High Efficiency RF Power Amplifier Designed With Harmonic Real-Time Active Load-Pull

Xian Cui, Seok Joo Doo, Patrick Roblin, *Member, IEEE*, Jeffrey Strahler, *Member, IEEE*, and Roberto G. Rojas-Teran, *Fellow, IEEE*

Abstract—In this work, a high efficiency p-HEMT radio frequency power amplifier (PA) is designed using a new multiharmonic real-time active load-pull using the large signal network analyzer. This technique synthesizes a large set of instantaneous load mismatches to quickly find the optimal harmonic impedances, so as to achieve high PA efficiency in a shortened design cycle. At 2 GHz a demo power amplifier implemented with a p-HEMT demonstrated a power added efficiency (PAE) of 68.5% for 18.0 dBm output power, while achieving a maximum PAE of 75% below the 1 dB compression point for 18.6 dBm output power.

Index Terms—Large signal network analyzer (LSNA), power added efficiency (PAE), power amplifier (PA), real-time active load-pull (RT-ALP).

I. INTRODUCTION

NOWADAYS the availability of high efficiency radio frequency (RF) power amplifiers (PA) is a critical factor for wireless mobile devices to extend battery operation time as well as reduce the heat sink form factor and weight. Nonlinear power amplifiers, such as Class F [1], [2], achieve high efficiency by minimizing the overlap between device voltage and current to reduce the power dissipation of the transistor itself. Conventional designs using circuit simulators are limited by the accuracy of nonlinear large signal models and device package parasitics [3]. Passive or active load-pull [4]–[6] can be used to determine the optimal load conditions for the power transistor but it is a time-consuming process, especially in nonlinear PA design where both the impedance amplitude and phase must be swept for each harmonic to find the optimal terminations.

A real-time active load-pull (RT-ALP) [7] based on the large signal network analyzer (LSNA) was introduced for ultra-fast impedance sweeping. It was extended to multiharmonics for nonlinear PA design in [8] and multiharmonic load conditions for CW operation of a GaN HEMT with PAE of 80% were extracted. This work indicated that RT-ALP offers great opportunities for speeding up the PA design.

This letter presents the first complete design of a 2 GHz high efficiency p-HEMT PA using the new multiharmonic RT-ALP for drastically shortening the PA design cycle. Real-time tunings not only at the fundamental but also the second and third harmonic frequencies are utilized to quasi-interactively determine

Fig. 1. Frequency domain representation of the incident wave a_2 actively injected (blue) at the transistor's output in RT-ALP measurements.

Fig. 2. Multiharmonic real-time active load-pull configuration.

the optimal load terminations. The goal is to select a load-line minimizing $\int_0^T i_D(t) \times v_{DS}(t) dt$ instead of $i_D(t) \times v_{DS}(t)$ for a given input power. The corresponding matching networks are then designed to realize the optimal multiharmonic reflection coefficients obtained. The constructed p-HEMT PA is then characterized and its performances are compared to the RT-ALP prediction.

II. MULTIHARMONIC RT-ALP TECHNIQUE

At the n -th harmonic frequency ($n\omega_0$), the RT-ALP method applies at the output a CW tone $a_2(n\omega_0 + \Delta\omega)$ with a frequency offset $\Delta\omega$ (Fig. 1) from the targeted $n\omega_0$ so that the phase of the harmonic load reflection coefficient is continuously swept in each LSNA measurement (e.g., 10 ms). The reflection coefficient loci is changed by varying the power levels for $a_2(n\omega_0 + \Delta\omega)$, producing a set of harmonic load reflection coefficient loci mapping the Smith chart (see Fig. 3 in Section III). Circles with larger radius are associated with larger injected power levels. The swept dynamic RF loadlines of the device are simultaneously obtained.

The configuration of the multiharmonic RT-ALP is shown in Fig. 2. The top RF source provides a CW signal at the fundamental frequency ω_0 to drive the transistor input. The two bottom sources inject signals at $(2\omega_0 + \Delta\omega, 3\omega_0)$ or

Manuscript received October 17, 2007; revised January 8, 2008. This work was supported in part by the National Science Foundation under Grant ECS-0622003.

The authors are with the Department of Electrical and Computer Engineering, The Ohio State University, Columbus, OH 43210 USA (e-mail: roblin@ece.osu).

Color versions of one or more of the figures in this letter are available online at <http://ieeexplore.ieee.org>.

Digital Object Identifier 10.1109/LMWC.2008.918913

Fig. 3. Loci of $\Gamma_L(2\omega_0, t)$ obtained from the $(2\omega_0 + \Delta\omega)$ real-time swept phase measurements by accounting for the two dominant tones (solid line) or all the sideband tones (dashed line) with (1). Each circle is measured in 10 ms.

$(2\omega_0, 3\omega_0 + \Delta\omega)$ in the RT-ALP case (swept phase measurement) or $(2\omega_0, 3\omega_0)$ in the constant phase measurement case. The bias voltage and current variations of the device are displayed on the oscilloscope. The diplexer provides a path for the injected harmonic signals to reach the transistor output while maintaining an impedance match for ω_0 at port 2.

