Active Harmonic Load-Pull With Realistic Wideband Communications Signals

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Abstract *— A new wideband open-loop active harmonic load–pull measurement approach is presented. The proposed method is based on wideband data-acquisition and wideband signal-injection of the incident and device generated power waves at the frequencies of interest. The system provides full, user defined, in-band control of the source and load reflection coefficients presented to the device-under-test at baseband, fundamental and harmonic frequencies. The system's capability to completely eliminate electrical delay allows it to mimic realistic matching networks using their measured or simulated frequency response. This feature enables active devices to be evaluated for their actual in-circuit behavior, even on wafer. Moreover the proposed setup provides the unique feature of handling realistic wideband communication signals like multi-carrier wideband code division multiple access (W-CDMA), making the setup perfectly suited for studying device performance in terms of efficiency, linearity and memory effects.*

In this work we describe the hardware and signal conditioning of the proposed setup. The high dynamic range, bandwidth and measurement speed of the system, together with its capability to engineer the large-signal operation of an active device, are demonstrated by measuring the improved RF performance of a multi-carrier W-CDMA driven laterally diffused metal–oxide–semiconductor device when the electrical delay in the setup is canceled.

Index Terms—Complex modulated signals, device characterization, large-signal characterization, laterally diffused metal–oxide–semiconductor (LDMOS), load–pull, mixed signal, wideband.

This article was first published November 18, 2008; republished 12 December 2008 in *IEEE Transactions on Microwave Theory and Techniques, Vol. 56, No. 12*; December 2008. Republished in this format 17 June 2010, with permission.

Introduction

LOAD–PULL techniques have been introduced in the past to characterize active devices for their large-signal properties when small-signal linear superposition techniques are no longer valid [1]. Originally, load–pull technique was targeted on finding the optimum matching condition, at the fundamental frequency, in order to achieve the highest gain or output power of the active device-under-test (DUT). Currently, with the growing importance of wireless applications and the increasing use of high data-rate protocols, the linearity and efficiency of the amplifier have become the dominant parameters to optimize. To serve this optimization in the best possible way, control of the higher harmonics and baseband (BB) terminations is required [2], [3]. Several works have been reported in literature providing extended capabilities for passive and active load–pull systems, to include harmonic and BB impedance control, yielding numerous innovations in device characterization [4]–[14]. Nevertheless, current load–pull solutions still suffer from various limitations, which we will discuss briefly below.

Passive harmonic load–pull systems [4], [5] [see Figure 1(a)] are currently still the preferred industry large-signal test bench choice due to their simplicity and high power handling capabilities. Unfortunately, passive load–pull setups are constrained by losses and electrical delay in the tuners, connecting cables and wafer probes, when on die measurements are performed. This limits the magnitude of the reflection coefficients that these systems can provide to the DUT (note that the reflection is reduced by twice the path insertion losses). These restrictions are most problematic when considering the harmonic termination requirements for the higher efficiency classes of operation (e.g., class-AB, class-D, and class-F). Here, the use of a shorted or open condition is often required for the harmonic terminations in order to achieve maximum efficiency or linearity. In addition to this, the electrical delay, introduced by the tuners and connecting cables, will generate a variation in the phase of the reflection coefficients offered to the DUT as function of frequency. This results in a varying in-band reflection coefficient under wideband modulated signal excitation. Since this delay is significantly larger in a (passive) load–pull setup than in a real circuit, the resulting phase change is unrealistically high and restricts the wideband properties of the measurement setup (typical numbers are 3°/MHz). If high-resonators or impedance tuners are used to control the harmonic terminations, the phase change with frequency is even worse. For this reason passive harmonic tuner systems have their most meaningful use in narrowband device characterization.

Active load–pull systems can, thanks to the use of amplifiers, overcome the reflection magnitude restrictions due to losses. In literature two basic concepts for active load–pull have been promoted over time, which we briefly discuss below.

1) Closed-Loop Active Load–Pull Systems: In these systems [6]–[9], to achieve the desired reflection coefficient, a portion of the wave generated by the DUT (e.g., b_1 or b_2) is coupled out of the signal path, adjusted in amplitude and phase and then injected back as an incident wave to the active device [see Figure 1(b)]. Since the response time of the loop is much smaller than the period of the signal envelope, closed-loop systems can provide real-time the desired reflection coefficients to the DUT, independent of the power and spectral content of the signal. This makes the closed-loop concept suitable for fast device characterization, but also prone to oscillations since the loop gain cannot be very selectively controlled over frequency. To avoid the risk of oscillation, most closed-loop systems include sophisticated filtering [6], [9]. To relax the oscillation problem, a recent work proposes to control the reflection coefficient at BB frequencies [8], where it is possible to implement better frequency selectivity.

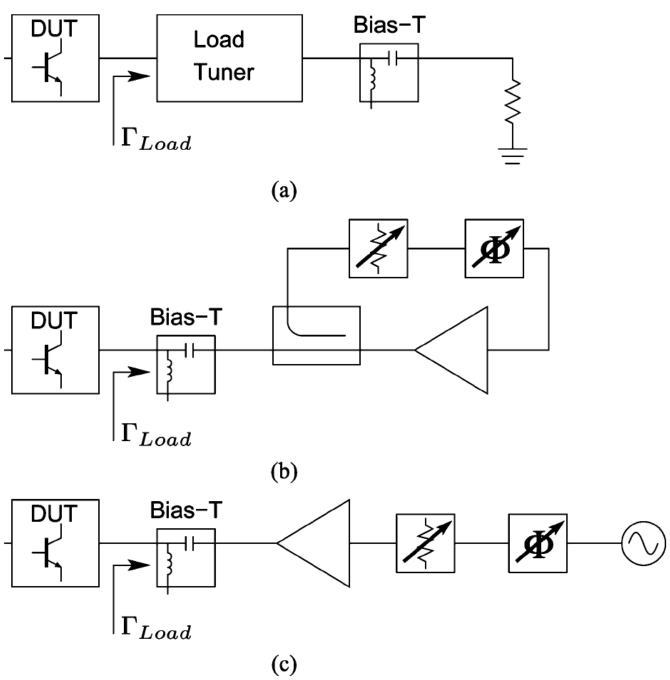

Figure 1. (a) Typical passive load–pull configuration, the bias-tee placed after the tuner reduces the impact of its losses, but yields larger memory effects due to BB inductance. (b) Closed-loop active load–pull configuration, here the bias-tee is placed as close as possible to the DUT, to reduce unwanted memory effects. (c) Open-loop active load–pull configuration, also here the bias-tee is placed as close as possible to the DUT, to reduce memory effects caused by bias path inductance.

