Design of a Broadband and Highly Efficient 45W GaN Power Amplifier via Simplified Real Frequency Technique

David Yu-Ting Wu, Farouk Mkadem, Slim Boumaiza

University of Waterloo, Waterloo, Ontario, N2L 3G1, Canada

Abstract — **A comprehensive approach for designing broadband and highly efficient power amplifier based on optimal impedance analysis and simplified real frequency technique (SRFT) is presented. Upon determining the impedances for highest efficiency across the bandwidth of interest, the SRFT is used to obtain the optimal matching topology and element values. The effectiveness of this design technique is illustrated using a commercially available 45W GaN device which achieved an average drain efficiency of 63% from 1.9 GHz to 2.9 GHz (~42%) with an average output power and gain of 45.8 dBm and 10.8 dB respectively. The PA with DPD yielded ACPR below -50dBc when driven with WCDMA and LTE at 2.14 and 2.6 GHz respectively**

Index Terms — **Broadband amplifiers, high efficiency,**

I. Introduction

Emerging wireless standards delivering higher data rate exhibit increased signal bandwidth (up to 100 MHz in Advanced Long Term Evolution) and higher peak to average power ratio (PAPR). This has led to recent interests in wideband and highly efficient power amplifiers (PA) with the potential of becoming the main amplifier in a Doherty or Envelop Tracking configuration.

Unfortunately, high efficiency mode such as class F utilizes harmonic stubs and thus exhibits extremely narrowband characteristic. With the recent introduction of class J [1], it appears a mode of operation with high efficiency and broadband potential has been found. Yet class J ultimately dictates a set of fundamental and harmonic impedance requirements at the device intrinsic drain, which is transformed to the device package plane as the impedance requirements for the matching network. Because this transformation depends strongly on the device packaging and output capacitance, the ease of matching of the impedance contour at the package plane is not immediately apparent. In [1], the broadband output matching network was derived empirically and the design of broadband network was not discussed in detail.

In this paper, a complementary approach has been developed in which the issue of optimal broadband matching takes center stage. The proposed approach focuses on designing the minimum complexity matching network capable of realizing the optimal impedances derived from either waveform engineering or empirical loadpull. The use of waveform engineering is preferred since it determines, a priori, the optimal impedances leading to non-clipping, thus more linearizable, output waveforms [2]. Nevertheless, this paper illustrates that simple source and loadpull impedance data can yield very satisfactory broadband PA performance using the matching technique described.

Traditionally, analytical approach for determining the optimal broadband matching between an arbitrary load and a resistive termination entails complicated modeling of the impedance contour and knowledge of the matching network transfer function [3]-[4]. In the late 1970s, a mathematical algorithm known as real frequency technique was introduced [5] which bypasses the complexity of the analytical approach and directly synthesizes the optimal LC ladder using real frequency impedance data. This algorithm was further refined in what became the simplified real frequency technique (SRFT) [6]-[7]. In this paper, SRFT is adopted as part of a comprehensive design approach for highly efficient broadband power amplifier. The impedances analysis carried out in this paper also shows that the observed tradeoff between efficiency, output power, and bandwidth is in fact a manifestation of the well known gain-bandwidth constraint in broadband matching as outlined in [3]-[4].

This design uses accurate large signal model in obtaining the optimal impedances and in subsequent optimization of the matching network. The design of the 45W GaN PA based the technique described achieved a minimum drain efficiency of 60% from 1.9 GHz to 2.9 GHz and showed good linearity with digitally pre-distorted (DPD) WCDMA and LTE signal at 2.14 GHz and 2.6 GHz respectively.

II. Optimal Impedance Analysis

For a single frequency design, there exists a set of source and load fundamental and harmonic termination deemed "optimal" for a given design goal, such as maximum output power or maximum efficiency.

