



Highly Reflective Load-Pull

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For the purpose of identifying the large-signal behavior of transistor devices, the use of linear S-parameter is often inadequate [1]. Large-signal characterization is essential for the estimation and determination of the device performance in the nonlinear domain, and the load-pull process [1]–[3] is one recommended approach for the characterization, optimization and design of transistor devices and radio-frequency (RF) power amplifiers (PAs).

The load-pull technique was first reported almost four decades ago [1]. It was a pioneering work, as it brought a paradigm shift in the characterization, measurement, design, and optimization of transistor devices and RFPAs. Although the load-pull setup reported in [4] can be considered very rudimentary, it has definitely advanced the way the design and optimization of RFPAs are carried out. These initial reports of the load-pull technique have resulted in subsequent advancements in the overall load-pull methodology [5]–[23].

The characterization and measurement of large-periphery transistor devices, which are employed in high-power modern wireless RFPAs, introduces two general problems. The first type of challenge includes stability and heating issues and relates to the device in combination with its assembly and packaging. These issues are commonly taken care of by suitable

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stabilization networks and by use of mechanically and thermally appropriate test fixtures to hold the devices [24]–[25]. The other type of challenge is due to the limitations of the associated load-pull measurement systems in presenting highly reflective impedances to the devices. This is the focus of this article, which elaborates on the problems encountered in the highly reflective load-pull characterization of high-density large-periphery devices, and also discusses the emerging solutions to overcome these challenges.

Load-Pull Basics

The load-pull technique utilizes systems that enable the synthesis of various load impedance environments in applications where transistor performance can be determined experimentally. In generic terms, the load-pull process refers specifically to creating an a priori known impedance to a device under test (DUT) in a precisely controlled fashion, in order to extract optimal performance of the device [17]. Load-pull systems allow for the determination of appropriate matching impedance values while physically changing the reflection coefficient, Γ_L , as shown in Figure 1. The synthesized matching impedance aids in the extraction of parameters that help to meet design goals such as output power, dc-to-RF power conversion efficiency, operating power gain and gain compression, and power-added efficiency, from the transistor DUT.

The load impedance, Z_L , at the output port of the device, the incident and reflected traveling waves, a_2 and b_2 , at the output port, and Γ_L are related by following relationships:

$$\Gamma_L = \frac{a_2}{b_2} \quad (1)$$

$$\Gamma_L = \frac{Z_L - Z_0}{Z_L + Z_0}, \quad (2)$$

where Z_0 is the characteristic impedance of the system in which the DUT is used (normally 50 Ω).

The function of the tuner is to vary the magnitude and phase of the reflected traveling wave, a_2 , so as to synthesize an appropriate Γ_L . The tuners can be active [4] or passive [5], depending on the applications. This functionality can be obtained by altering the tuner setting by way of moving the slug or stub up and down and back and forth in a passive tuner, as shown in Figure 2, or by actively injecting a magnitude- and phase-controlled signal in active load pull [18]–[23]. In passive tuners, as the probe/slug/stub is inserted into the tuner transmission line, it introduces mismatch by adding parallel susceptance. The parallel susceptance increases as the probe/slug/stub approaches the line and aids in the synthesis of the desired reflection coefficient. The magnitude of the impedance mismatch is determined by the probe position (depth) and the phase of the impedance mismatch is determined by the carriage position (length).

High Reflection Load-Pull

Large-periphery transistor devices possess extremely low output impedances, on the order of 1 Ω and often sub-1 Ω [26]–[28]. They therefore require load-pull systems capable of creating highly reflective environments, with $|\Gamma_L|$ approaching 1.0, for the extraction of optimal performance. In practical applications, however, the requirement to establish highly reflective environments puts severe constraints on traditional passive and active load-pull systems [29], [30].

The major drawback of the standard passive load-pull system is the limited maximum achievable reflection coefficient, $|\Gamma_L|$ (with a maximum usually around 0.75), due to the inherent losses in the tuner, the measurement network and the device fixture [31]. Furthermore, the high mismatch between the output impedance of large-periphery DUTs and the impedance environment of standard load tuners, which is normally around 50 Ω , has the potential to create a high voltage standing wave ratio (VSWR) in the measurement system. A high VSWR can generate very large voltage and current peaks in the measurement system, which can damage the DUT and the measurement system, due to corona discharge [32].

A high impedance mismatch between the load-pull system and the DUT also leads to increased measurement uncertainty [33], which is caused by the limitations of vector network analyzers (VNAs) in characterizing high-reflection, low-loss two-port networks, such as tuners. According to the well-known rule of thumb in metrology, the ratio between the reference impedance (here 50 Ω) and the test object should always remain less or equal to 10 (VSWR \leq 10:1 or $|\Gamma_L| \leq$ 0.8), in order to achieve a reliable measurement [33]. However, the required $|\Gamma_L|$ in highly reflective load-pull measurement systems approaches unity, thereby leading to increased measurement uncertainty.

Traditional active load-pull systems are capable of generating high reflection coefficients, even beyond the boundary of the Smith chart, but also suffer from high VSWR and increased measurement uncertainty. In addition, feedback and feed-forward active load-pull systems often suffer from unstable

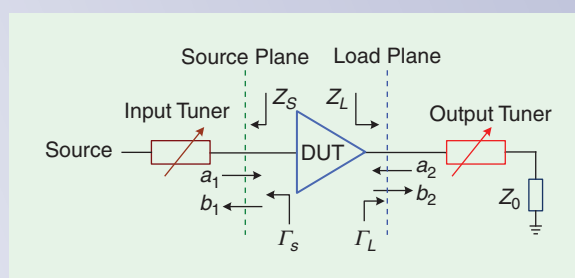


Figure 1. Block diagram depicting source and load reflection coefficients Γ_s and Γ_L , respectively. The impedance tuners at the input and output ports synthesize the required Γ_s and Γ_L .