In spite of the significant progress made in closed-loop systems, there are still two basic constraints, namely: the electrical delay of the active loops and their connecting cables (30°/MHz for a commercial system and 4.85 MHz for the custom optimized setup in [9]), as well as the limited linearity of the loop amplifiers. This latter problem requires the use of extreme power back-off conditions for the loop amplifiers in order to achieve the desired linearity. These two basic limitations disqualify closed-loop systems for the linearity characterization of devices with output power greater than 10 W and bandwidths larger than 10 MHz.

2) Open-Loop Active Load–Pull Systems: These systems [10]–[14] do not reuse the waves generated by the DUT, but directly inject the synthesized signals into the DUT to compose a desired reflection coefficient [see Figure 1(c)]. In their most basic implementation they are strictly narrowband. Since the injected signals are no longer directly dependent on the waves generated by the DUT, the realized reflection coefficients are power and phase dependent. Therefore iterations are needed during the measurements to find the optimal injection signals to offer the desired reflection coefficients to the DUT. This makes these systems slower than the closed-loop ones.

To control the phase and amplitude of the injected signals, early versions employed attenuators and phase shifters. More advanced versions include in-phase/quadrature (IQ) modulators to control the amplitude and phase relations [13]. When extending this single-tone principle to wideband signals the difficulty arises on how to determine the required content of the injection signals, at both the fundamental and harmonic frequencies. To tackle the problem, the use of multipliers in the signal path has been proposed [14]. However, when considering realistic devices, intermodulation distortion and memory effects will be present, making it impractical to obtain the required injection signals by a separate analogue multiplication process.

In this paper, we present a new open-loop load–pull measurement system with high dynamic range that solves for losses, electrical delay, power handling and

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linearity limitations present in current passive and active load–pull systems. The proposed method is based on wideband data-acquisition and wideband signal-injection of the incident and device generated power waves at the frequencies of interest. By monitoring and controlling the spectral content of these waves in the fundamental and harmonic frequency bands, it is possible to synthesize user defined reflection coefficients at the DUT reference planes. This allows users to mimic any matching network based on its frequency scattering parameters, measured or even simulated. Furthermore, by monitoring and controlling the spectral content of the waves at the DUT reference planes, the signal distortion resulting from the "loop" amplifier nonlinearities will be automatically compensated. For this reason, the system is able to perform high-linearity measurements up to the saturated power of the "loop" amplifiers used in the system. This should be regarded as a major advantage over closed-loop topologies. In conclusion, the elimination of both electrical delay and amplifier nonlinearities, opens the door for very advanced large-signal device testing over extreme wide bands. Something that is highly desirable for the development of the next generation of wireless applications.

Wideband Open-Loop Load–Pull Approach

As previously mentioned, we aim to identify a load–pull measurement approach capable of providing large signal device characterization with absolute control of the source and load reflection coefficients versus frequency in the BB, fundamental and harmonic frequency bands (bands centered at f_1, $2f_1$, etc.). Since reflection coefficients represent the ratio of waves, we basically want to control the linear ratios of the incident and device generated waves on the DUT (Figure 2).

The only wave known prior to the iteration procedure, with its (potentially) complex modulated content, is the source signal (a_s). The device generated waves (b-waves), are the result of the interaction of the a_s wave with the nonlinear properties (unknown) of the device. It is important to note that the spectral content of the generated waves will include not only the fundamental frequency band, but also the BB, higher order harmonics, and intermodulation distortion. However, knowing the (user defined) reflection coefficients and being able to accurately measure the spectral content of the device generated waves, it is possible to derive what the spectral content of all the injected waves should be. The waves that need to be generated and injected into the DUT, to provide the required reflection coefficient, can be linked to the device generated signals by the following equation:

$$a_{x,n}(f_n) = b_{x,n}(f_n) \cdot \Gamma_{x,n}(f_n) \tag{1}$$

where x is an index to indicate source (s) or load (l), n indicates the frequency band, e.g., BB (0), fundamental (1) and harmonic (2 and up), and $\Gamma_{x,n}(f_n)$ is the user defined reflection coefficient of the source or load versus frequency at the BB, fundamental, or harmonic frequency.

Although conceptually simple, there are a few issues to be solved in this approach. First of all this method requires very-fast data acquisition with high linearity and high dynamic range in order to handle the spectral content of the complex modulated signals with their related distortion products. Secondly, the a-waves need to be generated with a high dynamic range and must be optimized for their spectral content in order to satisfy (1). Both requirements place high demands on the capabilities of the hardware configuration as well as on the related software.

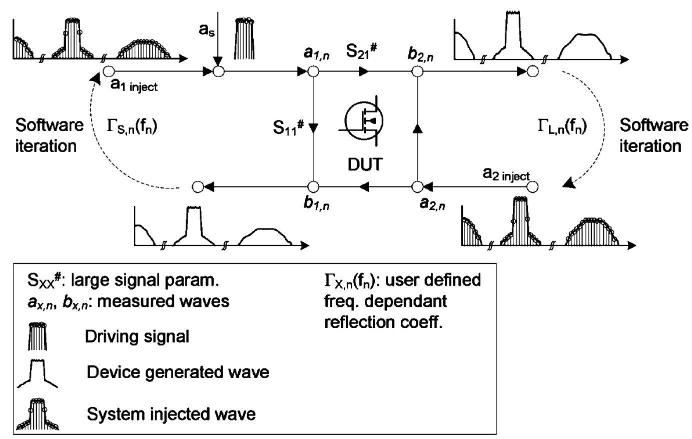

Figure 2. Principle of the proposed wideband open-loop active load–pull approach. When the nonlinear DUT is excited with a user-defined modulated signal a_s, it generates signals in the BB, fundamental and higher harmonic frequency bands. By measuring the device generated waves ($b_{1,n}$ and $b_{2,n}$) as well as the incident waves, the waves to be injected are estimated at every iteration. When the required reflection coefficient versus frequency (at every controlled band) is achieved, the iteration has converged and the large signal parameters (e.g., power added efficiency, output power, intermodulation distortion, etc.) are measured.

