# EM/Circuit Mixed Simulation Technique for an Active Antenna

G. Zakka El Nashef, F. Torres, S. Mons, T. Reveyrand, T. Monedière, E. Ngoya, R. Quéré

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# EM/Circuit Mixed Simulation Technique for an Active Antenna

Georges Zakka El Nashef, F. Torrès, S. Mons, T. Reveyrand, *Member, IEEE*, T. Monédière, E. Ngoya, *Member, IEEE*, and R. Quéré, *Fellow, IEEE*

***Abstract*—Today's increase of functions, improvement of performance, and cost reductions required on an active electronically scanned array (AESA), associated to the limited amount of available areas and volumes to implement the equipment, drive an approach leading to directly connect power amplifiers (PAs) to the antennas array without placing an isolator/circulator between them. In this letter, an electromagnetic/radio frequency (EM/RF) circuit mixed simulation technique will be theoretically introduced and experimentally demonstrated on transmission (Tx) chains to deal with the proposed challenge.**

***Index Terms*—Active electronically scanned array (AESA), behavioral model, electromagnetic (EM) macro-model, mutual coupling, power amplifier (PA).**

## I. Introduction

THE SIMULATION and analysis of a transmission (Tx) chain have to face two types of problems: electromagnetic (EM) characterization of the transmitting antennas and rigorous radio frequency (RF) circuit approach. The combination of these issues is commonly referred to as EM/RF circuit cosimulation. In general, cosimulation can be seen in many application areas as encompassing nonlinear circuit conception, antenna design, and signal integrity analysis in a global approach. In our case, cosimulation approach is used to study interactions between PAs and antennas.

Active electronically scanned array (AESA) design constraints are very critical since greater performances are required within less room or nonconvenient places to fit the equipment, and this compactness leads to many problems, such as heat dissipation, EM coupling, etc. More precisely, in our case, short relative distance between the array's antennas may result in a high level of mutual coupling between antennas, leading to large output loading mismatches [up to voltage standing wave ratio (VSWR) = 4:1 and more]. This mismatching affects PA behavior, which in turn directly impacts the gain and phase controls of each radiating element. Therefore, the cosimulation approach is of great help to investigate the effects of large output mismatching ($Z_{\rm load} \neq 50\ \Omega$) and high-level mutual coupling between antennas on the system performance.

However, such general approaches may require large and complex simulations. Therefore, an alternative method is used in this letter, the "mixed simulation approach." It consists in dividing and solving separately both the RF circuit model and EM model, which can then be combined to produce a general behavioral model for system approach, enabling an efficient synthesis and optimization.

Due to space limitations and already published theoretical works in [2] and [6], a brief summary of the two different approaches (Sections II-A and II-B) will be presented. Those approaches enable us to have simulation models of separated blocks [antennas and power amplifiers (PAs)] to ensure that overall performance will meet requirements in given worst-case conditions. Then, a preliminary mixed simulation approach combining both models is presented in Section II-C. An experimental validation is described in Section III. Conclusions are given in Section IV.

## II. Mixed Simulation Concepts: PA and EM Models

### A. PA Behavioral Model

In this letter, a PA is directly connected to an antenna, which then presents an unknown load impedance to the PA. The mutual coupling between antennas may induce strong variations in both the real and imaginary part of the antenna input impedance. Hence, without an isolator, the load presented to the PA will vary and the PA characteristics will change and cause a significant signal distortion [1]. Therefore, an extended bilateral behavioral model was developed in [2], based on nonlinear scattering functions [3], allowing to take into account large output mismatches (VSWR up to 4:1). The developed model [2] uses second-order Taylor expansion to improve robustness and accuracy. The phase of the four waves ($\tilde{a}_i$ *and* $\tilde{b}_i$) is normalized such as $\tilde{a}_1$ becomes real. Thus, assuming that $\tilde{a}_2$ is negligible compared to $\tilde{a}_1$, and that the PA is working at the fundamental frequency, the second-order Taylor development can be written as follows:

$$\begin{aligned}\tilde{b}_i =& S_{i1}(f_0,|\tilde{a}_1|)\cdot\tilde{a}_1 + S_{i2}(f_0,|\tilde{a}_1|)\cdot\tilde{a}_2\\ &+T_{i2}(f_0,|\tilde{a}_1|)\cdot\tilde{a}_2^* + T'_{i2}(f_0,|\tilde{a}_1|)\cdot\tilde{a}_2^2\\ &+T''_{i2}(f_0,|\tilde{a}_1|)\cdot\tilde{a}_2^{*2} + T'''_{i2}(f_0,|\tilde{a}_1|)\cdot\tilde{a}_2\cdot\tilde{a}_2^* \end{aligned}\quad (1)$$

where $S_{ij}(|\tilde{a}_1|)$, $T_{i2}(|\tilde{a}_1|)$, $T'_{i2}(|\tilde{a}_1|)$, $T''_{i2}(|\tilde{a}_1|)$, and $T'''_{i2}(|\tilde{a}_1|)$ are the nonlinear scattering functions.

The PA model can be extracted from continuous-wave (CW) measurements at the operating frequency, as described in [4]. The identification of nonlinear scattering functions defined in (1) can be obtained from waves' measurements for at least six

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The authors are with the XLIM Laboratory, OSA/C2S2 Departments, UMR CNRS 06172, University of Limoges, 87060 Limoges Cedex, France (e-mail: georges.zakka-el-nashef@xlim.fr).

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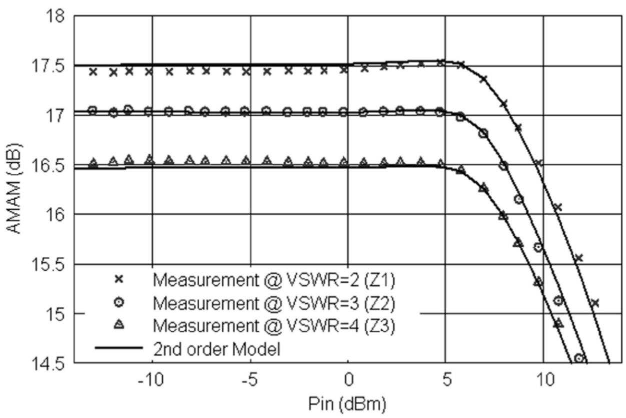


Fig. 1. Fundamental gain compression (AMAM) versus input power for several loading impedances ($Z1 = 36.8 - \mathrm{j}\cdot 27.1, Z2 = 35.2 - \mathrm{j}\cdot 5.8, Z3 = 14.8 + \mathrm{j}\cdot 10.4$). Second-order model (lines) compared to measurement (symbols). $F = 8.2$ GHz.

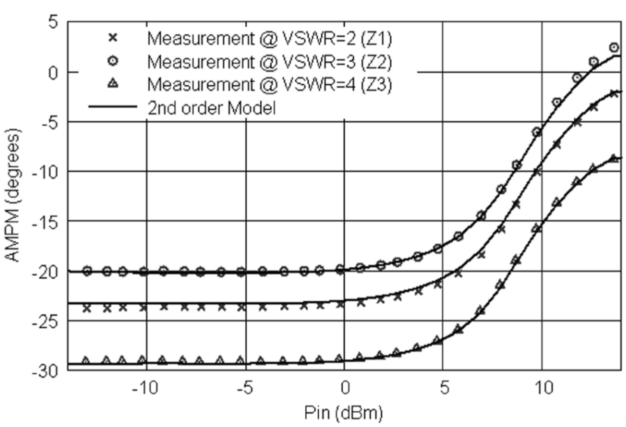


Fig. 2. Fundamental phase variation (AMPM) versus input power for several loading impedances ($Z1 = 36.8 - \mathrm{j}\cdot 27.1, Z2 = 35.2 - \mathrm{j}\cdot 5.8, Z3 = 14.8 + \mathrm{j}\cdot 10.4$). Second-order model (lines) compared to measurement (symbols). $F = 8.2$ GHz.

different impedances (12 parameters need to be extracted), as explained in [2]. Depending on available data, one may also consider more than six impedance measurements and solve (1) using least-square minimization.

