A Unified Theory for Nonlinear Distortion Characteristics in Different Amplifier Technologies

This article presents a unified theory of power amplifier nonlinear distortion characteristics under small- and large-signal regimes for a wide range of active device technologies. It shows that some handy interactions between mild and strongly nonlinear operation, the so-called large-signal intermodulation distortion sweet spots, are inherent to a wide variety of power amplifier technologies such as Si MOSFETs, Si LDMOS, GaAs-AlGaAs HEMTs, GaAs MESFETs, Si BJTs and GaN HEMTs, justifying their use in the design of highly linear and efficient power amplifiers.

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RF power amplifiers (PA) are the last active blocks in any wireless transmitter system, handling the highest levels of RF signal and supply power. Linearization enforcing techniques relying on either adding external circuitry to the PA, or simply improving its design, are thus necessary. However, since the first set of methods — known as external linearization — involves several drawbacks like cost, size, effective bandwidth or difficulty of adjustment, there has been a growing interest in directly optimizing the actual PA linearity.

One possible way to achieve this design goal is to rely on certain bias points and power operating conditions, the so-called large-signal intermodulation distortion (IMD) sweet spots, which lead to improved carrier-to-IMD ratios (IMR) near the zones where power-added efficiency (PAE) is maximized. In an IMD versus input power level (P_in) plot, they can take many forms from a barely noticeable decrease in the slope to mild valleys, or even sharp dips in the IMD characteristic.

Using recent developments in the understanding of PA IMD under small- and large-signal regimes, this article shows that large-signal IMD sweet spots are not particular to a specific transistor or PA topology, but are inherent to a large variety of PA circuits and active device technologies such as Si MOSFETs, Si LDMOS, GaAs-AlGaAs HEMTs, GaAs MESFETs, Si BJTs and GaN HEMTs. Measured and simulated results are presented showing that these large-signal IMD sweet spots constitute a powerful and reliable means of PA linearization.

Power Amplifier Classes of Operation Revisited

PA IMD and efficiency vary dramatically with the amplifier’s operation class, traditionally defined with the help of the conduction angle concept, 2?, which expresses the percentage of the waveform period the device is on. This definition is based on an idealized piece-wise linear form of the active device’s transfer function (TF), which is the transformation of the known nonlinear bi-dimensional dependence of the output current, I_O(t), on the input and output control voltages, v_I(t) and v_O(t), I_O[v_I(t), v_O(t)], into a
one-dimensional model, $I_0[v(t)]$, assuming a determined output boundary condition imposed by the load impedance.

Using this traditional conduction angle concept, if $2\pi < 180^\circ$ the amplifier is said to be in class C, if $2\pi = 180^\circ$ it is in class B, if $180^\circ < 2\pi < 360^\circ$ in class AB and if $2\pi = 360^\circ$ the PA is said to operate in class A.

Besides the typical piece-wise approximation of the active device TF, Figure 1 illustrates the input voltage and output current waveforms for each of the previously mentioned operation classes.

![Fig. 1 Typical piece-wise approximation of an active device TF and the corresponding Vin and Iout for classes C, B, AB and A.](http://www.mwjournal.com/Journal/Print.asp?Id=AR_358)
in which \( G_1 \) is the device’s transconductance, herein assumed independent of bias or excitation amplitude. Unfortunately, this is an oversimplified model of operation because no actual device presents such a discontinuous behavior, as illustrated in Figure 2 (\( V_T \) is, in fact, undefined). The assumed linear zone still presents some residual nonlinearity, and cannot go on forever, but tends to saturate when \( v_I \) loses control over the output current, transferring it to \( v_O \) (a FET enters the triode region and a BJT or HBT enters in saturation).

Therefore, before starting any explanation of the IMD characteristics versus PA operation class, revisiting the definition of PA operation regimes is needed, keeping in mind these observed smooth TFs. Provided saturation of the TF is included, any increase in model detail will not be paid back in terms of the prediction of fundamental output power or power-added efficiency (the reason why this model has been left unquestioned for so long); however, it is possible to accurately model IMD, only when the TF’s soft turn-on is described. Moreover, it will also be shown that, despite the variability in nonlinear device models and the levels of detail being dealt with, a large range of device technologies share a very similar set of IMD characteristics.