Hardware Description

A simplified block diagram of the measurement setup is shown in Figure 3. The S-parameter test-set is based on the five-coupler configuration, first described in [15], that allows the simultaneous measurements of the source, input and load reflection coefficients at the DUT reference planes. Wideband A/D converters (100 MHz sampling frequency) are used to acquire the down-converted waveforms, facilitating the measurement of the device reflection coefficients over a wide bandwidth in a single data acquisition.

Custom bias-tees with low inductance are placed directly at the wafer probe in order to minimize the electrical delay of the BB impedance, implemented here as a passive impedance switch bank [16]. On the BB board also the low-frequency test-set for the calibrated BB impedance measurement is implemented.

The fundamental and harmonic loads are synthesized by injection of arbitrary waveforms as described in detail in the following paragraph.

A. Signal Generation

The source signal and all injection signals needed to create the user defined reflection coefficients at the DUT reference planes, are originating from fully synchronized (200 MS/s) arbitrary waveform generators (AWGs), which share the same time base. Based on the phase coherency requirement between fundamental and harmonic injection signals, IQ up-conversion is preferred over digital IF techniques [17]. This allows to use a single local oscillator (LO) to generate the high frequency signals at fundamental and harmonic frequencies by means of multipliers (e.g., ×2 for the second harmonic) in the LO path (Figure 4). This guarantees that the active loads and the driving signal are phase coherent since

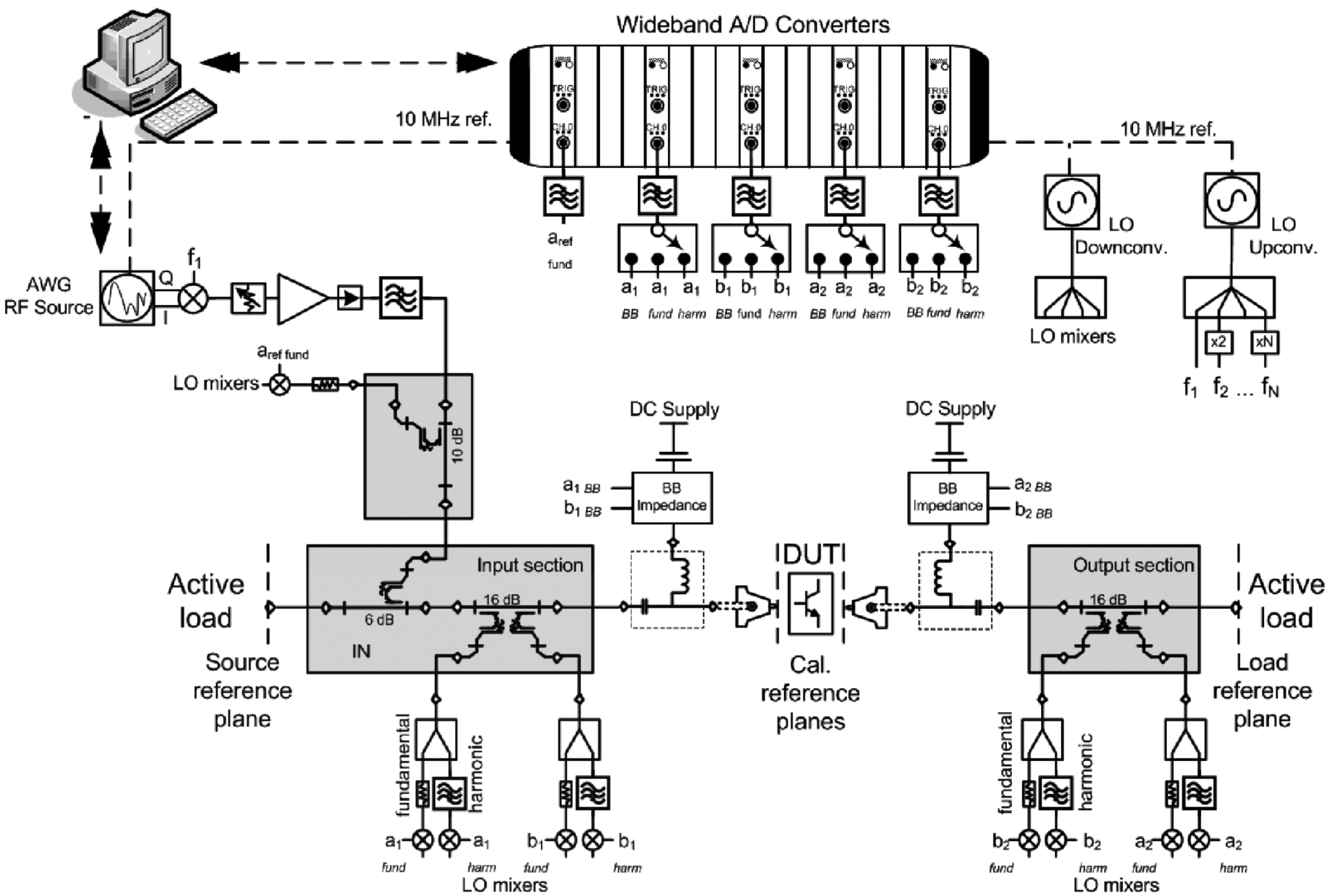

Figure 3. Simplified block diagram of the proposed wideband active harmonic open-loop load-pull system.

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this LO does not need to be swept. Consequently, the source and all injection signals are up-converted to the fundamental and harmonic frequencies and fed to the DUT to establish the driving signal and reflection coefficients.

Another advantage of the IQ approach, compared to other known signal generation techniques, is the relative limited length of the data records needed to fulfill the standard model requirements of complex modulated signals (e.g., wideband code division multiple access (W-CDMA) [20]), yielding a significant speed advantage in practical measurement situations.

