

Accurate and Rapid Measurement of IP_2 and IP_3

Ken Kundert

Designer's Guide Consulting, Inc.

Version 1b, 22 May 2002

Describes a new approach to computing IP_2 and IP_3 with SpectreRF that is both faster and more accurate than the traditional approach. Also given is a way of rapidly estimating CP_{1dB} .

Last updated on March 10, 2009. You can find the most recent version at www.designers-guide.org. Contact the author via e-mail at ken@designers-guide.com.

Permission to make copies, either paper or electronic, of this work for personal or classroom use is granted without fee provided that the copies are not made or distributed for profit or commercial advantage and that the copies are complete and unmodified. To distribute otherwise, to publish, to post on servers, or to distribute to lists, requires prior written permission.

1.0 Intermodulation Distortion

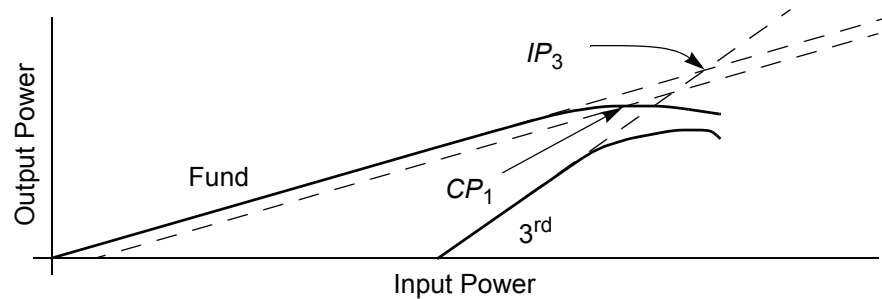
At low frequencies, it is common to quantify the nonlinearity of a circuit by indicating the distortion in the output signal when driven by a sinusoid that achieves a certain output level. At high frequencies, and particularly with narrowband circuits, it is more common to characterize the distortion produced by a circuit in terms of a compression point or an intercept point. These metrics characterize the circuit rather than the signal, and as such it is not necessary to specify the signal level at which the circuit was characterized.

2.0 Compression Points

To measure the compression point of a circuit, apply a sinusoid to its input and plot the output power of the fundamental as a function of the power of the input. The 1 dB compression point is the point where the gain of the circuit has dropped 1 dB from its small-signal asymptotic value. This is illustrated in Figure 1. iCP_{1dB} is the input power and oCP_{1dB} is the output power that corresponds to the 1 dB compression point. iCP is normally used for receivers and oCP for transmitters. Compression point is typically measured in dBm, which is decibels relative to one milliwatt.

FIGURE 1.

The 1 dB compression point (CP_1) is the point where the output power of the fundamental crosses the line that represents the output power extrapolated from small-signal conditions minus 1 dB. The 3rd order intercept point (IP_3) is the point where the third-order term as extrapolated from small-signal conditions crosses the extrapolated power of the fundamental.



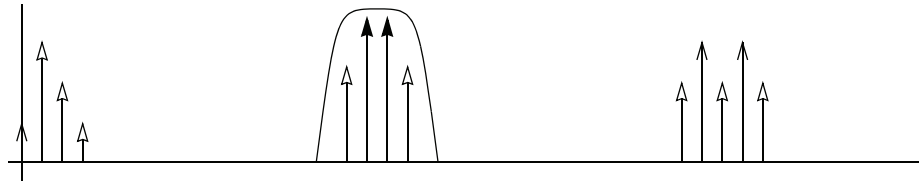
3.0 Intercept Points

In narrowband circuits distortion is commonly measured by applying two pure sinusoids with frequencies well within the bandwidth of the circuit. Call these frequencies f_1 and f_2 . The harmonics of these two frequencies would be outside the bandwidth of the circuit, however there are distortion products that fall at the frequencies $2f_1 - f_2$, $2f_2 - f_1$, $3f_1 - 2f_2$, $3f_2 - 2f_1$, etc. As shown in Figure 2 these frequencies should be within the bandwidth of the circuit and so can be used to measure the intermodulation distortion, or IMD, produced by the circuit.

The traditional approach to measuring a two-tone n^{th} -order IMD intercept point, IP_n , begins by applying two sinusoids to the circuit's input at f_1 and f_2 . Make the amplitude

FIGURE 2.

A narrowband circuit driven with two closely spaced sinusoidal tones (\longrightarrow) responds by generating harmonics (\longrightarrow) and intermodulation (\longrightarrow) tones. Distortion of the output signal results because several of the odd-order intermodulation tone fall within the bandwidth of the circuit.



of the two sinusoids the same, and increase their power while plotting the power at the output in a fundamental (either f_1 or f_2) and in an n^{th} -order intermodulation product (for IP_3 use either $2f_1 - f_2$ or $2f_2 - f_1$). This is illustrated for IP_3 in Figure 1. The n^{th} -order intercept point IP_n is defined in terms of the power levels of the fundamentals and the n^{th} -order products as extrapolated from their asymptotic small-signal behavior. When the input signal is small, a doubling of the input power doubles the fundamental output power and multiplies the output power of the n^{th} -order products by 2^n . Thus, the asymptotic slope of the fundamental is 1 dB/dB and the asymptotic slope of the n^{th} -order products is n dB/dB. The n^{th} -order intercept point (IP_n) is where the asymptotes for the n^{th} -order intermodulation product and the fundamental cross. iIP_n is the input power and oIP_n is the output power corresponding to the intercept point. They are generally measured in dBm. IP_3 is the most commonly used intercept point, but others, including IP_2 , IP_5 , and IP_7 , are also of interest.