These optimal impedances can be determined through empirical loadpull or waveform engineering. Empirical loadpull, which blindly varies the source and load impedance until a global maximum is achieved, has the advantage of simple implementation. However, without access to the intrinsic drain waveforms, the exact cause of any non-linearity cannot be determined. It would also be nontrivial to obtain impedances away from the maximum that could achieve the desired compromise between efficiency and linearity. On the other hand, waveform engineering [1], derives the optimal impedances directly from the voltage and current waveforms at the intrinsic drain and thus afford significant insights into PA characteristics such as clipping effects, class of operation, potential sources of nonlinearity, and maximum drain to source voltage. The difficulty in implementation is in de-embedding the device package and parasitic, both of which are often kept proprietary by the device manufacturer.

Regardless of how the optimal impedances are derived, they form an impedance contour over the design frequency which can be visualized on the Smith chart. Figs. 1 and 2 illustrate the optimal fundamental and second harmonic impedance contour for the load and source respectively, extracted in simulation via source and loadpull with the device biased at V_{ds}=28V and I_{ds} = 270 mA

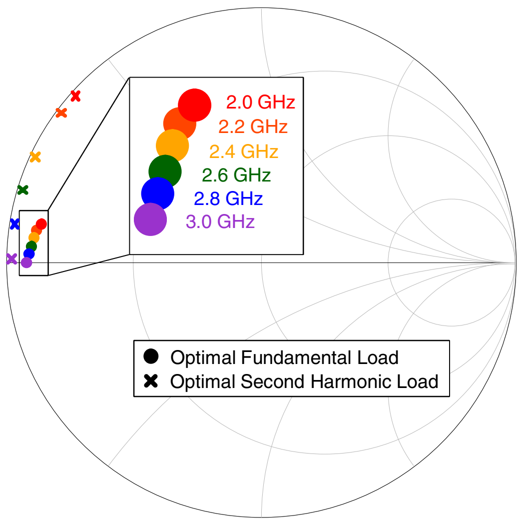

Fig. 1. Optimal fundamental and second harmonic load impedance derived from loadpull at the device package plane

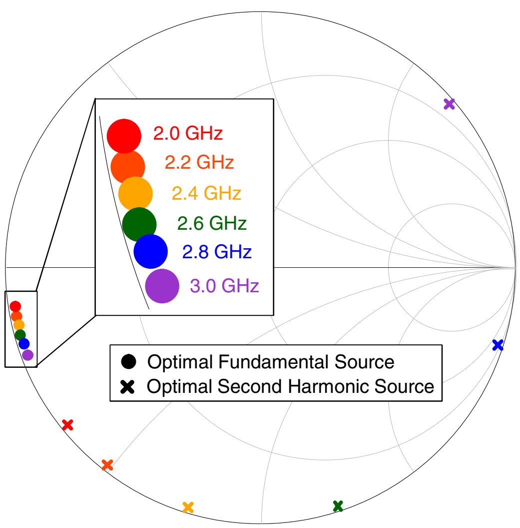

Fig. 2. Optimal fundamental and second harmonic source impedance derived from sourcepull at the device package plane

From the above figures, it appears that a matching network producing the exact optimal fundamental and harmonic impedances could theoretically yield a PA achieving maximum efficiency over the entire bandwidth. However, the *counter-clockwise* phase rotation of the contour as a function of increasing frequency reveals that such a matching network is not realizable without exotic negative capacitance and inductance. This is because LC-ladder matching network consisting of non-negative element value always produce clock-wise phase rotation (evident from plotting jωL and 1/jωC on the smith chart versus frequency).

Therefore, some impedance mismatch will be present, and the design objective is to determine the optimum network that minimizes the mismatch across the design bandwidth. A useful figure of merit indicative of the quality of match between an arbitrary load and a resistive termination, as illustrated in Fig. 3, is the transducer power gain (TPG) given in (1).