Finally, computer controlled attenuators and high power amplifiers are placed in the signal path, after the IQ up-converters, in order to level the power of the injection signals. This allows to make full use of the maximum dynamic range of the AWGs at all times, something that proves to be essential for meeting the spectral requirements of modern communication signals. With the current configuration the system is able to deliver up to 2 W of average power at the input reference plane of the DUT with a linearity that is sufficient for the communication standards currently used in industry.

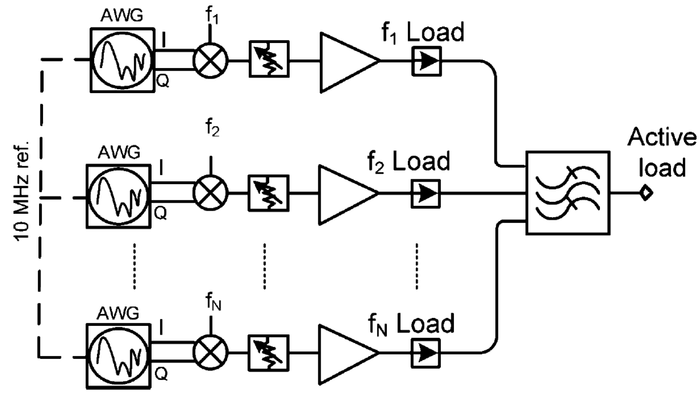

Figure 4. Simplified block diagram of the wideband active loads with phase coherent frequency up-conversion.

B. Data Acquisition

To obtain an accurate representation of the spectral content of the RF waves at fundamental and harmonic frequencies, the measured power waves need to be down-converted to a lower frequency prior to data acquisition. This is a common technique used in vector network analyzers and permits to achieve the highest possible dynamic range. However, to do this correctly for modulated signals, it is very important that the detection path is free from nonlinear errors that cannot be corrected for by linear calibration techniques. To maximize the detection dynamic range of the harmonic frequency components, power splitters and high-pass filters are used at the detection ports of the couplers in the input and output sections, as shown in Figure 3 [9]. By high-pass filtering the higher harmonic components, the mixer used for the down-conversion in this signal path is protected from the (high) power of the fundamental signal. This drastically relaxes the mixer linearity requirements for the second and higher harmonics, improving the quality of the acquired signals. By stepping the LO frequency, the frequency band centered on the harmonic of interest is down-converted to its low IF representation for data acquisition. The detection dynamic range of the system in the fundamental frequency band is maximized by using variable attenuators in the RF path and high-power high-linearity mixers. Note that DUT nonlinearities also give rise to BB signals, which do not need frequency conversion prior to the data-acquisition. The resulting low IF signals and the BB signals from the DUT are fed to multiplexing high-speed switches to reduce the number of high-performance wideband A/D converters needed.

C. System Calibration

The system calibration is relatively straightforward and is a combination of the techniques described in [15], [16], and [18]. A first calibration step uses a combination of standards at the source and DUT input reference planes, and allows the simultaneous measurement of the source and DUT input reflection coefficients [15]. At the same time a calibration for the measurement of the BB impedance is performed by use of a “short”, “open” and “load” standard at the DUT input and output reference planes [16]. A second calibration step employs a “short”, “open” and “load” standard at the load reference plane, when a “thru” is used as a DUT, and it allows the measurement of the DUT load reflection coefficient. Finally an absolute power calibration is done by use of a power meter at the load reference plane [18].

An optional calibration step can also be performed to correct for the IQ modulators imperfections. During this step the IQ modulators leakage is minimized by use of dc offsets while balance and quadrature errors are corrected by use of digital pre-compensation [19].

Signal Processing

A. Generation and Data Acquisition of the Waves at the DUT

1) Generation: When testing the performance of active devices for complex modulated signals (e.g., W-CDMA) the driving signal must comply with all related specifications of the standard test model [20]. This requires the use of a predefined data record length for the AWGs, which must be sufficiently long to store at least one full period of the waveform specified by the standard test signal model. Since in our approach the reflection coefficients are synthesized by signal injection, the injected signals must be fully coherent with the driving signal and should therefore have the same record length. The required record length in combination with the sampling speed sets an effective frequency bin size (Δf_{AWG}) for the signals preseint in the system

$$\Delta f_{\mathrm{AWG}} = \frac{fs_{\mathrm{AWG}}}{N_{\mathrm{AWG}}} = \frac{1}{T_{\mathrm{MOD}}} \tag{2}$$

where Δf_{AWG} represents the frequency bin size of the generated signals, and fs_{AWG} and N_{AWG} are respectively the sampling frequency and the number of samples used by the AWGs, and T_{MOD} is the time period of the source signal that is needed to meet the requirements of the modulation standard according to the given test model.

An example of the signals present in the proposed system is given in Figure 5 which illustrates the frequency binned spectral content of the I and Q signals to be delivered to the IQ modulator [see Figure 5(a)]. This block generates the RF source signal which drives the DUT with a given modulation (e.g., W-CDMA). Due to the always present nonlinearities of the active DUT, the DUT generated waves ($b_{1,fund}$ and $b_{2,fund}$) will contain besides the desired fundamental signal, also intermodulation sidebands. Moreover spectral content generated by nonlinearities will also be present in the BB and harmonic frequency bands (see Figure 2). When considering fundamental operation (same reasoning applies also for the harmonic frequency bands), the down-converted RF signal with intermodulation sidebands is given in Figure 5(b). In order to realize the desired reflection coefficients over the total bandwidth where spectral content is present, the I and Q injection signals must now include the third- and fifth-order intermodulation distortion (IM3 and IM5) sidebands [see Figure 5(c)]. Failing to provide the proper signal at the IM3 and IM5 frequency bands, would create an unrealistic 50Ω termination for those DUT generated signals, which totally invalidates any linearity performance measurement.