The place to start is by comparing the two most important and distinct groups of active device technologies used in current microwave PAs: bipolar and field-effect transistors. If the TF characteristic of a bipolar device was given in terms of the dependence of collector current on base-emitter voltage, \( i_C[v_{BE}] \), it would be approximately exponential.\(^6\) This is in contrast with a FET whose drain-source current dependence on gate-source voltage, \( i_{DS}[v_{GS}] \), is only approximately exponential in the sub-threshold region, and then shows a quadratic zone near turn-on, which is further linearized due to non-uniform channel doping profile and short-channel effects.\(^7\) This situation changes dramatically, however, if the TF of the BJT is not given as \( i_C[v_{BE}] \) but as \( i_C[v_S] \), where \( v_S \) is no longer the intrinsic, but the extrinsic base-emitter control voltage. Because the voltage drop in the total series resistance of the base-emitter mesh (both base and emitter parasitic resistances and input generator internal impedance) is proportional to the base current (also an exponential function of intrinsic \( v_{BE} \)), the overall effect is an exponential TF near turn-on followed by a linearized characteristic imposed by the series resistance,\(^1\) which is much more similar to the FET’s TF. In fact, as illustrated previously, the resemblance between the TF curves originated from FETs or BJT devices is so evident that they can be approximated by the same global equivalent model.

In order to obtain an unambiguous and consistent definition of the various PA operation classes, a low order Volterra series of the output current of an active device will be used, or its memoryless subset, the Taylor series:
Figure 3 presents the variation of these three small-signal coefficients with the bias point for a real active device. The variation of $G_3$ with bias indicates that the small-signal third-order IMD (which is directly related to the PA’s third harmonic and gain compression or expansion) will change with bias point, not only in amplitude, but also in phase (the sign of $G_3$ in our memoryless nonlinearity).

By comparing the last two figures, one can conclude that a null in the $G_3(V_I)$ characteristic can be found close to the position of the ideal threshold voltage, $V_T$. Thus, biasing the PA at that point implies a null in the output third-order IMD. This result is consistent with the one obtained if the ideal PA was biased exactly at the break point of the piece-wise approximation, that is at $V_T$, originating the so-called linear (for odd-order distortion) class B PA.

This observation leads to the desired and more precise definition of a generalized cut-off voltage, and thus of PA operation classes, if the bias point of this $G_3$ null (the so-called small-signal IMD sweet spot) is taken as the ideal $V_T$. Class C would then be the operating regime of a PA biased below that bias point, class B would correspond to a PA biased exactly at the null, and classes A and AB would be the operating regimes of PAs biased above that point.

This refinement of PA operation class is still consistent with every other property of the circuit, as shown in Figure 4. As a matter of fact, this figure presents a comparison between the main PA characteristics, of DC power consumption ($I_{DC}$), output fundamental power ($I_{ds1}$), and second and third harmonic content ($I_{(R^2)}$ and $I_{J^2}$), obtained from three PA active device approximations: a FET-based PA ($I_{ds}$), a BJT ($I_c$)-based PA and the ideal piece-wise model ($I_p$). The similarity of the three curve families is obvious except for the region close to $2\pi = 360^\circ$ (class A). This is an indication that, contrary to the piece-wise linear model that only represents the devices’ strong nonlinearities, the actual devices also manifest mild nonlinearities. They still show some residual distortion even when operated in the ideally linear class A regime.

Large-signal IMD Sweet Spots

With the PA operation classes precisely defined this way, the focus can be on the large-signal IMD sweet
spots. They are critical points of the IMD versus $P_{in}$ characteristic, and can be explained as interactions between small- and large-signal nonlinear distortion, or between the device’s mild and strong nonlinearities. It is therefore convenient to study small- and large-signal nonlinear characteristics separately.

The small-signal IMD presents a rise of 3 dB/dB, its phase is determined by the TF local derivatives and can be controlled by changing the active device’s bias point. Thus, as seen in the plot of the coefficients of the Taylor expansion, the small-signal IMD, determined by the coefficients of Equation 1, can be either in phase with the fundamental (a symptom of small-signal gain expansion), in opposition (gain compression), or of null amplitude. This last situation would correspond to highly linear class A regimes, as previously reported in Pedro, et al.\(^8\)

Under large-signal operation, the nonlinear response is determined by the PA energy balance considerations. As the PA becomes short in supply power, the phase of the large-signal IMD sidebands tends to a constant value of 180°, describing the inevitable gain compression.\(^3\)

Now, three different scenarios, corresponding to the three discussed PA operation classes, are possible:

In the first one, the PA is biased for class C, in which small- and large-signal IMD phases are in opposition, as illustrated in Figure 5a. Thus, at the onset of PA saturation, the IMD must reverse its phase and there will be at least one IMD null (a large-signal IMD sweet spot), as depicted in Figure 5b.

![Fig. 5 Typical $P_{out}$ and IM3 versus $P_{in}$ characteristics for different small-signal and large-signal IMD phases. (scenario 1)](image-url)
In the second scenario, the PA is biased for class A. As seen in Figure 6a, small- and large-signal IMD phases are now coincident, and no large-signal sweet spot can occur (Figure 6b).