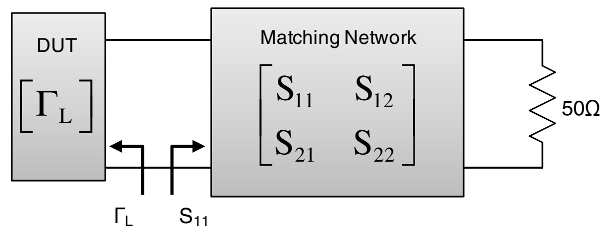

Fig. 3. Matching an arbitrary load to a resistive termination

$$T(w) = \frac{|S_{21}|^2 \left(1 - |\Gamma_L|^2\right)}{|1 - S_{11}\Gamma_L|^2} \quad (1)$$

It can be shown that high TPG and consequently better match, is more obtainable when the load impedances are closer to the center of the Smith chart and with small impedance spread versus frequency. TPG also increases with lower relative matching bandwidth. Given such, optimal harmonic matching will be difficult due to wide bandwidth and spread. Fortunately, second harmonic loadpull in Fig. 4 indicate large region of impedance tolerance in which high efficiency can be achieved across the band. Consequently, a design decision was made to consider broadband matching at the fundamental only while the harmonics are kept out of the low efficiency regions through proper placement of the bias line and subsequent optimization of the matching network.

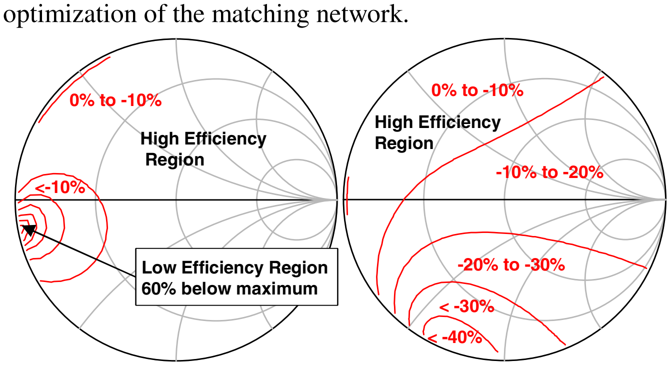

Fig. 4. Second harmonic load pull PAE contours at $2f_0$=4.0GHz (left) and $2f_0$=6.0GHz (right) indicating drain efficiency percentage below maximum in different regions of the Smith chart.

III. SIMPLIFIED REAL FREQUENCY TECHNIQUE

For a lossless, reciprocal two-port LC matching network, the S-parameter can be expressed in the Belevitch form [8] on the s-plane with polynomial *h(s)*, *g(s)*, and *f(s)* in the form

$$x(s) = x_0 + x_1 s + x_2 s^2 + \cdots + x_N s^N \qquad (2)$$

For the special case where *f(s)* = 1, the S-parameter in Belevitch form reduces to

$$S = \begin{bmatrix} S_{11} & S_{12} \\ S_{21} & S_{22} \end{bmatrix} = \frac{1}{g(s)} \begin{bmatrix} h(s) & 1 \\ 1 & -h(-s) \end{bmatrix} \qquad (3)$$

Substituting the expression for S_{11} and S_{21} in (3) into (1), and utilizing the lossless condition given by

$$g(s)g(-s) = h(s)h(-s) + 1 \qquad (4)$$

TPG in terms of *g(s)* and *h(s)* versus frequency can be derived as

$$T(\omega) = \frac{1 - |\Gamma_L|^2}{h \cdot h^*[1 + |\Gamma_L|^2] - 2\,\mathrm{Re}[\Gamma_L \cdot h \cdot g^*] + 1} \qquad (5)$$

For a given set of coefficient $\{h_0, h_1, h_2 \ldots h_N\}$ in *h(s)*, *g(s)* and TPG are uniquely defined via (4) and (5). Therefore, by utilizing nonlinear optimization routines available in commercial numerical environment, the coefficients of *h(s)* yielding the maximum TPG across the frequency band can be determined. The solution is convergent and does not depend on the initial condition because the order of nonlinearity in terms of the real coefficient of *h(s)* is always quadratic in nature [7].

Upon finding the optimal coefficients of *h(s)*, the number of elements in the LC ladder corresponds directly with the order of *h(s)*. In addition, the elements of the low-pass LC ladder network can be found by expanding the continued fraction of the driving impedance $S_{11} = h(s)/g(s)$.

By applying the SRFT to the optimal source and load impedances, LC ladder element yielding the best match is obtained. The normalized element value at ω=1, Z_0=1 is given in Table 1.