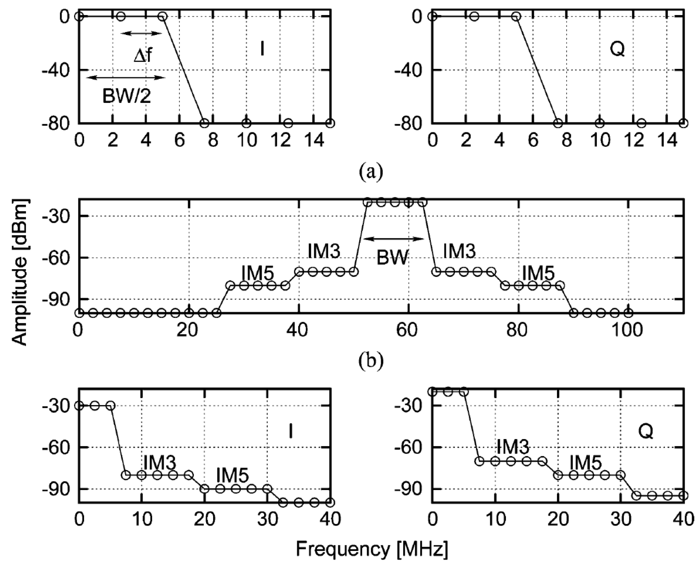

Figure 5. Explanation of the signals present in the proposed load–pull system. (a) Frequency binned spectral content of the I and Q waveforms for generating the drive signal of the DUT. (b) Down-converted low IF representation of the spectrum in the fundamental band at the output of the DUT. (c) Spectral content of the I and Q waveforms for generating the active load injection signal to achieve the user defined reflection coefficient over the fundamental band.

2) Data Acquisition: As previously discussed, to achieve a functional wideband load–pull system, all frequency bands of interest need to be measured with high accuracy and frequency resolution. Wideband analog-to-digital (A/D) converters are used to acquire the down-converted signals, and an FFT is applied to obtain their spectral content. Also in this case the sampling speed of the A/D converter and the time span used for the data-acquisition set an effective frequency bin size which must be compatible with the applied test signal, as described by the following expression:

$$\Delta f_{\mathrm{A/D}} = \frac{fs_{\mathrm{A/D}}}{N_{\mathrm{A/D}}} = \frac{\Delta f_{\mathrm{AWG}}}{k} = \frac{1}{k \cdot T_{\mathrm{MOD}}} \tag{3}$$

where, in addition to the previously introduced quantities, $\Delta f_{\mathrm{A/D}}$ is the frequency bin size of the acquired signals, and $fs_{\mathrm{A/D}}$ and $N_{\mathrm{A/D}}$ are respectively the sampling frequency and the number of samples used by the A/D converters, and k is an integer.

For a correct operation of this approach, the frequency resolution of the A/D converter should be set equal ($k = 1$), or an integer factor better (smaller frequency bin size) than that of the generated signals.

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With the spectral content of the incident and device generated waves at the DUT reference planes available, we are able to calculate the actual measured reflection coefficients versus frequency at the fundamental and harmonic frequencies.

B. Optimization of the Reflection Coefficients

With the principles of signal injection and acquisition of the waves at the DUT reference planes explained, we still need to find the exact waveforms to be loaded in the AWGs in order to achieve the desired (user defined) reflection coefficients. As discussed in Section II, only the content of the driving waveform (a_s) is known prior to the acquisition. All other injection signals ($a_{1\text{inject},n}$ and $a_{2\text{inject},n}$) need to be found by optimization. By monitoring the deviation of the measured reflection coefficient with the actual desired one for each frequency bin, the spectrum of the injected wave is optimized and found by subsequent iterations as shown in Figure 6. The definition, error checking and optimization are done in the frequency domain, while the actual injection signals are loaded and acquired in the time domain.

The open loop approach guarantees that no sustained oscillations can occur. In practice when the target, user-defined, reflection coefficients (at its input or output) force the device to operate in an instable region, the system will simply fail to converge to its end solution. In all other situations normal convergence of the optimization algorithm will occur.

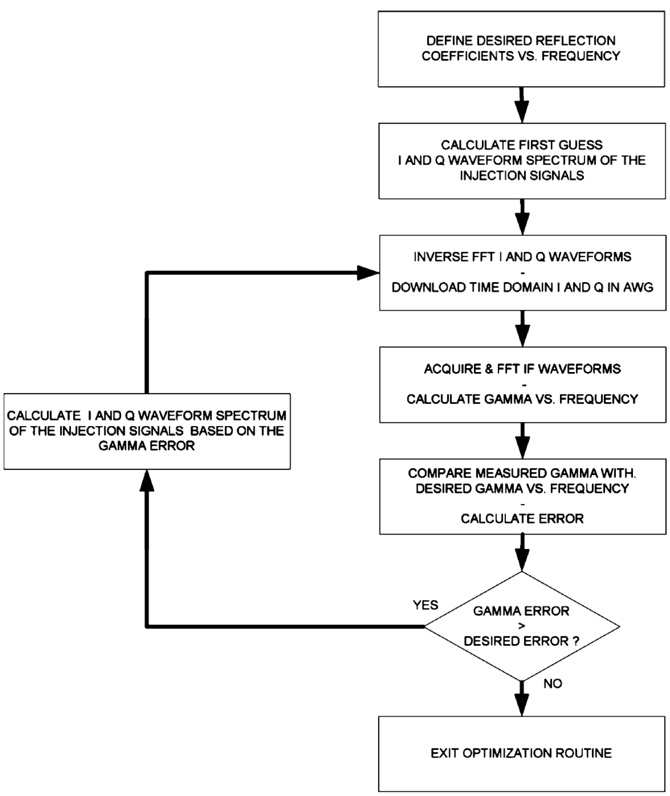

Figure 6. Flow diagram of the optimization for the reflection coefficients.

The use of the combined IQ up-conversion and low IF down-conversion requires frequency mapping and IQ decomposition, which must be handled by the controlling software. If there is a difference in frequency bin size between acquisition and signal generation [see (3)], additional action is required by the software to match the effective frequency bin size.

C. System Linearity Considerations

As mentioned in Section I, the open-loop approach will lower the linearity requirements of the drive amplifiers used for the signal injection. This can be understood as follows: the waveforms of the AWGs are optimized so that they exactly fulfill, in combination with the device generated waves, the user defined reflection coefficients at the DUT reference planes, over the frequency-band of interest. Since the reflection coefficient is just a linear relation between the incident and reflected wave, potential distortion products resulting from the open loop amplifier will be automatically compensated during the iteration of the reflection coefficients versus frequency.

