A Methodology for multi-band class E RF PA design.

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***Abstract* — This paper presents a simple methodology for the design of a highly efficient PA in the context of multi-radio devices. This PA belongs to the radiofrequency reconfigurable part within a transmitter. Our goal is to improve the PA efficiency over the bandwidth. The method is based on the extraction of the transistor's non-linear output parameters and the optimisation of a given class E topology to an ideal impedance load. Herein, the method is illustrated by an integrated design using between 2 and 4 GHz and the results of the simulations (in terms of efficiency) agree well with the transient simulation and the S-Parameters optimisation. Considerations are given for the complete development of this design and the interest for future multi-band front end designs.**

***Index Terms* — RF architecture, multi-radio, class E PA.**

I. THE MULTI-RADIO CONTEXT

Wireless communications are widely used for our daily needs. There are numerous examples which use between 2 and 4 GHz and standard names like BLUETOOTH, WiFi, WiMAX, UMTS and LTE are well-known. An important aspect of today's applications is the increasing data rate need, especially in connectivity standards (WiFi, WiMAX), because of the user's high Quality of Service (QoS) demands. To increase the data rate, we tend to use wideband or multi-standard architecture [1] [5]. The concept of software radio includes a self-reconfigurable radio link. While focusing on the radiofrequency (RF) part, the term multi-radio is preferred. A multi-radio RF device is supposed to use different RF flexible blocks, whose parameters are the centre/carrier frequency, bandwidth, modulation scheme and average power, see figure 1.

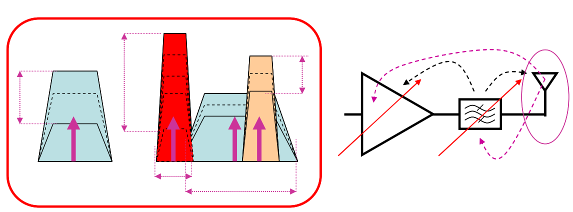

figure 1 : Multi-radio front-end challenges

Also, the signal dynamics (represented by the Peak to Average Power Ratio) should not decrease the transmitter's performance. Due to the different PAPRs of the standard signals (OFDM for example), the transmitter has to be linearised, because the power amplifier (PA) introduces crippling non-linear effects (NL) [13] [4]. The present challenge is to linearise the RF architecture while keeping high efficiency, because this is related to battery-life (especially in our context of nomadic transceiver design). Lots of linearization techniques were proposed [1] and recent works made EER/polar-based architecture very popular [6]. The envelope information can be coded by a Pulse Width Modulation or a Sigma Delta ($\Sigma\Delta$) process in order to present a constant power property which is beneficial for the architecture efficiency. This coding enables its recombination by a multiplication with the carrier signal, which is modulated by the phase information. This multiplication can be achieved before the PA or by supply modulation, if the PA is in a switched (SW) class [4] such as class D, S, E or F [3].

As RF blocks are flexible, the size of the transmitter is optimized and the system efficiency is directly dependant on the different configurations. Moreover, a major design challenge is to keep a highly efficient PA for the different bands specified by the standards (carrier frequency and bandwidth). The idea of the front end design is an optimization (frequency sweeping) of a Class E PA. The Class E was chosen for its high efficiency performance and for the small number of reactive components needed, whatever the topology is [3] [7] [8] [9] [10] [11] [12] [13].

As was discussed in [2], PAs for multi-radio can be of three types: (i) broadband, (ii) reconfigurable (for example: [5]) or (iii) multi-band matching PAs. We present in this paper a simple methodology based on previous studies of Class E and optimisation processing. A slight increase in the number of reactive components can improve the bandwidth of such a PA and drive us to a dual band PA optimisation.

II. PROPOSED DESIGN METHOD

The proposed design method is summarized in figure 2. Once the given transistor and frequency bands are chosen, the idea is to first extract the non-linear (NL) output of the transistor **in switched mode of operation**, in part one of the method. In that order, we performed a transient simulation where the transistor input voltage implies a hard switching behaviour (a FET is supposed here). This hypothesis is achieved by a closed loop on the gate-to-source voltage that maintains the saturation of device whatever the frequency is, see figure 2. Thanks to this simulation, we calculated the frequency-dependant gate-to-source voltage, and the resulting input power *Pin* (unmatched case). A Large Signal S-Parameters (LSSP) frequency simulation, **including the parametric input power *Pin***, computes the LSSP output matching coefficient (*S22 NL*) of the transistor. In part two of the method, we model the real transistor as an ideal switch with its *S22 NL* (impedance) in parallel: a transistor and a capacitor, as is drawn in figure 2 and confirmed by the Smith chart. This supposes that we consider the unilaterality (*S12 = 0*).