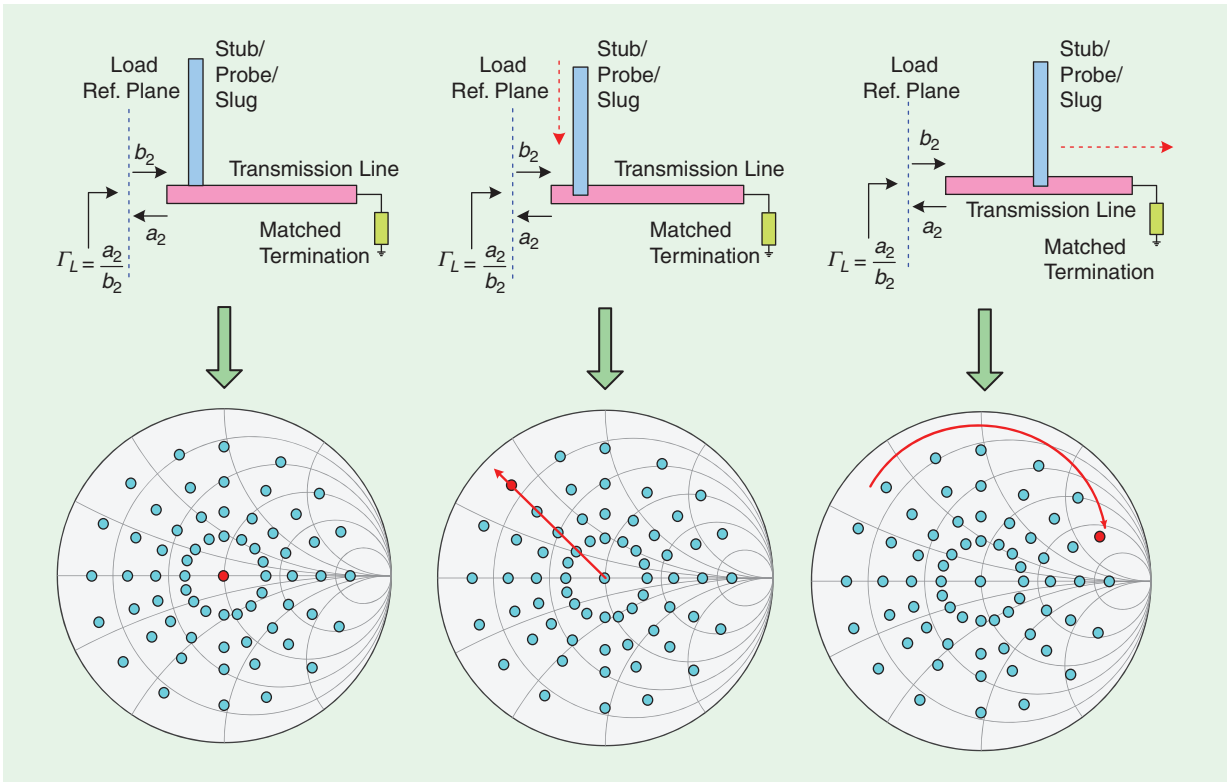


Figure 2. Probe/slug/stub position and its movement defines the load reflection coefficient, Γ_L , at the load reference plane in a passive load-pull system.

operation, which has the potential to damage the DUT and the measurement setup. The instability in the feedback active load-pull system [6], shown in Figure 3, occurs mainly due to a large gain ripple in the characteristics of the loop amplifier, which is generally eliminated by inserting a highly selective bandpass filter, such as a yttrium iron garnet (YIG) filter. The YIG filters possess flat gain over a limited useful bandwidth, thereby eliminating the effects of a large gain ripple in the response of the loop amplifier. The downside of the YIG filter insertion in the loop, however, is the resulting limited bandwidth of the load-pull system. The active feed-forward

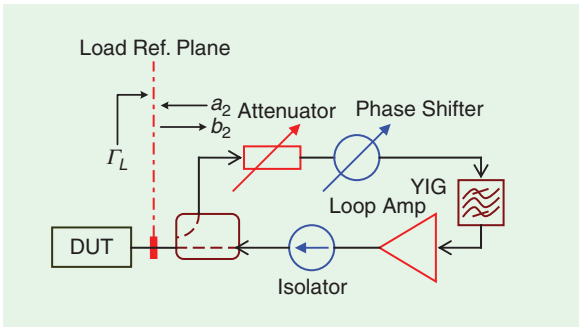


Figure 3. Block diagram of a feedback active load-pull system which modifies the traveling wave, b_2 , and feeds it back as a phase coherent traveling wave, a_2 , in order to synthesize reflection coefficient, Γ_L , at the load reference plane.

load-pull system [4] remains stable for the majority of practical applications, but can oscillate when the loop gain tends to extremely high values, which may be the case during large-periphery DUT measurement and characterization. In order to guarantee unconditional stability, an additional condition at the DUT input is necessary. Active open-loop load-pull systems, shown in Figure 4, usually do not suffer from instability issues.

Nevertheless, the required load-pull power, P_{LP} , either from the load-pull source in active open-loop or from the feedback loop in active closed-loop load-pull systems increases greatly, due to the large

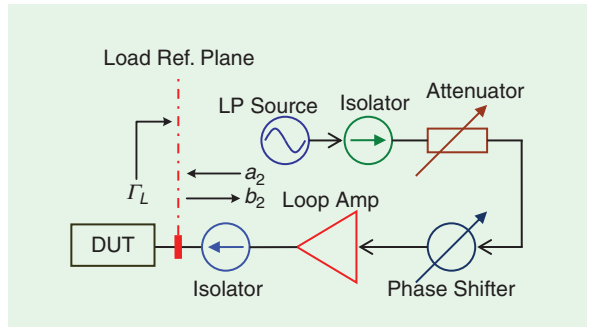


Figure 4. Block diagram of an open-loop active load-pull system which injects a phase coherent traveling wave, a_2 , from an external signal generator, LP Source, in order to synthesize Γ_L at the load reference plane.

impedance mismatch between the large-periphery DUT and the load-pull test set. This increase in P_{LP} directly translates to the limited usefulness of these systems in practical applications, as they rely on expensive loop amplifiers to synthesize high reflection coefficients.

The quarter-wave-transformer method, pre-matched-tuning technique, and broadband-impedance-transformer technique [19], [24], [34]–[41] are the three most common improvements to have been incorporated in traditional active and passive load-pull systems, in order to overcome the above-mentioned problems and ensure that the demands of large-periphery DUT load-pull characterizations are met. The technique of using probing couplers is also gaining in popularity, as it directly captures the traveling waves from the DUT plane, thus, the synthesized reflection coefficient is not affected by the loss in the measurement probes and network [42].

The most common solutions to the problems encountered in the use of high-reflection load-pull systems presented in the literature thus far can be found in [19], [34]–[41], and [45]. As a case study, analysis has been carried out on a 100 W device, but the techniques presented here are equally applicable to any power device. A simple hybrid load-pull configuration is described that may be configured for use in large-periphery device characterization.

Problems Due to a High Voltage Standing Wave Ratio

The problem of a high VSWR appears in both passive and active load-pull systems. The effect of a high VSWR on the overall test and measurement setup can be analyzed by considering a generic model of a passive load-pull system, as given in Figure 5. The DUT is represented by an equivalent voltage source, V_d , and a series impedance, Z_d , which is connected to the passive tuner via a lossless $50\ \Omega$ transmission path. In a standard load-pull setup, a passive network consisting of couplers is used for measuring the traveling waves at the DUT ports; therefore, the load-pull

setup always has a transmission path between the DUT and the load tuner.