To validate the model, we have considered a commercial 8–14-GHz PA from NEXTEC-RF (NB00422). The fundamental measured gain (AMAM) and phase (AMPM) characteristics at the operating frequency (8.2 GHz) are compared to the model's response for different loading impedances, up to VSWR $=$ 4:1 (Figs. 1 and 2). The results show a good agreement between the second-order model and PA measurements. They clearly demonstrate the efficiency and accuracy of the second-order model, even in the presence of a large output loading impedance (e.g., VSWR $=$ 4:1), which is not the case for a first-order bilateral model [2].

### *B. EM Macro-Model*

Concerning the EM part, the mismatching problem between the antennas and PAs will modify the performance of the array in terms of radiation pattern: Mismatching affects PA behavior as explained in Section II-A, and consequently, the necessary magnitude and phase weights for the array in a given direction will be also modified once applied to PAs, degrading the array efficiency and its radiation performance [1]. Therefore, an array macro-model is needed in order to calculate, for each array element, the matching impedance ($Z_{\mathrm{active}} \neq 50\ \Omega$) as well as the feeding weights (magnitude and phase), according to pointing angle and frequency.

The macro-model mathematical development leading to (2) and (3) is explained in [6]

$$[Z_{\mathrm{active}}] = \left[\mathrm{diag}\left(\sqrt{[Z_L]}\right) \times [I + S_c] \times \mathrm{diag}[a]\right] \times [A] \times \left(\left[\mathrm{diag}\left(1/\sqrt{[Z_L]}\right) \times [I - S_c] \times \mathrm{diag}[a]\right] \times [A]\right)^{-1} \quad (2)$$

$$[a] = [U] \times \sqrt{\Re e\left([Z_{\mathrm{active}}]\right)} \Big/ [Z_{\mathrm{active}}] \quad (3)$$

where matrices $[U]$ and $[I]$ respectively are the driven voltages and currents. $[a]$ and $[Z_L]$ are respectively the waves and ports impedances that are used to supply each antenna during the EM analysis of the array structure. $[S_c]$ is the array's scattering matrix extracted through a rigorous EM simulation (CST Microwave Studio). Matrix $[A]$ is the complex coefficients (weights) for each antenna, calculated with an array synthesis software (SARA [6]), in order to obtain the best combination of the $N$ individual radiation patterns. $\sqrt{[Z_L]}$ and $\sqrt{\Re e([Z_{\mathrm{active}}])}$ denote the square root of each matrix element.

One of the originalities of this EM macro-model, besides taking into account the mutual coupling, calculating the matching impedances and their corresponding weights, is the integration of a synthesis software (SARA) that enables us to optimize the radiation pattern of the array in directivity, gain, and sidelobe level and to obtain a reflected power wave $(b_i)$ almost equal to zero, which will not disturb the array nor PA functioning. It is thus possible to eliminate isolators between PAs and antennas. The experimental validation described in Section III will prove the reliability of this concept.

An experimental/numerical evaluation of the macro-model performance is made. A $1 \times 4$ patch array working at 8.2 GHz has been designed and built. The interelement spacing was reduced to $0.4\lambda_0$ in order to increase the mutual coupling. Then, the matching impedances and complex coefficients (magnitude and phase) were calculated for different pointing angles using the EM macro-model. From the experimental point of view, a setup including a power splitter (one- to four-way), four attenuators, and four phase shifters connected to the $1 \times 4$ patch array was used (Fig. 3).

On the numerical side, the whole array has been modeled using CTS MWS, each antenna being fed by the calculated weights (magnitude and phase) and loaded with corresponding matching impedance as well. Figs. 4 and 5 respectively compare the simulated results and measured ones for the considered array with a $-10°$ and $+15°$ pointing angle.

We can observe a good agreement between the simulated and measured results. The pointing angle is maintained, and sidelobe level is a little bit degraded. These results prove that EM macro-model is efficient and can be used as a synthesis tool as well.