In the third and last scenario, the PA is biased for class AB, a more or less imprecise region of quiescent points just above the $G_3$ null. Despite the fact that small- and large-signal IMDs are still coincident, depending on the difference between the contribution of the positive lobe of $G_3$ and the negative one, it can be proved that a transition from 180° to 0° can occur at lower values of output power\(^4,5\) (see Figure 7a) generating an unexpected IMD sweet spot. Beyond this signal level, the IMD presents an opposite phase to the one imposed by the large-signal asymptote, and thus a new IMD sweet spot will have to appear at the onset of saturation. Thus, in this case, and depending on the PA quiescent point, two sweet spots can be generated (see Figure 7b). Fager, et al.\(^4,5\) give further details on the theoretical explanation of this behavior.
Simulated IMD Behavior for Different PA Technologies

In order to illustrate the ability of this analysis in describing IMD behavior in several PA technologies, various harmonic balance simulations of PAs biased for classes C, AB and A were performed. The models used were: BSIM3v3 model\(^9\) for Si MOSFET, Fager, et al.\(^4\) for Si LDMOS and GaN HEMTs, Angelov-Zirath\(^10\) for GaAs-AlGaAs HEMTs, Pedro\(^11\) for GaAs MESFETs and the Gummel-Poon\(^6\) for the Si BJTs.

Since each of the above mentioned PA technologies is analyzed for three different operation classes, the simulated IMD results will be presented in the form of carrier to IMD ratios (IMR) versus Pin for class C, AB and A, instead of the usual PIMD vs P\(_{in}\), since this enables a faster and more obvious comparison between them.

Si MOSFET

From Figure 8 it is possible to see that, for this Si MOSFET-based PA, a large-signal IMD sweet spot appears at class C, for high values of input power, while a double IMD sweet spot appears at class AB, and no sweet spot is visible in class A,\(^5\) as predicted.
As depicted in Figure 9, this Si LDMOS-based PA presents similar results to the ones shown by the Si MOSFET PA.

![Simulated IMR for Si LDMOS PA in three operation classes: A, AB, and C.](image)

**Fig. 9** Simulated IMR for Si LDMOS PA in three operation classes: A, AB, and C.

**GaAs-AlGaAs HEMT**

*Figure 10* shows the results for this GaAs-AlGaAs HEMT-based PA. These plots are similar to the ones already obtained for Si MOSFET and Si LDMOS.

![Simulated IMR for GaAs AlGaAs HEMT PA in three operation classes: A, AB, and C.](image)

**Fig. 10** Simulated IMR for GaAs AlGaAs HEMT PA in three operation classes: A, AB, and C.

**GaAs MESFET**

*Figure 11* shows the results obtained for this GaAs MESFET-based PA, in which IMR for classes A and C present the same aspect as seen before. However, class AB no longer has two peaks, but a rather smooth one. That slight increase in IMR at medium signal level regime can be attributed to an interaction between the negative $G_3$ and the positive higher orders’ contributions. Nevertheless, they were found not strong enough to generate the previous phase reversal, and thus neither a strong IMR maximum at medium signal excursions is visible, nor is there any large-signal IMD sweet spot.

![Simulated IMR for GaAs MESFET PA in three operation classes: A, AB, and C.](image)

**Fig. 11** Simulated IMR for GaAs MESFET PA in three operation classes: A, AB, and C.

**Si BJT**

As seen from *Figure 12*, the results obtained for the Si BJT-based PA are similar to the ones observed for the GaAs MESFET.
GaN HEMT
As seen from Figure 13, the results obtained for the GaN HEMT-based PA are again similar to the ones observed for the GaAs MESFET and Si BJT. As it is possible to see from the last six figures, class A presents the best small-signal linearity but, for high values of input power, IMR in classes AB and C is better than in class A. This fact, associated with the low gain and PAE recognized for microwave PAs biased in deep class C, justifies their use in class AB where optimized linearity and efficiency can be simultaneously obtained.

Measured IMD Behavior for Different PA Technologies
In order to provide experimental illustration of these simulated predictions Figures 14, 15 and 16 present measured results for two-tone IMD performance of Si CMOS, a GaAS MESFET and a GaN HEMT-based PAs in classes C, AB and A at 950 MHz, 2 GHz and 900 MHz, respectively. The experimental observations clearly support the simulated predictions shown previously for the corresponding PA technologies, validating the unified IMD theory presented above.
Conclusion

In this article, a unified theory of nonlinear distortion characteristics in different PA technologies was presented. It was shown that large-signal IMD sweet spots are inherent to a broad range of physical device characteristics, and that their presence is related with the PA classes of operation, which were, therefore, more precisely defined. The usefulness of these special IMD properties as a means to optimize the compromise between PA linearity and efficiency was also discussed.

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References


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