The voltage, $V(z)$, and the maximum voltage, V_{\max} , on the transmission path are given by (3) and (4), respectively [43].

$$V(z) = \sqrt{\frac{2 \cdot P_d \cdot Z_0}{1 - |\Gamma_d|^2}} (1 + \Gamma_d \cdot e^{-2j\beta z}) \quad (3)$$

$$V_{\max} = \sqrt{2 \cdot P_d \cdot Z_0 \cdot \text{VSWR}} \quad (4)$$

where

$$\text{VSWR} = \frac{1 + |\Gamma_d|}{1 - |\Gamma_d|} \quad (5)$$

and P_d is the maximum power delivered by the DUT to the load, β is the propagation constant along the transmission path, and Γ_d and Γ_{lp} are the reflection coefficients at the DUT and load-pull planes, respectively. The term $e^{-2j\beta z}$ represents the phase of the reflection coefficient along the transmission path.

The data in Table 1 are the estimated maximum voltage values, V_{\max} , along the measurement system transmission path, obtained from (4), for transistor devices with arbitrary low impedances. For a simpler analysis, only real impedances have been assumed. It can be observed from Table 1 that a high VSWR in the load-pull setup used for characterizing large-periphery devices can generate very high peaks of voltage waves along the transmission path, which can damage the DUT and the measurement system.

When used for plotting voltage $V(z)$ along the transmission path for a 100 W ($1\ \Omega$) device at a frequency of 2.1 GHz, (3) gives the results shown in Figure 6, which shows that massive voltage peaks of around 707 V may occur along the transmission path. A 100 W device with $1\ \Omega$ output impedance can handle a maximum voltage of 100 V; therefore, in order to avoid a peak voltage of 707 V, the tuner should be placed as close as possible to the DUT [32], which, in this case, corresponds to position 1 in Figure 6. However, the length of the transmission path corresponding to position 1 is

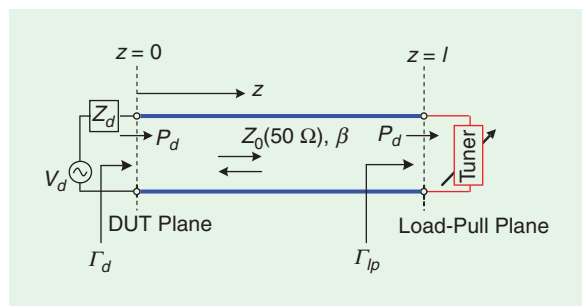


Figure 5. Generic model of the passive load-pull characterization setup which consists of an equivalent voltage and impedance of the DUT at one end and an impedance tuner at the other end of a lossless transmission section.

TABLE 1. VSWR and maximum voltage generation along the measurement system transmission path.

Device Output Power (W)	Assumed Output Impedance (Ω)	VSWR	V_{\max} (V)
50	2	24.974	353.37
100	1	49.891	706.34
200	0.5	99.503	1,410.69
250	0.4	124.786	1,766.25
400	0.25	199	2,821.35
500	0.2	249	3,528.46

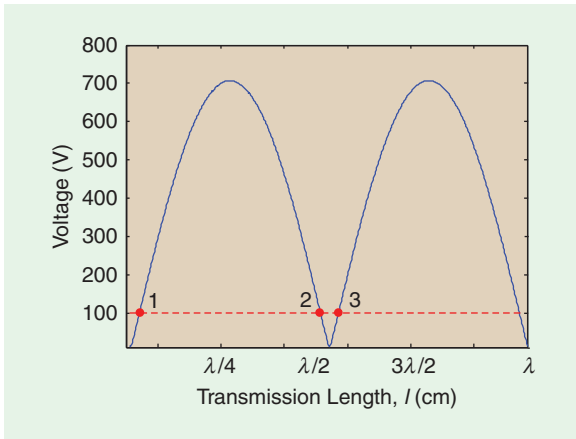


Figure 6. Voltage generation in a passive load-pull measurement system when characterizing a 100 W (1 Ω) device at 2.1 GHz (similar curves can be obtained for an active load-pull system).

too small at 2.1 GHz; that is, it is not possible to place a measuring network such as a coupler between the DUT and the load tuner if the length of the transmission path corresponding to position 1 is chosen.

The closest the tuner can be placed to the DUT corresponds to either position 2 or position 3 in Figure 6; therefore, there will be at least one voltage peak in the measurement system. This large voltage peak can cause a breakdown between the transmission path and the ground, due to the corona effect [44], which can eventually damage the tuner, the whole measurement setup and/or the DUT. The problem of high VSWRs will get worse with the increase in operating frequency, as well as with the power rating of the device being characterized. Therefore, the load-pull approaches, active or passive, that employ 50- Ω transmission lines between the DUT and the load-pull component are not suitable for high-power device characterization.

The Problem of High Load-Pull Power

The load-pull power, $P_{LP} = |a_2|^2$, requirement for the synthesis of the load reflection coefficient at the DUT plane can be analyzed through a generic model of an

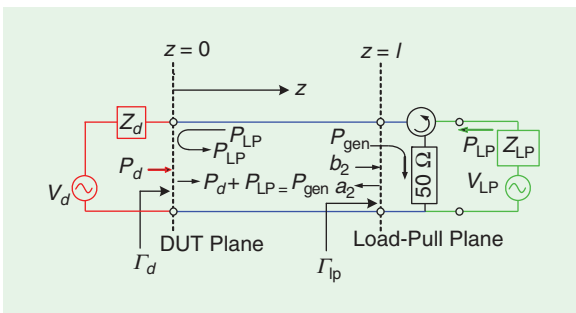


Figure 7. Generic model of an active open-loop load-pull measurement setup. It contains the equivalent voltage and impedance of the DUT at one end and the active load-pull component at the other end of a transmission line section [32].

open-loop active load-pull setup, as shown in Figure 7. In this model, the terms V_{LP} and Z_{LP} represents the voltage and the impedance of the active load-pull tuner. In this model, Z_{LP} is set to the system characteristic impedance $Z_0 = 50 \Omega$. The circulator isolates the DUT and the active load-pull tuner.

Equation (6) is the relationship between the desired load reflection coefficient and the required load-pull power, P_{LP} , in order to probe a device that has a maximum output power of P_d . It clearly establishes that the P_{LP} will become extremely high during the characterization of large-periphery low-impedance devices.

$$P_{LP} = P_d \frac{|\Gamma_d|^2}{1 - |\Gamma_d|^2}. \quad (6)$$

The plot in Figure 8 depicts the load-pull power, P_{LP} , requirement for active load-pulling of a 100 W (1 Ω) device. It is apparent that the P_{LP} required for load-pulling a 100 W (1 Ω) device is around 1,175 W. However, the required P_{LP} of 1,175 W makes active load-pull systems impractical, due to the high costs involved. It is these incurred costs that limit the use of active load-pull systems in their standard configuration [13]–[16] in high-power device characterization.