The load reflection coefficient $\Gamma_L(n\omega_0, t)$ can be easily extracted from the RT-ALP data by either accounting for only the two dominant tones ($n\omega_0$) and $(n\omega_0 + \Delta\omega)$, or all the sideband tones with *SSB* the total number of single side band modulation tones measured by the LSNA using

$$\Gamma_L(n\omega_0, t) = \frac{\sum_{p=-SSB}^{SSB} a_2(n\omega_0 + p\Delta\omega)e^{jp\Delta\omega t}}{\sum_{p=-SSB}^{SSB} b_2(n\omega_0 + p\Delta\omega)e^{jp\Delta\omega t}}. \quad (1)$$

The contours of output power and PAE can then also be plotted to select the optimal harmonic load conditions.

III. PA DESIGN WITH RT-ALP AND MEASUREMENT RESULTS

The above multiharmonic RT-ALP methodology is applied to a packaged pHEMT (Agilent ATF-54143) biased around cut-off ($V_{DS} = 30$ V, $V_{GS} = 0.31$ V). In this study the LSNA acquired spectral data up to the fourth harmonic each with 99 tones ($SSB = 49$). The modulation frequency Δf is 200 kHz.

With a RT-ALP first run at $f_0 = 2$ GHz with 6.5 dBm input power, the optimal load at the fundamental is selected as $\Gamma_L(\omega_0, t) = -0.2$ for the subsequent $2f_0$ and $3f_0$ RT-ALP.

Fig. 3 shows the extracted $\Gamma_L(2\omega_0, t)$ result from the second harmonic RT-ALP data at 4 GHz. The dashed line loci are obtained when accounting for all sideband tones in (1), and the solid line loci when considering only the two dominant tones. At low power levels (small radii) the latter is a good approximation to the former as expected. The PAE contours are shown in Fig. 4. As can be seen, the realistic second harmonic impedance with the highest PAE (case A) near the Smith chart's edge differs from the class B/F ideal short. This high PAE is verified with the constant phase measurement to be 71.92% with an output power of 60.8 mW.

Fig. 4. PAE contour plot in the $\Gamma_L(2\omega_0)$ plane, obtained from the $(2\omega_0 + \Delta\omega)$ real-time swept phase measurements (based on dashed line contour in Fig. 3).

TABLE I
COMPARISON OF THREE TYPICAL CASES: CASE A, B, C

$\Gamma_L(2\omega_0)$	PAE (%)		Output Power (mW)	
	Swept Phase	Constant Phase	Swept Phase	Constant Phase
Case A : (optimal $0.6 + j0.7$)	77.04	71.92	60.26	60.80
Case B: (short)	62.70	63.48	52.10	51.40
Case C: (50 Ohm)	60.44	61.90	46.93	47.10

Note: The transistor's input power remains as 6.5 dBm at 2 GHz.

Fig. 5. Output matching network to supply the optimal reflection coefficients predicted by RT-ALP: (a) S parameter at f_0 , $2f_0$ and $3f_0$ frequency (marker 1, 2, and 3, respectively) and (b) layout (left end to transistor and right end to 50 Ω).

Table I lists the three typical cases (A, B, C) of the PAE contours in Fig. 4 selected for comparison. In each case both the PAE and output power results from the RT-ALP are consistent with those from the constant phase measurement. The small difference indicates that memory effects are small for the 200 kHz RT-ALP modulation frequency used.

With the RT-ALP tuning at $3f_0$, it was found that the performance improvement from controlling the third harmonic is almost insignificant for this particular device. The resulting optimal load terminations for the output matching are then: $\Gamma_L(\omega_0, t) = -0.2$, $\Gamma_L(2\omega_0, t) = 0.6 + j0.7$, $\Gamma_L(3\omega_0, t) = 0.30 - j0.85$.

Fig. 5 shows the designed output matching network with microstrip substrate of Duroid ($\epsilon_r = 2.2$, $H = 31$ mil). The source

Fig. 6. Simplified schematic of the designed pHEMT PA including bias tees.

Fig. 7. RF load lines of the transistor for swept phase (blue) with the optimal one (green) and the final constant phase load line (red).

Fig. 8. (a) Drain current and voltage waveforms and (b) instantaneous power dissipation at the output: the efficiency is optimized when the difference between positive and negative areas is the smallest fraction of the DC power dissipated.

matching network is a conjugate match of the input impedance. The fabricated amplifier with bias tees (Fig. 6) is measured with the reference planes de-embedded to the planes A and B given in Fig. 6. Under the input power of 6.5 dBm at 2 GHz, the PA achieves PAE of 68.5% with 64.1 mW fundamental output power and 15.9 dB gain. The PAE and Pout results are very close to the findings in Table I, revealing the reliability of the RT-ALP design methodology.

The associated RF loadline is shown in Fig. 7. Note that this is the orthogonality of the time-domain drain voltage and current waveforms in Fig. 8(a) with power generation and absorption in Fig. 8(b) balancing which yields the high PAE.

Fig. 9. Pout, Gain, and PAE versus the input power sweep at 2 GHz.

From the power sweep measurement in Fig. 9, we can observe a PAE of 68.5% with 18 dBm output power and 16 dB gain at 2 GHz. Below the 1dB compression point, this PA achieves a PAE up to 74.6% with 18.6 dBm output power.

IV. CONCLUSION

We have presented the design of a high efficiency p-HEMT RF PA using a novel multiharmonic RT-ALP. The methodology enables to efficiently investigate the optimal load mismatches at the fundamental and harmonics to achieve high PAE in a shortened design cycle. A multiharmonic matching network was then designed to realize the p-HEMT PA. The measurement result of 68.5% PAE at 2 GHz is in good agreement (around 3% difference) with the RT-ALP prediction. This demonstrates the efficacy of real-time tuning for the interactive design of power efficient PAs.

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