System Performance

A. System Operating Frequency and Bandwidth

In principle the system operating frequency has no strict limitation, since it is only bounded by the frequency handling capabilities of the IQ modulators, test amplifiers and multipliers, which are commercially widely available up to K-band. The fundamental operating frequency of our current system implementation can be chosen arbitrarily within an octave bandwidth (1.5–3 GHz), without requiring any hardware replacement. When the diplexers in the system are replaced, a wider frequency range can be covered.

When considering the modulation bandwidth, two different bandwidths should be considered, namely the signal generation bandwidth and the signal detection bandwidth. The maximum analogue frequency of the AWGs sets the upper limit for the input signal as well as the bandwidth over which the reflection coefficients can be controlled. In the proposed setup the AWGs can produce signals up to 80 MHz, which in IQ up-conversion results in a maximum controllable bandwidth of 160 MHz. The A/D converters were selected for maximum dynamic range and therefore have a slightly lower analogue bandwidth, namely, 45 MHz. Nevertheless, this does not affect the usable bandwidth of the system. In fact, since the only information needed for the optimization process are the reflection coefficients versus frequency at the DUT reference planes, no phase coherence is required between the data acquisition and the signal generation. Therefore, when it is desired to capture an extreme bandwidth around the fundamental or harmonics of interest, which is larger than the bandwidth of the acquisition A/D card itself, the LO frequency used for down-conversion can be simply stepped. In this way each time a different part of the frequency spectrum is brought into the reach of the A/D converter. Therefore, the only true bandwidth limitation of the system is given by the bandwidth of the AWGs.

To demonstrate the functionality of the setup, a test signal composed of 161 sinusoidal tones in the bandwidth between 2060 MHz and 2220 MHz is fed to a "thru," while the output active load is set to provide an open condition over the whole 160 MHz bandwidth. Figure 7 shows the measured reflection coefficient at the output reference plane of the "thru" as a function of frequency. It is clear from the plot that the desired reflection coefficient ($\Gamma_L = 1$) is achieved for all frequencies without any phase delay or amplitude unbalance.

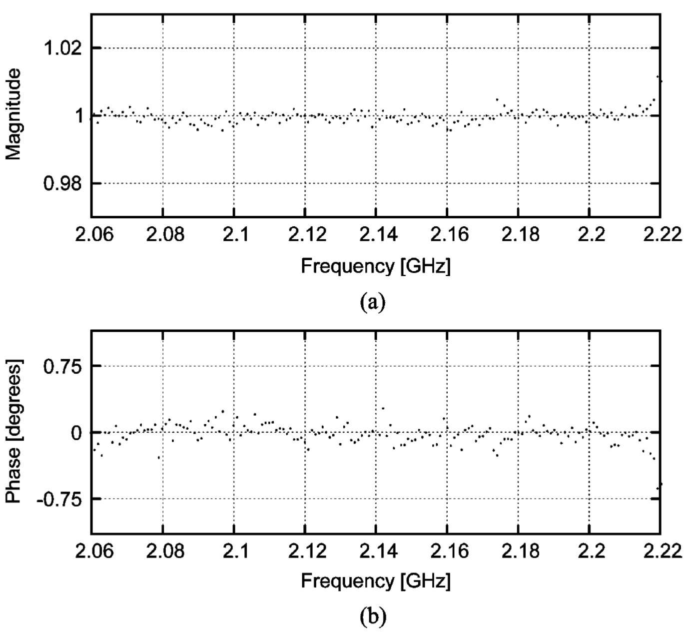

Figure 7. Measured reflection coefficient at the output reference plane of the DUT for a signal composed of 161 sinusoidal tones in a 160 MHz bandwidth.

B. Measurement Linearity and Dynamic Range

Since mixer based down-conversion is used to measure the reflection coefficients at the reference planes of the DUT in a wideband fashion, the dynamic range of the signal detection will be limited by the system noise floor at lower powers and by the mixers nonlinearities at very high power levels. Consequently, we have optimized our system for low noise and made use of high linearity mixers in the signal detection path. The resulting dynamic range is confirmed by IM3 measurements on a "thru" for a two-tone stimulus, yielding a spurious free dynamic range of 80 dB in the signal detection.

C. Active Load Dynamic Range

In order to control and optimize the reflection coefficient offered to the DUT for a particular frequency bin, it is necessary to measure the power waves for that frequency bin with high-accuracy and repeatability. For this reason one should avoid the use of noise data since this would make the optimization of the reflection coefficient unstable. Therefore the reflection coefficient is controlled only for those frequency bins in which the power exceeds a user specified threshold level. The selection of these power levels sets the dynamic range over which the load reflection coefficient can be controlled. We will refer to this as the "active load dynamic range." As mentioned in the Hardware Description Section, computer-controlled stepped attenuators are used to guarantee the full use of the voltage swing of the AWGs at all times. Therefore, the active load dynamic range will be limited only by the measurement noise, the IQ mixer imperfections and the robustness of the optimization algorithm used.

To give the reader an estimate of the "active load dynamic range" than can currently be achieved, a W-CDMA input signal centered at 2.14 GHz is fed to a heterojunction bipolar transistor (HBT) with an emitter area of 528 μm^2. On purpose the power of the signal is chosen so high that the active device starts to clip and generate intermodulation sidebands. The active load at the DUT output for the fundamental is set to provide a constant reflection coefficient ($\Gamma_L = |0.283|\angle{-135°}$) over the whole 20.4 MHz bandwidth. Figure 8 shows the measured reflection coefficient and power spectrum at the DUT output. Note that the reflection coefficient can be accurately controlled over the whole (broadened) spectrum over a 60 dB power range.

D. Measurement Speed

A commonly mentioned disadvantage of the open-loop concept over the closed-loop is the measurement

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speed, since in an open-loop system the reflection coefficients need to be optimized at every power level and loading condition. Modern equipment like the PXI express platform can overcome this classical speed limitation, since it provides very high data transfer rates in the up-and down-loading of the waveforms to the AWGs and A/D converters. Typical A/D download times for a measured complex modulated signal waveform are in the order of 60 ms, when using the maximum record length of the cards for averaging. The total upload time for all the AWGs (fundamental + second harmonic at input and output) is about 10 ms. Therefore the total time needed for setting source and load reflection coefficients around the fundamental and harmonic frequency bands (t_{meas}) based on the hardware specifications can be estimated as

$$t_{\mathrm{meas}} = (D_t \cdot 2 + S_t + U_t) \cdot N \qquad (4)$$

where D_t represents the total download time per point (60 ms), U_t is the total upload time per point (10 ms), S_t is the total switching time between fundamental and harmonic measurements (30 ms) due to switches and LO stepping settling time, and N is the number of iterations needed to optimize the reflection coefficients.