Quarter-Wave Impedance Transformation Technique

The basic idea behind this approach is the transformation of the load-pull environment from 50 Ω to a lower impedance value by using a quarter-wave ($\lambda/4$) impedance transformer. Figure 9 gives a pictorial representation of the quarter-wave impedance transformation concept. Before transformation, the tuning range is limited, with the matched condition at point *a* in the Smith chart. The load tuner is connected to the quarter-wave transformer via a 50 Ω line; and, after the transformation, the matched condition moves to point *c* in the Smith chart. The transformation of the measurement system's impedance to a lower value

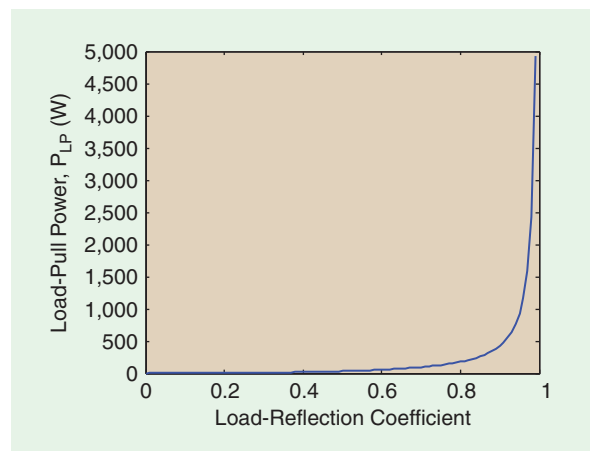


Figure 8. Load-pull power, P_{LP} , requirement as a function of the load reflection coefficient, Γ_d , at the DUT reference plane for a 100 W (1 Ω) device.

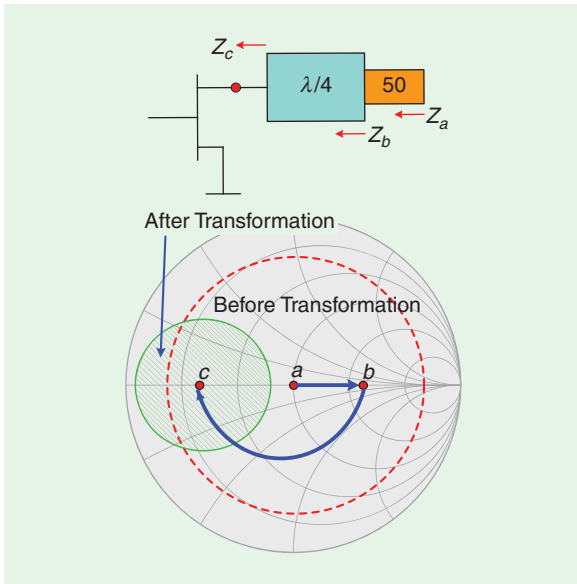


Figure 9. Pictorial representation of the quarter-wave transformation network.

results in reduced impedance mismatch between the system and the DUT.

The reduced impedance mismatch removes the possibility of a high VSWR and, hence, large voltage peaks in the measurement system. Figure 9 also shows that, by using the quarter-wave transformer, the range of maximum achievable reflection coefficient is enhanced. Thus, the quarter-wave impedance transformer serves the purposes of both minimizing high VSWR and enhancing the tuning range [34], [35]. High reflection coefficients are created in two steps using this technique [34]; and, because the VNAs do not have to measure very high reflection coefficients, the measurement accuracy using the quarter-wave transformation network also improves. These improvements, however, come at the cost of reduced Smith chart coverage. Therefore, the quarter-wave transformer needs to be redesigned for cases when different DUTs need to be characterized and measured or the frequency of operation changes, so that the required Smith chart regions are not precluded. Further, the quarter-wave transformer approach is narrowband in nature and is, therefore, primarily useful for fundamental load-pull characterization of high-power DUTs. This technique requires additional hardware and calibration steps, increasing the size and cost of the system and the calibration time.

Prematched Tuner-Based Load-Pull Technique

Figure 10 depicts the concept of prematched tuning. Prematching tuners employ two independent RF probes, which are used to generate high reflection coefficients. These tuners are capable of individually generating smaller reflection coefficients, but when used in combination with one another enhance the

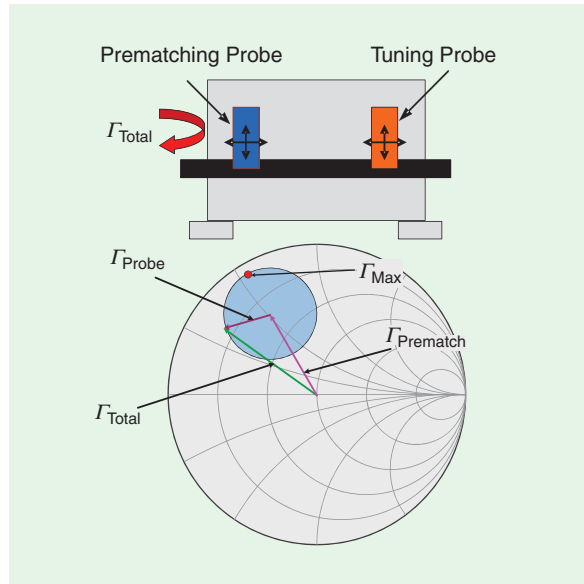


Figure 10. Prematched impedance tuning concept (courtesy of Focus Microwaves).

achievable reflection coefficient, as shown in Figure 10. In this technique, one of the tuners moves the matched condition to the desired region of the Smith chart, while the other tuner forms the loci of the reflection coefficient around this new matched position. State-of-the-art prematched automated passive load-pull tuners [36], [37] are able to synthesize reflection coefficients in the order of 0.90–0.92 magnitude, which is higher than what can be achieved using a standard passive tuner, as shown in Figure 11.