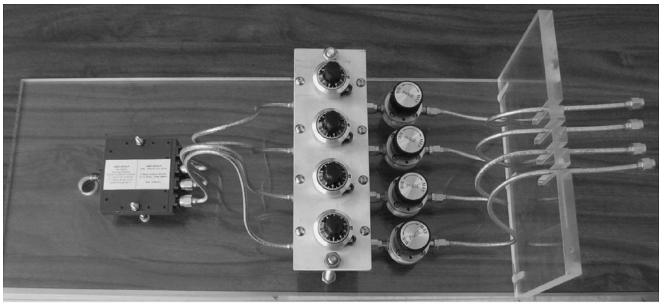

Fig. 3. EM macro-model demonstrator.

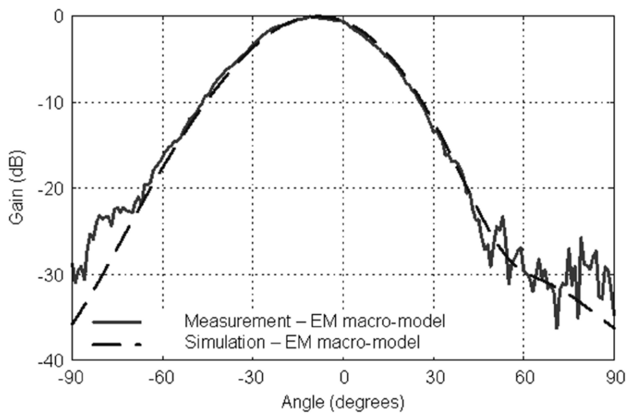


Fig. 4. Radiation pattern comparison $(-10^\circ)$. EM macro-model measurement (solid line). EM macro-model simulation (dashed line).

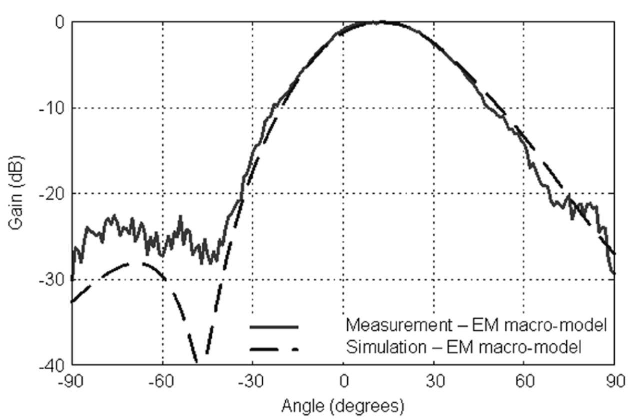


Fig. 5. Radiation pattern comparison $(+15^\circ)$. EM macro-model measurement (solid line). EM macro-model simulation (dashed line).

### C. Mixed Simulation Approach

Here, a mixed simulation approach is evaluated by integrating the EM macro-model into Advanced Design System (ADS). This approach is realized by defining the calculated impedances obtained from the EM macro-model as an output load for PA and measuring/simulating the overall performance of the studied "virtual" system characterized by PA and output loads corresponding to each antenna.

The PA model has been implemented in ADS using frequency-domain defined device (FDD) nonlinear blocks [2]. A variable load is connected to the output to define the input impedance of the antenna. Equation (1) appears in a text file that contains all the measured data and a special tool "Data Access Component" (DAC), which links the text file to the model and performs a two-dimensional interpolation during the simulation. Figs. 6 and 7 show the comparison between the PA model (implemented into ADS) and measurements in terms of gain (AMAM) and phase (AMPM) for two different antenna impedances calculated by the EM macro-model. The measurements of the PA incident and reflected waves were performed using a load-pull bench, as explained in [4]. The calculated impedances were controlled by a tuner.