In practice it is not necessary to sweep the input power to determine IP_n as long as you are confident that you have chosen an input power P_{in} that results in an accurate extrapolation. P_{in} should be chosen small enough that the output power of the first order term and the output power of the n^{th} -order term are both in their asymptotic ranges. And P_{in} should be chosen large enough so that they can be measured accurately. Once an appropriate P_{in} is applied, the output intercept point is computed with

$$IP_n = P + \frac{\Delta P}{n - 1} \quad (1)$$

where IP_n is the n^{th} -order intercept point (dBm), P is the power in the fundamental in dBm, and ΔP is the difference between the desired output signal and the undesired n^{th} -order output product in dB. P is the input power P_{in} if iIP_n is desired and it is the output power P_1 if oIP_n is desired [4].

With a two tone test, there are two different IP_3 results to be had, one computed from $|2f_1 \pm f_2|$ and the other from $|2f_2 \pm f_1|$. Generally, both results are the same. There are two things that would cause them to be different. First, if the tones are not well within the bandwidth of the circuit so that small-signal frequency response was different at the two intermodulation frequencies. Second, if the input powers at f_1 and f_2 are not matched, but this error can be avoided by following the directions given below.

3.1 Unequally Sized Test Tones

Tradition dictates that the same power be used for both of the fundamental tones when performing this test. Doing so gives the largest intermodulation terms and also allows

one to be a little sloppy in choosing the terms that make up P and ΔP in (1). However, as will be shown later for IP_2 and IP_3 , if you are careful when choosing the powers used for P and ΔP , the actual result is not sensitive to differences in the power levels of the two input tones. This fact will be used to construct a method of computing IP_n that is faster and more accurate when using an RF simulator than the traditional approach. The rules for choosing which powers are used in P and ΔP follow. When computing IP_2 , if P is the power at f_1 , then ΔP should be the difference (in dB) between the power at f_2 and the power at $|f_1 \pm f_2|$. Or, if P is the power at f_2 , then ΔP should be the difference (in dB) between the power at f_1 and the power at $|f_2 \pm f_1|$. Similarly, when computing IP_3 , if P is the power at f_1 , then ΔP should be the difference (in dB) between the power at f_2 and the power at $|2f_1 - f_2|$. Or, if P is the power at f_2 , then ΔP should be the difference (in dB) between the power at f_1 and the power at $|2f_2 - f_1|$. These rules are summarized in Table 1 and Table 2.

Generally, if the power in f_1 is significantly larger than the power in f_2 , then the power at $|2f_1 - f_2|$ will be significantly larger than the power at $|2f_2 - f_1|$. Being at lower power means that the signal is closer to the noise floor and so is determined less accurately, and so its use should be avoided by making the power at f_1 be P . Similarly, if the power at f_2 is larger than that at f_1 , then it should be chosen to be P .

TABLE 1.

Formulas for computing IP_2 . p_{i1} and p_{i2} represent the input power for the tones at f_1 and f_2 , p_{o1} and p_{o2} represent the output power for the tones at f_1 and f_2 , and p_{o12} represents the output power for the tone at $|f_2 \pm f_1|$.

Large Tone	iIP_2	oIP_2
f_1	$iIP_2 = p_{i1} + (p_{o2} - p_{o12})$	$oIP_2 = p_{o1} + (p_{o2} - p_{o12})$
f_2	$iIP_2 = p_{i2} + (p_{o1} - p_{o12})$	$oIP_2 = p_{o2} + (p_{o1} - p_{o12})$

TABLE 2.

Formulas for computing IP_3 . p_{i1} and p_{i2} represent the input power for the tones at f_1 and f_2 , p_{o1} and p_{o2} represent the output power for the tones at f_1 and f_2 , and p_{o12} and p_{o21} represent the output power for the tones at $|2f_1 - f_2|$ and $|2f_2 - f_1|$.

Large Tone	iIP_3	oIP_3
f_1	$iIP_3 = p_{i1} + \frac{p_{o2} - p_{o12}}{2}$	$oIP_3 = p_{o1} + \frac{p_{o2} - p_{o12}}{2}$
f_2	$iIP_3 = p_{i2} + \frac{p_{o1} - p_{o21}}{2}$	$oIP_3 = p_{o2} + \frac{p_{o1} - p_{o21}}{2}$

4.0 Derivation

This section derives IP_2 , IP_3 , and an approximation of CP_1 for a circuit assuming that its nonlinear transfer characteristics can be described with a simple power series,

$$x = au + bu^2 + cu^3 + \dots, \tag{2}$$

where u is the input signal and x is the output signal. While this model does not include the dynamic effects common in most circuits, including such effects would not change the result and would make the analysis needlessly more complex.