TABLE I
IDEAL LC MATCHING ELEMENTS EXTRACTED WITH SRFT

Element	Source		Load	
	Type	Value	Type	Value
Element #1	C	8.035 F	L	0.148 H
Element #2	L	0.645 H	C	8.412 F
Element #3	C	1.993 F	L	0.421 H
Element #4	L	0.907 H	C	2.659 F
Termination	R	1 Ω	R	1 Ω

To verify that SRFT yielded an acceptable match, the impedances generated by the LC ladders are overlaid with the optimal impedances in Figs 1 and 2 and shown in Fig. 5 as black traces. As expected, physically realizable networks generate clockwise phase rotation. Therefore, the optimum match involves circling around the region of optimal impedances. Each black dot represents a one of six frequency points evaluated in Figs 1 and 2. Notice that none of the six dots lie in their respective optimal. The unavoidable impedance mismatch translates into efficiency degradation, as well as degradation in gain due to input mismatch. This, along with the lack of specific harmonic termination, is the underlying cause between the tradeoff seen in narrowband and high efficiency designs, and lower efficiency and broadband designs.

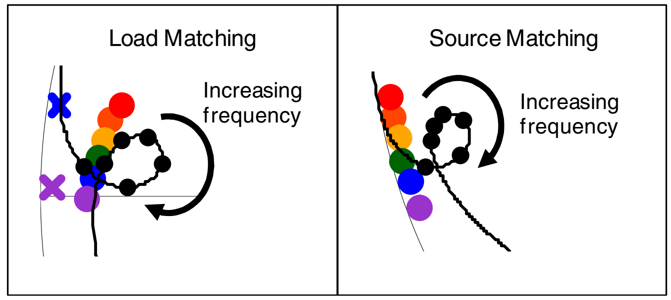

Fig. 5. SRFT generated fundamental impedances vs. optimum matching in Fig. 1 and Fig. 2

IV. MATCHING NETWORK SYNTHESIS

To realize the LC ladder in microstrip form, fairly straightforward conversion from LC to step impedance is used. Upon converting to transmission line, optimization algorithm fine tunes the line width and length to maximize efficiency. This is necessarily to take into account the parasitic introduced by the steps. Moreover, the optimization places the biasing line to ensure the harmonics stay in relatively high efficiency regions of Fig. 4. The optimization is fast and converges quickly because the initial condition was derived from the optimal LC ladder via the SRFT.

The fabricated input and output matching networks are shown in Fig. 6. As it can be seen, the network is simple and compact and required no additional tuning to obtain high performance.

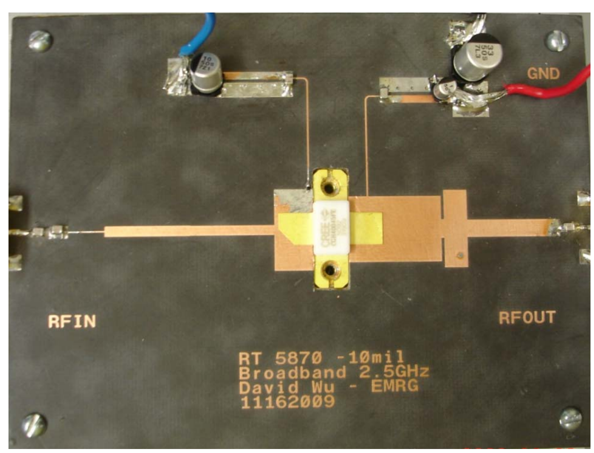

Fig. 6. Fabricated input and output matching network

V. MEASUREMENT VS. SIMULATION RESULTS

The fabricated PA without tuning underwent continuous wave (CW) stimulation from 1.8 GHz to 3.1 GHz. Fig. 7 shows the output power, gain, and drain efficiency.

The measurement matches closely versus simulation with the exception of gain which deviated by 2dB. The operating frequency also shifted below the design frequency by 0.1 GHz. The average output power, gain, and drain efficiency from 1.9 to 2.9 GHz is 45.8 dBm, 10.8 dB, and 63% respectively.

