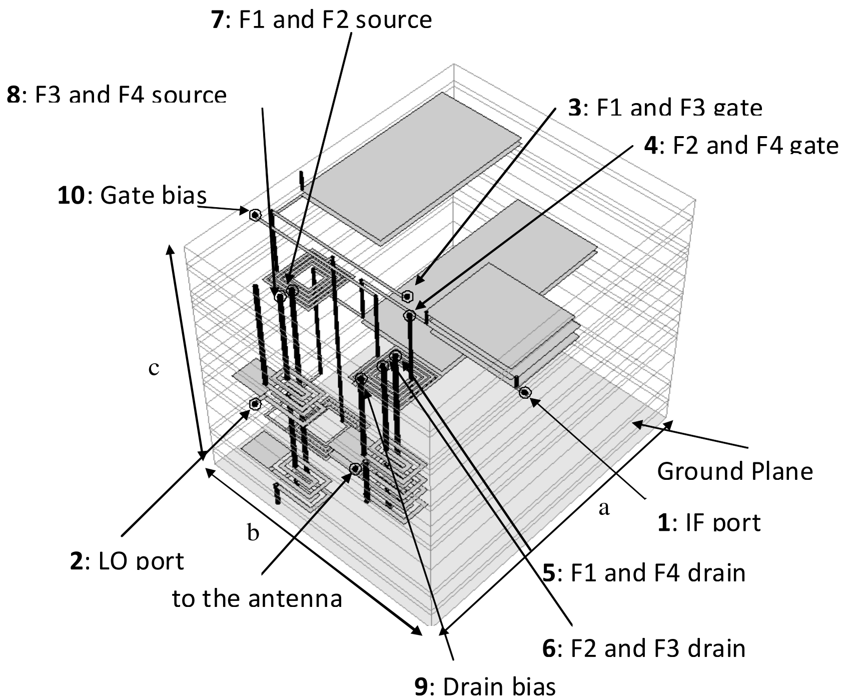
(a)



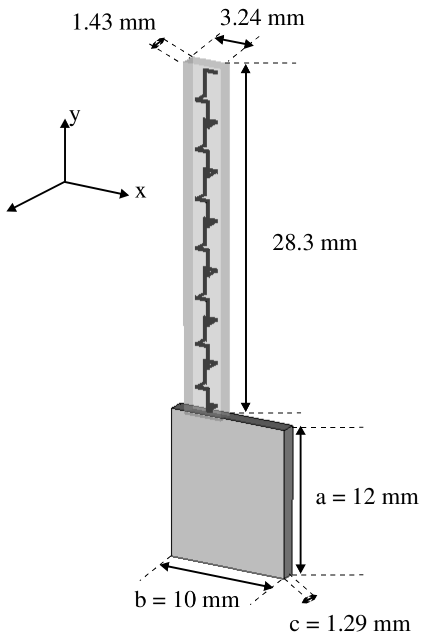
(b)

Fig 3. (a): EM analysis and harmonic neural network of the LO/RF balun, (b): Balun performance before and after optimization.

The ANN-based optimization takes about 2 CPU seconds on a 2.8 GHz PC, and leads to the final performance also shown in fig. 3(b), meeting all the design goals. The LO and IF baluns are treated in a similar way. The front-end assembly of fig 4 (a), consisting of a total of 11 ports, is then analysed as a whole by EM simulation [1].



(a)



(b)

Fig. 4. (a) LTCC front-end layout to be analysed by EM tools [1], (b) view of the antenna-transmitter assembly.

To design the integrated monopole antenna we first have to train an auxiliary HNN in order to approximate the relationships (1) in the neighbourhood of the RF, so that far-field related quantities may be included in the design process. The antenna (see fig. 4 (b)) consists of a multilayer spiral monopole whose cross-sectional dimensions are small with respect to the wavelength, so that the overall device is equivalent to a magnetic monopole parallel to the spiral axis. The designable layout parameters are the spiral cross-sectional width, the number of turns, and the spiral pitch. The auxiliary HNN is trained on the basis of the far-field information produced by each EM analysis [2] of the main training set by means of cheap linear operations defined by (7). The inputs to this neural network are the designable layout parameters, while the outputs are the entries of the matrices A_θ, A_ϕ.

The antenna is integrated on the same dielectric layers as the LTCC linear subnetwork, and its input line is connected to the unbalanced RF port of the mixer. The assembly of figure 4 (b), is now treated as a whole by means of EM simulation. A nonlinear optimization with respect to both the mixer and the antenna layout parameters (8 variables overall) using the HNN model of the entire assembly is then carried out. This process converges in about 83 s and yields the final performance given in fig. 5, satisfying all the design goals. A comparison between the performance computed by the HNN models and by a direct EM simulation of the entire device is also provided in the same figures. A small discrepancy is observed due to the approximation errors that are unavoidable in the neural models of such a complex configuration, but the EM results are still well within the design specifications.

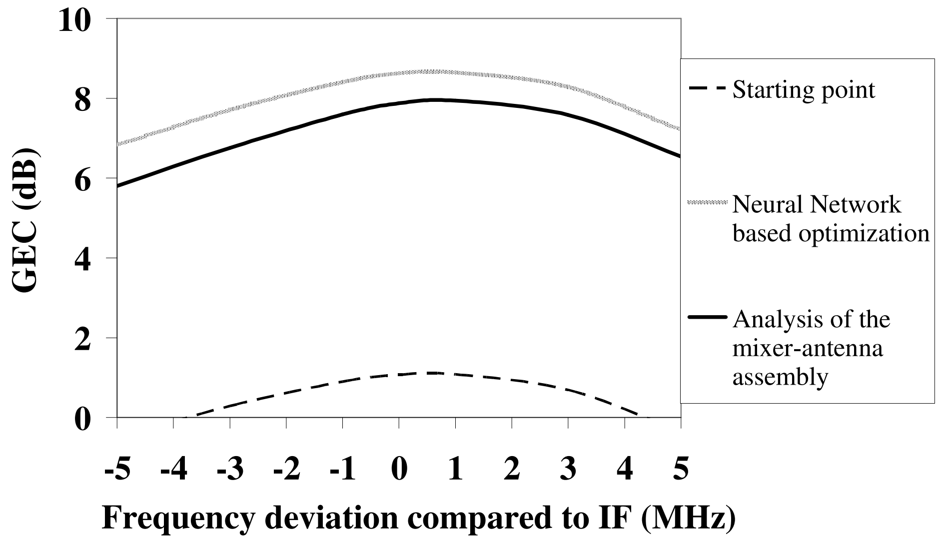


Fig. 5. Effective conversion gain of the mixer-antenna assembly.

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