The prematched load-pull system can tune to any angle and overcome the typical limitations of quarter-wave transformers [34], [35]. Furthermore,

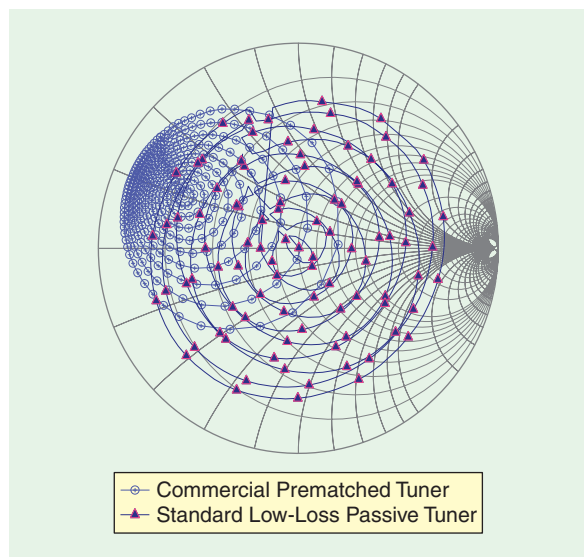


Figure 11. Comparison of the reflection coefficients achieved using a standard passive tuner and a commercial prematched tuner [37] at 2.425 GHz.

the prematched load-pull setups possess large bandwidth, in principle the same as a single-slug tuner (about one decade), and are, therefore, suitable for harmonic applications. Moreover, the measurement uncertainty is also reduced, as the high reflection coefficients are created in two steps; and, as a result, measurements made under high-reflection conditions are not significantly affected by the limitations of the VNA's accuracy in measuring low impedances. The major drawback of this technique is the dependence of the accuracy of the synthesized impedances on an interpolation algorithm when synthesizing desired reflection coefficients that do not lie on the calibration

grid [46]. Sometimes, the overall measurement speed also is reduced, due to the precharacterization step of the ensemble of both impedance tuners [46].

Enhanced Loop Load-Pull Architecture

The enhanced loop load-pull architecture [38], as shown in Figure 12, proposed recently by the University of Calgary, is another simple and effective technique for achieving a highly reflective impedance environment. This setup can synthesize load reflection coefficients up to 0.97, which is higher than the prematched load-pull system, as shown in Figure 13.

In this setup, the total reflection coefficient, Γ_L , is dependent on the contribution from the load-pull tuner, *Tuner2*, and the additional contribution Γ_{LOOP} from the passive loop. The flow diagram in Figure 12 depicts the process of generation of the Γ_L values at the DUT plane contributed by the load-pull tuner and the passive loop, while (7) gives the relationship for the generation of the enhanced load reflection coefficient, Γ_L . The factor Γ_{LOOP} contributed by the passive loop is given in (8).

$$\Gamma_L = \frac{a_2}{b_2} = S_{11TUNER} + \frac{S_{12TUNER} \cdot S_{21TUNER} \cdot \Gamma_{LOOP}}{1 - S_{22TUNER2} \cdot \Gamma_{LOOP}} \quad (7)$$

$$\Gamma_{LOOP} = \frac{b_3}{a_3} = |\Gamma_{LOOP}| \cdot e^{-2j\beta L_2}, \quad (8)$$

where Γ_{LOOP} is a complex factor that depends on the passive loop component characteristics, that is, the transmission factors of the coupler and the circulator. It is also dependent on the phase velocity, β , of the traveling waves and the length of passive loop, L_2 . The length of the loop can be altered by deploying cables of appropriate lengths. It has been found that only three cables of varying lengths, L_2 , are required to cover the entire Smith chart using this setup [38]; therefore, calibration and measurement times are reduced when compared with the state-of-the-art prematched load-pull techniques [37].

This topology is equally useful for DUTs requiring high reflection coefficients at their input and output ports [38]. Furthermore, given that the reflection coefficients are generated in two steps, the VNA does not have to measure high reflection coefficients; therefore, the measurements are not affected by the VNA's limitation in accurately characterizing low-loss, high-reflection two-port networks. It is anticipated that the enhanced loop source- and load-pull architecture topology in conjunction with the broadband impedance transformer technique, explained later in this article, will be useful for applications in all-passive harmonic load-pull setups.

Broadband-Impedance-Transformer Technique

The diagram in Figure 14 depicts the model of an active open-loop load-pull system with an impedance

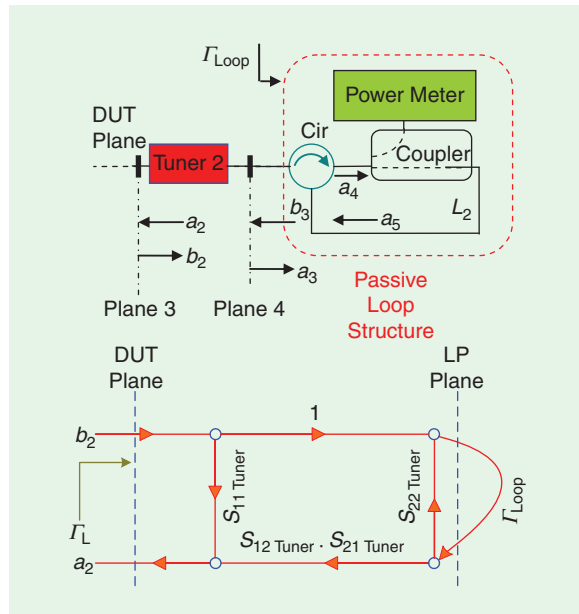


Figure 12. Enhanced loop load-pull tuner and the associated error flow model [38].

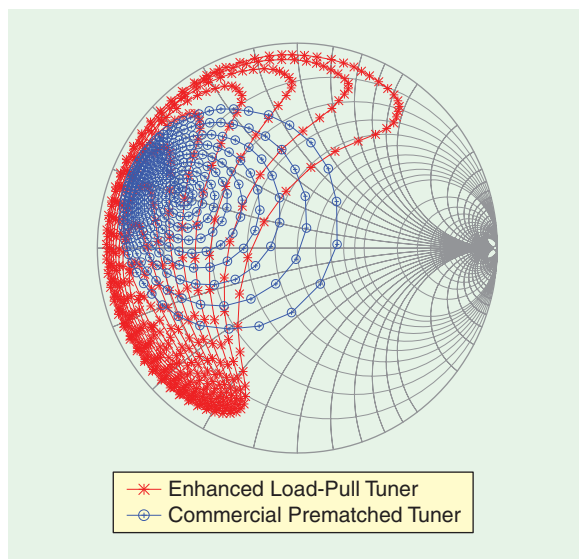


Figure 13. Comparison of achievable load reflection coefficient, Γ_L , using a commercial prematched tuner [37] and the enhanced loop tuner [38] at 2.425 GHz.