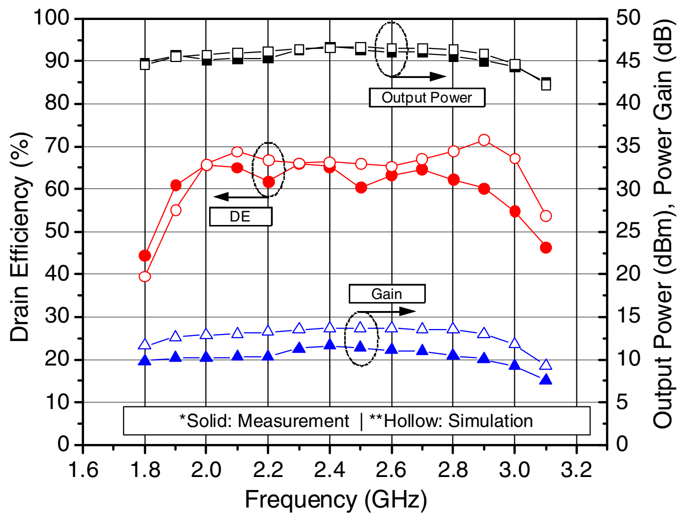

Fig. 7. Measurement versus simulation results showing drain efficiency, output power, and gain from 1.8 GHz to 3.1 GHz

VI. DPD LINEARIZATION

To assess the linearizability of the PA, two modulated signals are applied to the PA: 1001 WCDMA signal with 7.24 dB PAPR at 2.14 GHz and 10 MHz LTE signal with 9.2 dB PAPR at 2.6 GHz. Figs. 8 and 9 show the output spectrum of the PA with and without memory polynomial DPD.

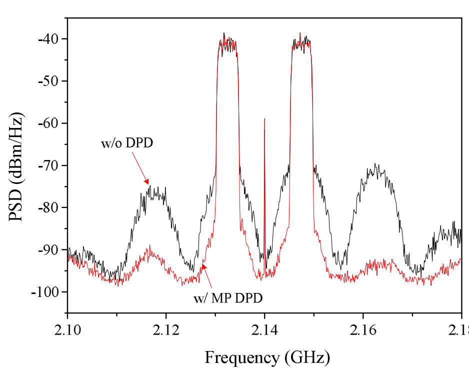

Fig. 8. Output spectrum before and after linearization for 1001 WCDMA signal at 2.14 GHz

The use of DPD significantly lowered the adjacent channel power ratio (ACPR) to below -50 dBc for both modulated signals. For WCDMA, the PA achieved an average output power of 37.78 dBm at 34.2 % drain efficiency, while the LTE signal achieved an average output power of 36.8 dBm at 27% drain efficiency.

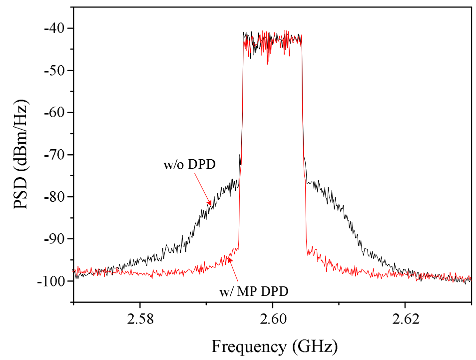

Fig. 9. Output spectrum before and after linearization for 10 MHz LTE signal at 2.6 GHz

VII. CONCLUSION

A comprehensive, first-pass design approach for broadband amplifier was presented. The analysis of the optimal impedance contour revealed fundamental tradeoffs between efficiency and bandwidth due to imperfect matching. The use of SRFT resulted in a simple and compact matching network. The fabricated PA has an average efficiency of 63%, with 10.8 dB gain and 45.8 dBm output power from 1.9 GHz to 2.9 GHz under CW stimulation. With the use of DPD, the PA was shown to be very linearizable with ACPR below -50 dBc for both WCDMA and LTE signals. The broadband PA designed is thus a good candidate for future research in broadband Doherty or Envelope tracking PA.

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