At this point, we can compute the load facing the ideal switch, made up of the *S22 NL*, the output network (to be designed) and the antenna, as seen in Figure 2.

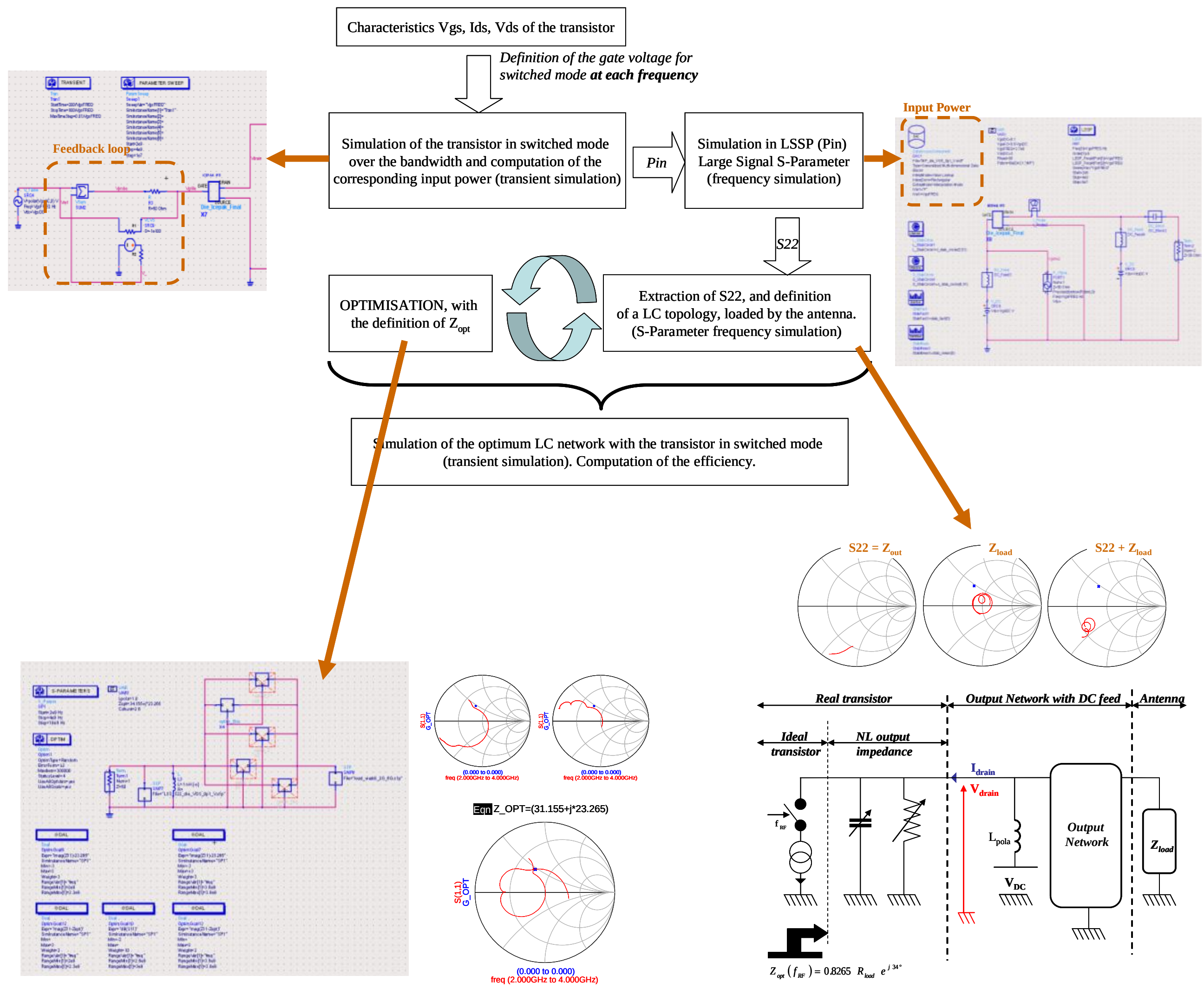

Figure 2: Proposed design method

Based on numerous theoretical developments concerning class E topologies, [13] [12] [7] [8] [9] [10] and [11], we focused on parallel-based topologies with finite DC feed inductor. The ideal load that must face the ideal transistor (the "perfect switch" part) of this type of topology [5] [11] is $Z_{opt}(\omega_{RF}) = 0.8265\ R_{load}\ e^{j34°}$, where R_{load} is the reactive part of the antenna (radiating part) and ω_{RF} the centre/carrier frequency. **The optimisation process goals are to match the impedance of the circuit (S22 NL + output network + antenna) to the optimal class E load Z_{opt}.** We compute the matching coefficient Γ_{classE} according to this criteria (50 Ohms is replaced by $Z_0 = Z_{opt} = 0.8265\ R_{load}\ e^{j34°}$). Once the optimisation process meets the goals, the network is incorporated into the final transient simulation in order to evaluate the resulting efficiency over the bandwidth.