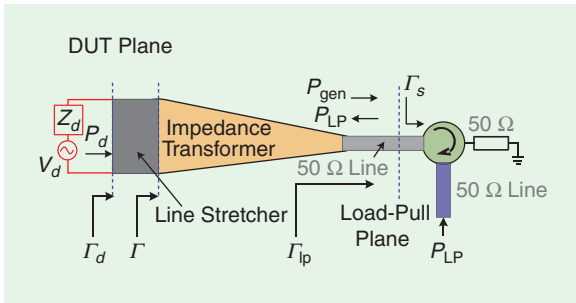


Figure 14. Representation of a modified active open-loop load-pull system with the incorporation of an impedance transformer.

transformer. The larger geometry and, hence, the smaller impedance of the impedance transformer is connected to the output port of the DUT. The signal flow diagram in Figure 15 depicts the mechanism of load reflection coefficient creation at the DUT reference plane.

In Figure 15, $a_s e^{j\varphi_s}$ represents the signal injected at the high-impedance end of the impedance transformer. Traveling wave $b_d e^{j\varphi_b}$ is the output of the device, and traveling wave $a_d e^{j\varphi_a}$ represents the actual power injected by the load-pull generator at the DUT plane. The term $|\Gamma|$ represents the impedance transformer reflection coefficient, that is, the transformation ratio of the impedance transformer; whereas, α is its phase. The synthesized reflection coefficient at the DUT plane is given by

$$\Gamma_d = \frac{a_d e^{j\varphi_a}}{b_d e^{j\varphi_b}} = |\Gamma_d| e^{j(\varphi_a - \varphi_b)}. \quad (9)$$

According to (9), the load synthesized at the DUT plane in the active load-pull process is a function of the relative phase between φ_a and φ_b ; therefore, without loss of generality, φ_b can be taken as a reference and set to zero in (9). Now the required load-pull power, P_{LP} , for the synthesis of the required reflection coefficient at the DUT plane can be deduced from Figure 15 [32].

$$P_{LP} = \frac{P_d (|\Gamma_d|^2 + |\Gamma|^2 - 2|\Gamma_d||\Gamma|\cos(\varphi_a - \alpha))}{(1 - |\Gamma_d|^2)(1 - |\Gamma|^2)}. \quad (10)$$

The load-pull power, PLP, is related to the transformation ratio of impedance transformer, $|\Gamma| e^{j\alpha}$, and the desired reflection coefficient, $|\Gamma_d| e^{j\varphi_a}$ in (10). It is evident from the results in Table 2 that the increasing transformation ratio, Γ , reduces the required PLP for the synthesis of the desired load reflection coefficient, Γ_d . In this case, it is assumed that phases φ_a and α are equal (the best-case scenario) and, thus, the cosine term in (10) is unity.

Equation (10) also conveys that the P_{LP} is regulated by the difference between the phase of the reflection coefficient at the DUT plane, φ_a , and the phase of the impedance transformer reflection coefficient, α . The

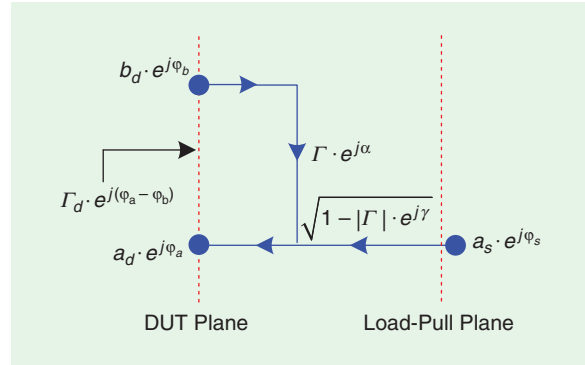


Figure 15. Signal flow diagram for an active load-pull setup with an impedance transformer between the DUT and the load-pull signal generator [48].

TABLE 2. Impact of transformation ratio on the load-pull power required for optimal synthesis of the load reflection coefficient (100 W (1 Ω) device).

Transformation Ratio	Required P_{LP} (W)	Transformation Ratio	Required P_{LP} (W)
50:5	78	50:20	442
50:7	126	50:30	686
50:9	174	50:40	931
50:10	198	50:50	1,175

required P_{LP} gets smaller as $\varphi_a - \alpha \rightarrow 0$ and reaches a minimum when φ_a equals α .

Figure 16 shows a plot of (10) for the swept value of $\varphi_a - \alpha$ for a 100 W device with the incorporation of a 50:7 impedance transformer. It is apparent from Figure 16 that a difference of 180° between φ_a and α results into a very high value of P_{LP} for synthesis of the optimal reflection coefficient. It is also apparent that

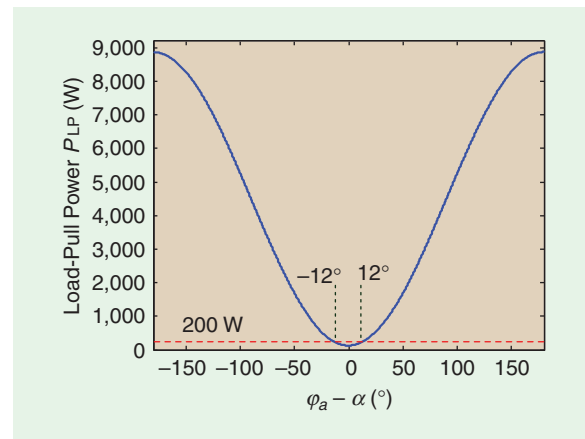


Figure 16. Load-pull power requirements as a function of the relative phase difference between the phases of Γ_d and Γ for a 100 W (1 Ω) device by incorporating a 50:7 impedance transformer [32].

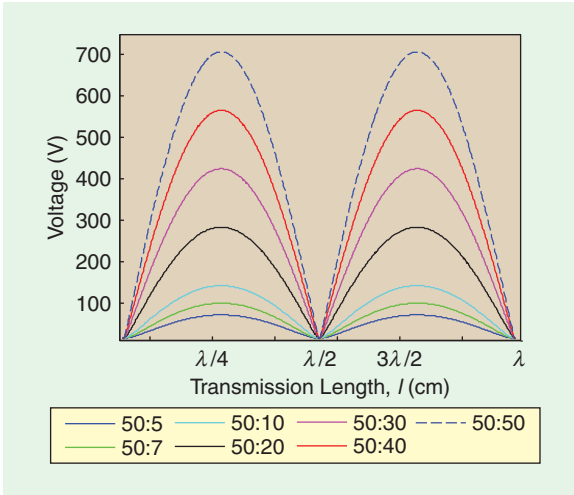


Figure 17. Estimated voltage for different transformation ratios in active and passive load-pull measurement systems when characterizing a 100 W (1 Ω) device at 2.1 GHz.

an active load-pull setup with 200 W of available P_{LP} requires that the term $\varphi_a - \alpha$ be below $\pm 12^\circ$.

The term $\varphi_a - \alpha$ can be minimized by adding a length of low-impedance line or a line stretcher at the DUT output port [32]. The effect of this additional line can be removed through the TRL calibration technique [47]. Thus, the position of the impedance transformer between the DUT and the load-pull is not that significant. Angle α is a function of the transformation ratio; hence, careful consideration is required when designing the impedance transformer.