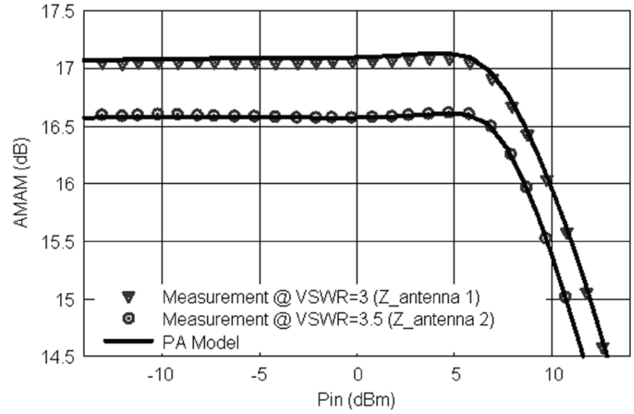


Fig. 6. Fundamental gain compression (AMAM) versus input power for two calculated impedances ($Z\_\mathrm{antenna1} = 21.1 - j \cdot 18.8, Z\_\mathrm{antenna2} = 48.4 + j \cdot 55.9$) using the EM analysis tool. PA model (lines) compared to load pull measurement (symbols). $F_0 = 8.2$ GHz.

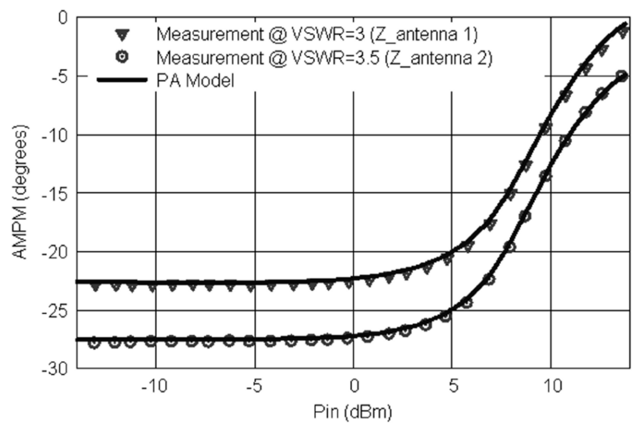


Fig. 7. Fundamental phase variation (AMPM) versus input power for two calculated impedances ($Z\_\mathrm{antenna1} = 21.1 - j \cdot 18.8, Z\_\mathrm{antenna2} = 48.4 + j \cdot 55.9$) using the EM analysis tool. PA model (lines) compared to load pull measurement (symbols). $F_0 = 8.2$ GHz.

Figs. 6 and 7 show a perfect agreement, and this again demonstrates the efficiency and accuracy of the PA model, even in the presence of output loading impedances characterizing the antennas. Thus, the knowledge of magnitude and phase variations due to PAs feeding antennas' mismatched loads will then allow us to correct the feeding weights of the array to finally obtain the desired radiation pattern, and it enables robust, accurate, and useful simulation of distortion at the circuit level.

## III. Experimental Validation

An experimental demonstration of the mixed simulation approach capabilities is presented here. Fig. 8 depicts the used architecture where we considered the same passive design used to

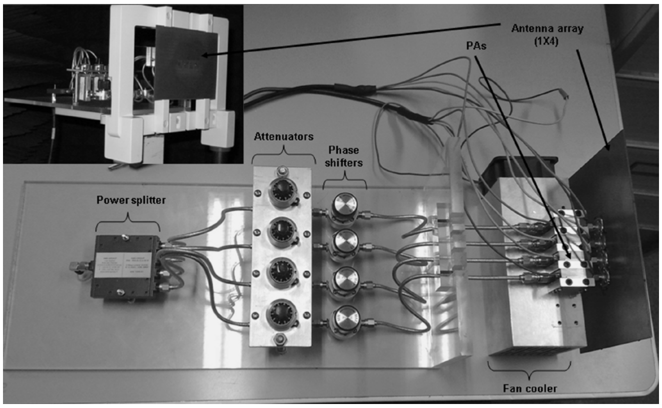


Fig. 8. Active antenna demonstrator.