We applied this method with the model of a MESFET Avago ATF50189 and a designed antenna (Vivaldi type). Results of the S22 NL (Z_{out}) extraction are reported in Figure 2 (left) and the antenna impedance is drawn in the left, middle smith chart. Three networks are computed: for 2 to 2.5 GHz (LTE, UMTS band), 3.2 to 3.8 GHz (WiMAX) and for both frequency bands. Figure 3 shows the results of the optimisation for the three networks, by computing the matching coefficient Γ_{classE} . We reported the result of the final transient simulation only with the network matching both frequency bands. This networks uses one DC feed inductor of 1 nH, two grounded shunt inductors of 1.62 and 2.36 nH, two series capacitors of 0.8 and 0.9 pF and two shunt capacitors of 0.4 and 1 pF. The efficiency and output power are drawn in Figure 3. We notice high theoretical drain efficiency in the range of 90%, corresponding to the goal parameters of our optimisation process. In practice, this will be lowered by the practical loss of the circuit. Maximum efficiencies agree well with matched frequency bands predicted by the S-Parameters simulation.

Drain to source voltages and currents are drawn for 2.3 and 3.5 GHz, in order to demonstrate the class E operation of the amplifier. We notice the negative current due to the importance of the output transistor capacitance (intrinsic) at these frequencies of operation.