Figure 17 shows voltage variation, plotted using (3), for the example 100 W (1 Ω) device for different transformation ratios. This figure demonstrates that the increased transformation ratio of the impedance transformer also decreases the maximum voltage. Additionally, the incorporation of a broadband impedance transformer makes this architecture appropriate for high-power harmonic load-pull applications [19], [32], [48].

The results in Table 2 and Figure 17 reveal that the higher the transformation ratio, the smaller the P_{LP}

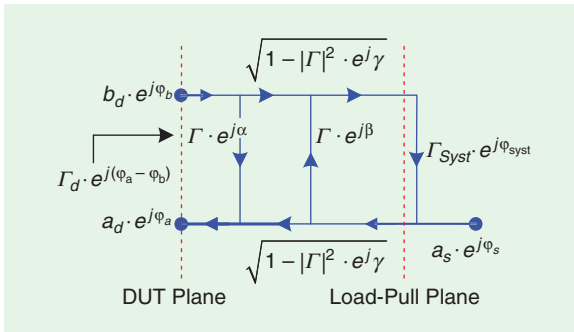


Figure 18. Signal flow diagram for a non-50- Ω measurement system with the impedance transformer between the DUT and the load-pull system [32].

requirement and the voltage peak in the measurement system. In practice, however, the highest transformation ratio is not the optimal choice. For an unknown device, the choice of optimal transformation ratio is dictated by its output impedance. Therefore, it is necessary to carry out load-pull measurements on an unknown device without the impedance transformer and then determine where the DUT output impedance could be located on the Smith chart. When this output impedance is known, an appropriate impedance transformer is then chosen such that the P_{LP} demand and voltage peak are minimized at the same time. The major drawback of this technique is that a separate impedance transformer is needed if the DUT or frequency of operation changes, so that the required Smith chart region is not precluded.

Hybrid Load-Pull System

It is apparent that the incorporation of an impedance transformer between the DUT and the load-pull setup helps in overcoming the problems of high voltage peaks in active and passive setups. The impedance transformer also aids in reducing the load-pull power requirements in an active load-pull setup. However, in the preceding sections, it was assumed that the active load-pull measurement setup components have a 50 Ω characteristic impedance, but in practice, they do not have perfect 50 Ω impedances. Therefore, the system reflection coefficient, Γ_s , is not zero. Thus, in the non-50 Ω environment, the signal flow diagram also changes, which is shown in Figure 18. Equation (11) for the P_{LP} can be derived [32] from the flow diagram in Figure 18. Once again, the reference phase φ_b has been set to zero.

$$P_{LP} = |a_s|^2 = \frac{P_d \left(\begin{array}{l} \Gamma_d^2 + |\Gamma|^2 + |\Gamma_s|^2 + |\Gamma_d|^2 |\Gamma|^2 |\Gamma_s|^2 - \\ 2|\Gamma_s| \times |\Gamma| \times (1 + |\Gamma_d|^2) \times \cos(\beta + \varphi_s) - \\ 2|\Gamma_d| \times |\Gamma| \times (1 + |\Gamma_s|^2) \times \cos(\varphi_a - \alpha) + \\ 2|\Gamma_d| \times |\Gamma_s| \times |\Gamma|^2 \times \cos(\beta + \varphi_s + \varphi_a - \alpha) - \\ 2|\Gamma_d| \times |\Gamma_s| \times \cos(\varphi_a - 2\gamma - \varphi_s) \end{array} \right)}{(1 - |\Gamma_d|^2)(1 - |\Gamma|^2)}$$

where γ is the phase of the impedance transformer's transmission coefficient, and β is the phase of the transformer reflection coefficient.

To see the effect of Γ_s on P_{LP} , (11) is plotted with a 50:7 impedance transformer and a measurement system reflection coefficient of $\Gamma_s = 0.1$ for a 100 W (1 Ω) device. For the overall measurement system, both phase terms φ_a and α are set to π , as this gives the optimal and lowest P_{LP} as per (10). Assuming that the phase γ of the impedance transformer transmission coefficient is equal to π , then sweeping the phase φ_s of the measurement system reflection coefficient in (11) gives the result for P_{LP} shown in Figure 19. It can be deduced from Figure 19 that even a small deviation in the system reflection coefficient Γ_s dramatically

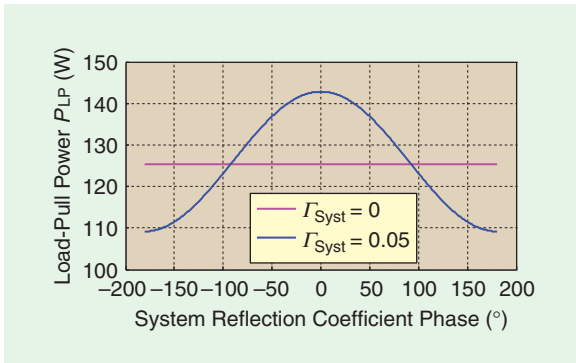


Figure 19. Load-pull power requirements as a function of the measurement system reflection coefficient ($\Gamma_s = 0.05$) phase compared to a perfect $50\ \Omega$ system with a 50:7 impedance transformer (for a 100 W ($1\ \Omega$) device).

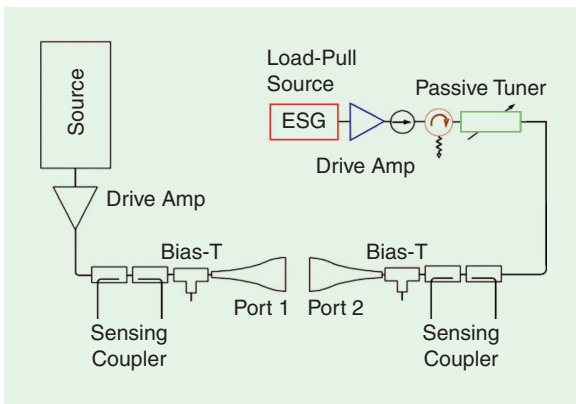


Figure 20. Block diagram of a hybrid load-pull measurement setup that incorporates an impedance transformer, active open-loop load-pull components and a passive tuner.

changes the P_{LP} requirement for the synthesis of the desired impedance at the DUT plane. Interestingly, this mismatch problem can be utilized in such a way that the P_{LP} requirement in an active load-pull system gets significantly reduced. This can be achieved by making a hybrid of the active and passive load-pull components, as shown in Figure 20.