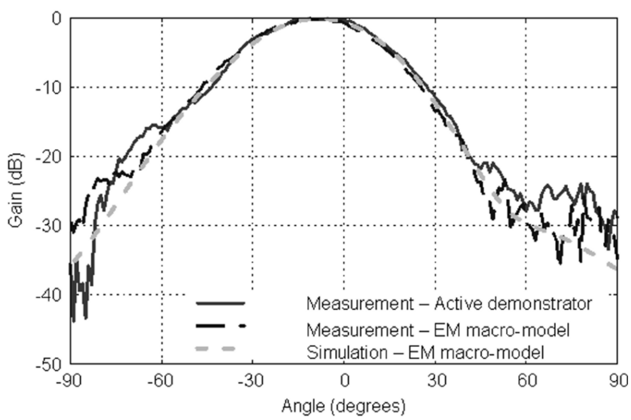


Fig. 9. Radiation pattern comparison $(-10^{\circ})$. Active demonstrator measurement (solid line). Measurement without PA (big dashed line). EM macro-model simulation (small dashed line). $F_0 = 8.2$ GHz.

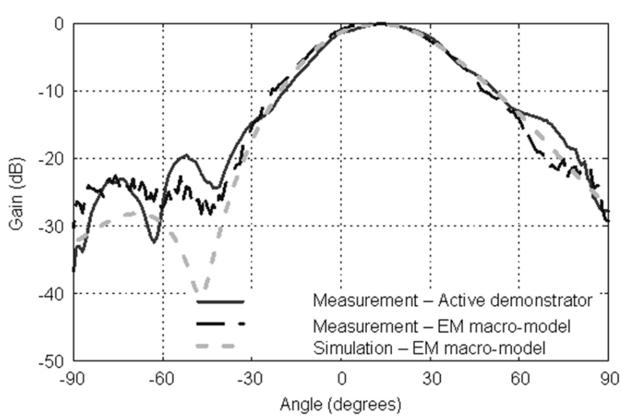


Fig. 10. Radiation pattern comparison $(+15^{\circ})$. Active demonstrator measurement (solid line). Measurement without PA (big dashed line). EM macro-model simulation (small dashed line). $F_0 = 8.2$ GHz.

validate the EM macro-model concept (Fig. 3). PAs are now directly inserted between the phase shifters and array's antennas, the absence of isolators between PAs and antennas being a special feature of this architecture. As previously mentioned, the idea is to correct signal—i.e., magnitude and phase—at each PA input in order to ensure both the desired weights at each antenna access and a null reflected wave: From the PA's model response in terms of AMAM and AMPM at an available input power range $(\mathrm{P_{in}})$, AMAM $(\delta|\mathrm{a}|)$ and AMPM $(\delta\phi)$ values are obtained. The desired weights ($|\mathrm{a}|$ and $\phi$) that were calculated by the EM macro-model can be established by feeding each PA by a corrected magnitude $(|\mathrm{a}| - \delta|\mathrm{a}|)$ and phase $(\phi - \delta\phi)$.

Figs. 9 and 10 compare the measured radiation patterns with PA (active antenna demonstrator), the measured radiation patterns without PA (EM macro-model validation), and simulated radiation patterns without PA (EM macro-model validation) for a $-10^{\circ}$ and $+15^{\circ}$ pointing angle.

Results show a good agreement between the main lobes for the measured results (with and without PA) and simulated ones. We can notice slight differences on the sidelobe levels, which are likely caused by the asymmetry of feeding antennas probes. Another reason for these differences may come from the inaccuracies during the measuring procedure. Eventually, we have demonstrated the efficiency of the mixed simulation approach proving the possibility of a direct connection between PAs and antennas without the use of an isolator.

## IV. Conclusion

This letter described briefly the theory and measurement procedure needed to accurately study the global interactions between the antennas and PAs. This work was founded on two complementary approaches. First, a bilateral behavioral model that deals with large output mismatches was experimentally validated. Then, an EM array macro-model that takes into account the mutual coupling between antennas and provides the input impedances and feeding weights was also experimentally validated. In both cases, excellent agreement is achieved between measurements and models results. Later, a mixed simulation approach was presented and validated. Finally, we have proven that the mixed simulation concept is effective. Measurements for different radiation patterns, with or without PA, demonstrate the robustness and accuracy of the mixed simulation concept and provide an elegant solution to new designs capabilities for AESA applications with the elimination of isolators from a Tx chain.

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