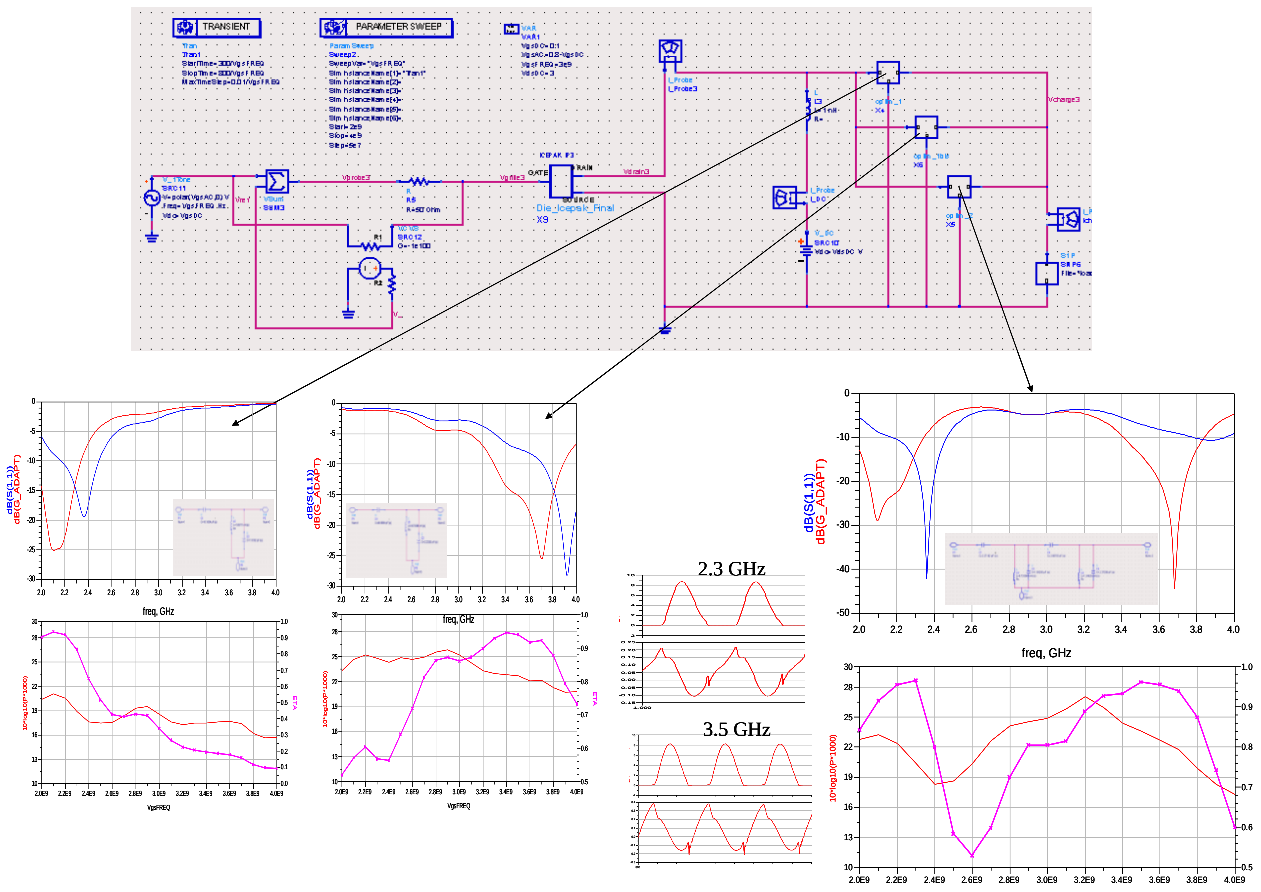

Figure 3 : Switched mode simulation of the transistor (transient) for three types of NT : (i) single band WiFi, (ii) single band WiMAX and (iii) dual band WiFi/WiMAX with transient drain to source voltage and current at 2.3 and 3.5 GHz. Efficiency (ETA in purple) and Output power (red) are reported on the same figure. Matching coefficient for 50 Ohms (blue) and Z_{opt} (red) are draw with frequency alignment.

The drain-to-source voltage is varied from 1 to 3 V and shows the independence of the efficiency and the quasi-linear variation of the output power, characterizing the saturation of the transistor. These results are shown in figure 4. It is important to evaluate this efficiency of the dual band PA over the voltage and the frequency because the multi-radio idea implies a power control whatever the signal parameters are.

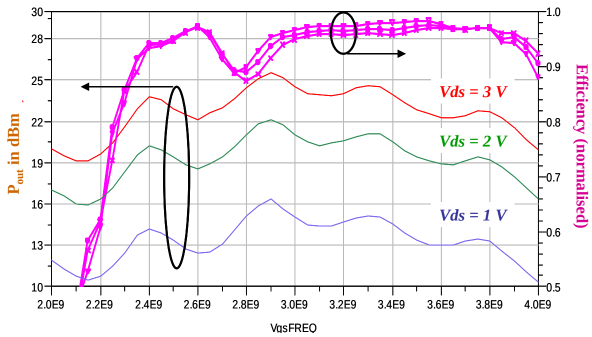

Figure 4 : Results of transient simulation with the transistor and the optimized dual-band network: efficiency and output power in function of the frequency and for Vds = 1, 2 and 3 V

III. CONCLUSION

We presented a simple methodology for class E efficiency optimisation over frequency. This method helps drive the study of broadband or multi-band PA for multi-radio. Results based on a real transistor model reach almost a one octave class E PA with reactive element values in an acceptable range for integrated design. The small number of reactive component is attractive for the design of a dual-band PA, as presented. The possibility of output power variations are simulated for slow (average power) or fast (envelope information) amplitude control. We will extend the number of frequency bands according to multi-radio front end needs.

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