The hybrid load-pull system consists of passive tuners and vector signal generators and other components from an active open-loop load-pull system. This configuration is quite useful for the situation when the signal generator in the active load-pull setup has limited P_{LP} . This hybrid setup can be deliberately mismatched by the passive tuners to dramatically reduce the P_{LP} requirements.

The plot in Figure 21 presents the P_{LP} requirement for load-pulling a 100 W ($1\ \Omega$) device using the hybrid load-pull setup with a 50:7 impedance transformer for different system reflection coefficients, Γ_s . It is clear that an increase in the system reflection coefficient magnitude results in an increase or decrease in the P_{LP} requirement, depending on phase φ_s of the system reflection coefficient. Therefore, the optimal use

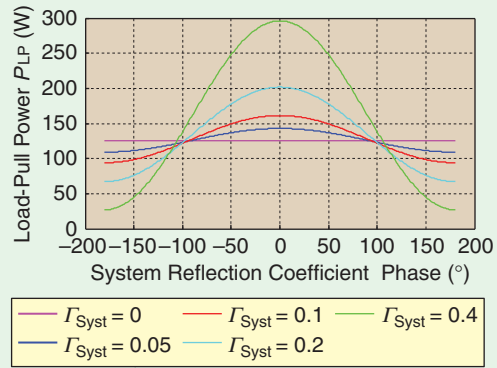


Figure 21. Load-pull power requirements as a function of the system reflection coefficient phase, φ_s , for different system reflection coefficients when using a 50:7 impedance transformer (for a 100 W ($1\ \Omega$) device).

of the system depends on the available impedance transformer, the tuning resolution of the passive tuner, and an extra line stretcher between the DUT and the impedance transformer. These all are important in determining the optimal phase and magnitude of system reflection coefficient, Γ_s , in order to find the P_{LP} requirement for synthesizing the specified load reflection coefficient, Γ_d .

The effectiveness of the hybrid load-pull setup in Figure 18 becomes clear when characterizing a 100 W laterally diffused metal oxide semiconductor (LDMOS) device with a specified optimal impedance of $1.7-j2.55\ \Omega$ using a 50:7.15 impedance transformer at a fundamental frequency of 2.14 GHz. The results shown in Figures 22 and 23 convey that, although the Smith chart coverage is reduced due to the incorporation of an impedance transformer, the setup is able to successfully extract the optimal output power from the

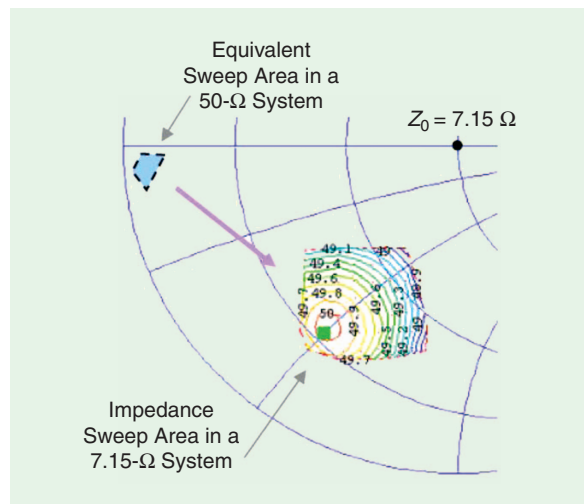


Figure 22. Load-pull power contours in dBm for a 100 W LDMOS device (MRF5S21100H) with an optimal impedance of $1.643 - j3.562\ \Omega$ at 2.14 GHz [48].

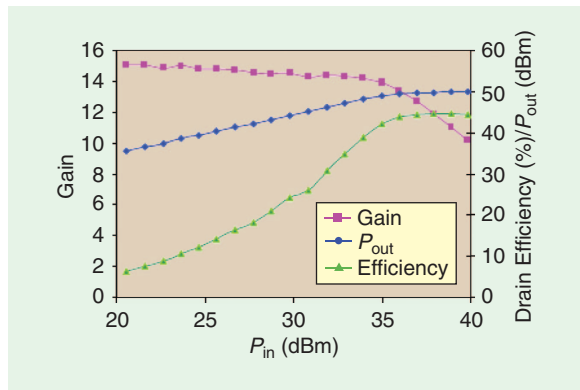


Figure 23. Measured output power (P_{out}), drain efficiency and gain versus input power (P_{in}) when load-pulling the device in Figure 22 at its optimal impedance, delivering over 100 W at a frequency of 2.14 GHz [48].

device. Furthermore, it should be noted that, according to (10), the required P_{LP} using the 50:7.15 impedance transformer is around 122 W for matching this specific 100 W device. However, in the hybrid setup of Figure 20, the required P_{LP} is only 35.5 W from the active load-pull source. The substantial reduction in P_{LP} is achieved by appropriately tuning the phase and magnitude of the system reflection coefficient, Γ_s , using the passive tuner in Figure 20.

Conclusion and Discussion

This article presented a review of the most common load-pull approaches adopted in the measurement and characterization of large-periphery microwave transistor devices. Although the quarter-wave transformer technique is simple and very cost-effective, it has serious limitations in terms of scalability, thereby necessitating a separate quarter-wave transformer for each device and frequency. Furthermore, due to the limited bandwidth of the quarter-wave transformer, its usage in harmonic load-pull applications is severely limited.

The state-of-the-art prematched load-pull setup is very versatile, rapid and capable of providing a highly reflective environment for harmonic load-pulling of high-power devices. The major limitations are its cost and its dependence on interpolation algorithms to synthesize a required reflection coefficient when it is not part of the calibration grid. Further, the prematched condition is device specific; therefore, the prematched tuner needs to be reset when the DUT is changed and, therefore, requires recalibration.

The broadband impedance transformer technique encompasses all of the features of the quarter-wave transformer technique. Additionally, it is readily applicable for harmonic load-pulling of high-power devices. The loop-enhanced load- and source-pull technique is capable of emulating highly reflective conditions, but is limited by the power rating of the components used in the measurement setup.

Overall, a hybrid setup consisting of passive and active load-pull components, as reported in [15] and discussed in this article, seems better suited for high reflection load-pull applications. The main issue that needs consideration in the hybrid system is the design and utilization of the associated components in an appropriate manner, so that the involved cost and complexity do not drastically increase. The recent commercial advancements in the form of high gamma tuners (HGT) [49] and multipurpose tuners (MPT) [50] appear to achieve high reflection coefficients at the fundamental and harmonics, but their potential use and the associated problems in high-power microwave device characterization remains to be investigated. In addition, it is important to note that the design and optimization of modern higher power PAs, which utilize large-periphery transistor devices, can be expedited by combining load-pull-generated data and a suitable nonlinear simulator [51]. This aspect, however, requires a thorough discussion and has been left for a future article.

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