As the simple optimization algorithm currently used allows convergence with an average of about 100 iterations for a complex modulated signal, the total time for setting source and load reflection coefficients around the fundamental and harmonic frequency bands is restricted by the hardware to 16 s for any arbitrary complex modulated signal. Note that when considering conventional single-or multitone input signals, convergence is usually achieved in a much lower number of iterations. At this moment the software implementation still affects the overall time, however benchmarking of the critical software routines showed that major improvements are still feasible, which will bring the final speed of the system close to the numbers given by the hardware constrains.

Measurement Results

To demonstrate the unique capabilities of the realized active open-loop load–pull setup we have measured an NXP GEN 6 LDMOS device with a gatewidth of 1.8 mm. In our experiments, we use a drain current and voltage of 13 mA and 28 V, respectively. In this experiment the optimum fundamental load and source matching conditions are set to $\Gamma_{L,f1} = |0.6|\angle 45°$ and $\Gamma_{S,f1} = |0.5|\angle 90°$, while the input and output BB impedances enforce a short condition and the input

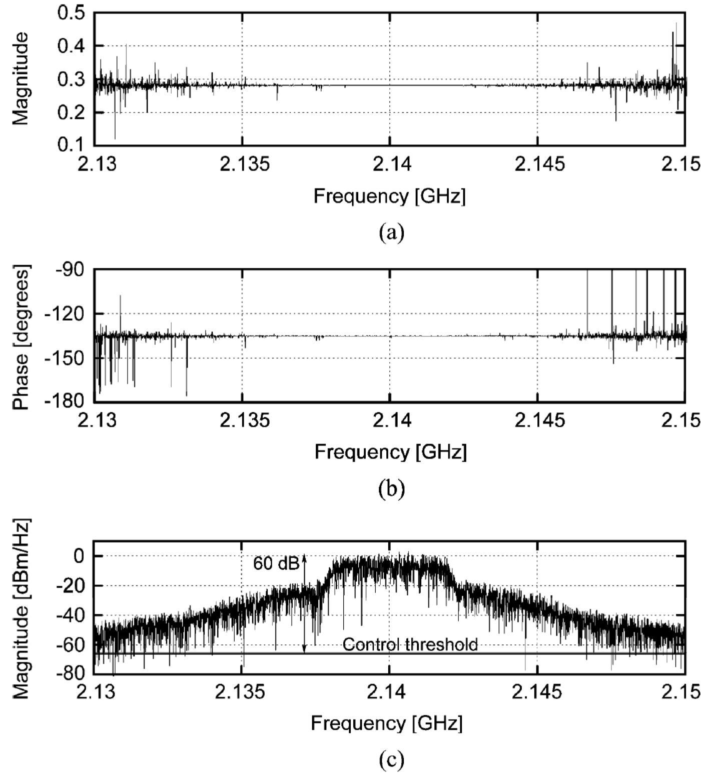


Figure 8. Measured load reflection coefficient and power spectrum, with a resolution bandwidth of 6 kHz, at the output reference plane of a HBT device over-driven with a W-CDMA signal yielding spectrum broadening. The user specified reflection coefficient is $\Gamma_L = |0.283|\angle -135°$. The power threshold level for the load reflection coefficient is set to 065 dBm. Note that the output reflection coefficient can be effectively controlled over a 60 dB power range.

and output second harmonics are set to circuit-like open conditions ($\Gamma_{L,f2} = \Gamma_{S,f2} = |0.95|$) to optimize the efficiency [21].

To highlight the excellent controllable bandwidth and the electrical delay free operation of the new measurement setup, a comparison is made with a previously developed state-of-the-art active harmonic load–pull system [9], which was especially optimized for minimum electrical delay. For this purpose, we use a two channel W-CDMA signal (centered at 2.135 and 2.145 GHz) and set the input and output reflection coefficients in the newly developed setup to the following two cases.

Case 1) Without electrical delay.

Case 2) With an electrical delay of 4.85 MHz for the fundamental source and load and 4.6 MHz for the second harmonic source and load.

Figure 9 illustrates the source and load matching conditions provided to the active DUT for the two different cases. Note that the filled markers represent the source and loading conditions for the two-carrier W-CDMA signal without any electrical delay, yielding completely overlapping points in the Smith chart. As shown in Figure 9, for the case with electrical delay the fundament load trajectory has been shifted, such that the optimum matching condition is now centered at 2.135 GHz. This was required to avoid the unstable region of the active device.

It is important to note that this is a comparison to the "best possible case" of a classical closed-loop active load–pull system since practical closed loops will be subject to amplitude variations within the control frequency bands. Moreover oscillation conditions in closed-loop systems for these very large bandwidths are difficult to avoid, due to the usage of wideband loop filters. Passive load–pull systems with harmonic tuning will have a comparable or even worse phase variation of the reflection coefficients versus frequency than the closed-loop system used in our comparison.

The measurement results are summarized in Table I. There is significant performance degradation for the active device when measured with electrical delay present in the reflection coefficients. This is also evident from Figure 10(a) and (b), which shows the power spectral density at the device output reference plane for the fundamental and second harmonic frequency bands.

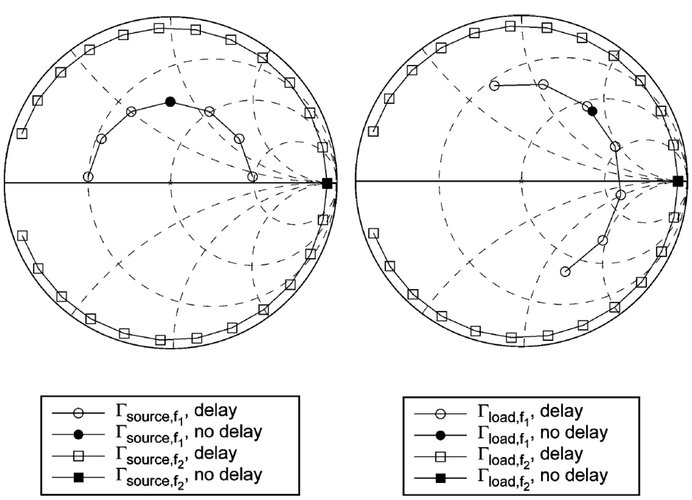
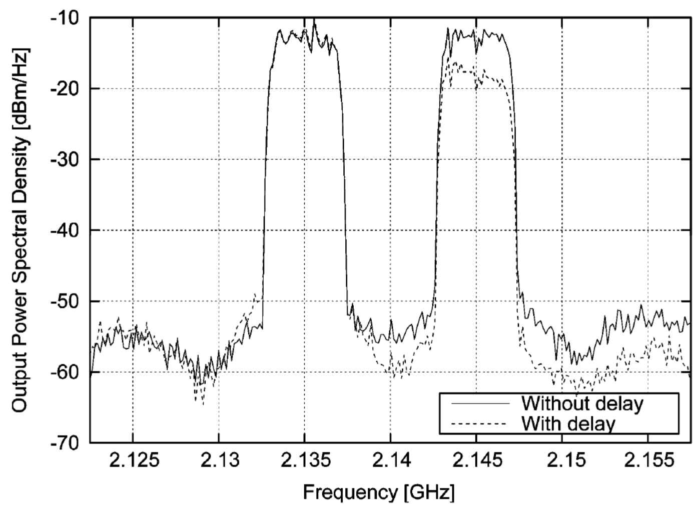


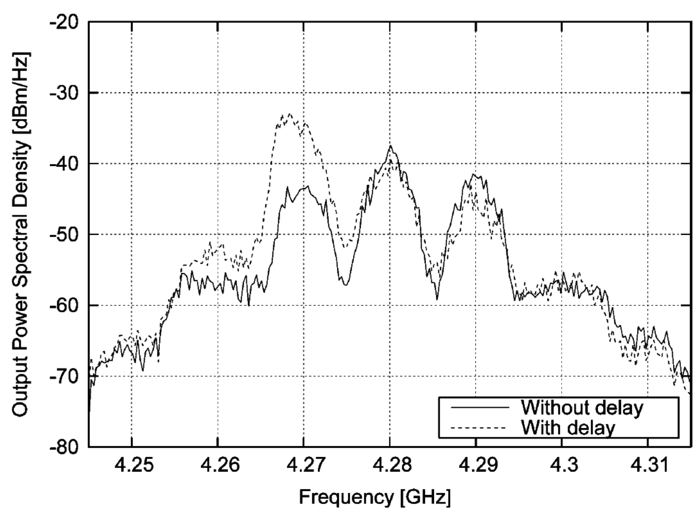
Figure 9. Source and load reflection coefficients at the device reference plane in the fundamental (2.1225–2.1575 GHz) and harmonic (4.245–4.315 GHz) frequency range, with electrical delay (open symbols) and without electrical delay (filled symbols).

TABLE I   MEASUREMENT RESULTS

	Without electrical delay	With electrical delay
PAE	24.2%	16.3%
P_{OUT} Ch. 1	20.3 dBm	20.5 dBm
P_{OUT} Ch. 2	20.6 dBm	15.4 dBm
ACLR1 Ch. 1	-43.9 dBc	-43.0 dBc
ACLR2 Ch. 1	-42.2 dBc	-41.6 dBc
ACLR1 Ch. 2	-42.1 dBc	-41.8 dBc
ACLR2 Ch. 2	-39.6 dBc	-39.2 dBc



(a)



(b)

Figure 10. Measured output power spectral density (dBm/Hz) versus frequency [GHz] of an NXP GEN 6 LDMOS device (gatewidth 1.8 mm) in the proposed load–pull setup. (a) At the fundamental frequency band using a 3 kHz resolution bandwidth. (b) At the second harmonic frequency band using a 6 kHz resolution bandwidth. The measurement is shown for the two cases with (dashed line) and without electrical delay (drawn line). The reflection coefficients offered to the DUT are given in Figure 9.

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Note that a 5 dB output power drop and close to 8% degradation of the power-added efficiency (PAE) can be observed, when comparing to the situation with no electrical delay.

Conclusion

A novel active harmonic load–pull setup has been presented which is suitable for large-signal device characterization under "real life" modulated signal stimulus. The new system is capable of synthesizing arbitrary source and loading conditions at the fundamental and harmonic frequencies over a frequency bandwidth of 160 MHz, and therefore permits testing of active devices under realistic (circuit-like) conditions. In fact, even user-defined reflection coefficients versus frequency can be downloaded to the measurement system in order to perform active device testing.

The IQ signal up-conversion and low IF down-conversion provide a very high dynamic range, which proves to be sufficient for the communication standards used today in industry. Furthermore this approach can be easily adjusted to any frequency band of interest (e.g., X-band) since all required parts are commercially available.

The IQ based open-loop approach, in combination with the frequency binning technique described, resolves all the conventional drawbacks of current load–pull techniques while still being competitive in terms of measurement speed.

The unique system capabilities have been demonstrated by testing an NXP GEN 6 LDMOS device under a two carrier W-CDMA input stimulus. To the best of the authors' knowledge, this is the first time that such a signal condition is used within an active harmonic load–pull system. Until now two-carrier W-CDMA testing, which is one of the most severe tests in industry for base station amplifiers, is only performed at the circuit board level, since conventional load–pull systems fail to handle these extreme bandwidths. The advantage of controlling the electrical delay has been made evident by comparing the performance difference in output power and power added efficiency of an LDMOS device, with a previous state-of-the-art system.

Finally, the selected hardware configuration leaves still room for numerous improvements and innovations and will hopefully lead, in the near future, to the development of new and innovative measurement techniques.

Acknowledgment

The authors would like to thank Auriga Measurement Systems, National Instruments, Agilent Technologies, and the PANAMA Project for their support of this project. The authors would also like to thank S. Theeuwen, NXP Semiconductors, Nijmegen, The Netherlands, for providing the GEN 6 LDMOS devices and J. Gajadharsing, NXP Semiconductors, for useful discussions and suggestions.

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