These derivations will show that when (1) is applied properly, the values computed for IP_2 and IP_3 are independent of the level of the input signals applied during the test.

4.1 IP_2

When computing IP_2 , only the first two terms of (2) need be considered.

$$x = au + bu^2. \quad (3)$$

Apply a two-tone test signal,

$$u(t) = \alpha \cos \omega_1 t + \beta \cos \omega_2 t. \quad (4)$$

Then

$$x(t) = a(\alpha \cos \omega_1 t + \beta \cos \omega_2 t) + b(\alpha \cos \omega_1 t + \beta \cos \omega_2 t)^2, \quad (5)$$

$$x(t) = a(\alpha \cos \omega_1 t + \beta \cos \omega_2 t) + b(\alpha^2 \cos^2 \omega_1 t + 2\alpha\beta \cos \omega_1 t \cos \omega_2 t + \beta^2 \cos^2 \omega_2 t). \quad (6)$$

The following trigonometric identities will be used to separate the equation into terms associated with individual frequencies.

$$\cos^2 \phi = \frac{1}{2}(1 + \cos 2\phi) \quad (7)$$

$$\cos \phi_1 \cos \phi_2 = \frac{1}{2}(\cos(\phi_1 + \phi_2) + \cos(\phi_1 - \phi_2)) \quad (8)$$

Now (6) is rewritten as

$$x(t) = a(\alpha \cos \omega_1 t + \beta \cos \omega_2 t) + b\left(\frac{\alpha^2}{2}(1 + \cos 2\omega_1 t) + \alpha\beta \cos(\omega_1 + \omega_2)t + \alpha\beta \cos(\omega_1 - \omega_2)t + \frac{\beta^2}{2}(1 + \cos 2\omega_2 t)\right). \quad (9)$$

Define $p_{o1} = (a\alpha)^2$ to be the output power at ω_1 , $p_{o2} = (a\beta)^2$ to be the output power at ω_2 , and $p_{o12} = (b\alpha\beta)^2$ to be the output power at $|\omega_1 \pm \omega_2|$. Now apply (1) with $n = 2$ to compute oIP_2 .

$$oIP_2 = \text{dB}_{10}(p_{o1}) + \text{dB}_{10}(p_{o2}) - \text{dB}_{10}(p_{o12}), \quad (10)$$

$$oIP_2 = \text{dB}_{10}(p_{o1}p_{o2}/p_{o12}), \quad (11)$$

$$oIP_2 = \text{dB}_{10}((a\alpha)^2(a\beta)^2/(b\alpha\beta)^2), \quad (12)$$

$$oIP_2 = \text{dB}_{20}(a^2/b). \quad (13)$$

To instead compute iIP_2 use (1) with $p_{i1} = \alpha^2$ as the input power at ω_1 , then

$$iIP_2 = \text{dB}_{20}(a/b). \quad (14)$$

Notice that both oIP_2 and iIP_2 are completely independent of α and β , the amplitudes of the two test tones. This, of course, is true in theory only. In practice, at least one of the tones must be large enough so that the resulting intermodulation tone is accurately measurable. However, there is no requirement that the two input tones be of the same size, or anywhere near the same size.

4.2 IP_3

When computing IP_3 , only the first two odd-order terms of (2) need be considered.

$$x = au + cu^3. \quad (15)$$

Apply a two-tone test signal,

$$u(t) = \alpha \cos \omega_1 t + \beta \cos \omega_2 t \quad (16)$$

Then

$$x(t) = a(\alpha \cos \omega_1 t + \beta \cos \omega_2 t) + c(\alpha \cos \omega_1 t + \beta \cos \omega_2 t)^3, \quad (17)$$

$$x(t) = a(\alpha \cos \omega_1 t + \beta \cos \omega_2 t) + \quad (18)$$

$$c(\alpha^2 \cos^2 \omega_1 t + 2\alpha\beta \cos \omega_1 t \cos \omega_2 t + \beta^2 \cos^2 \omega_2 t)(\alpha \cos \omega_1 t + \beta \cos \omega_2 t),$$

$$x(t) = a(\alpha \cos \omega_1 t + \beta \cos \omega_2 t) + \quad (19)$$

$$c(\alpha^3 \cos^3 \omega_1 t + 3\alpha^2\beta \cos^2 \omega_1 t \cos \omega_2 t + 3\alpha\beta^2 \cos \omega_1 t \cos^2 \omega_2 t + \beta^3 \cos^3 \omega_2 t).$$

The following trigonometric identities will be used to separate the equation into terms associated with individual frequencies.

$$\cos^2 \phi = \frac{1}{2}(1 + \cos 2\phi), \quad (20)$$

$$\cos^3 \phi = \frac{1}{4}(3 \cos \phi + \cos 3\phi), \quad (21)$$

$$\cos \phi_1 \cos \phi_2 = \frac{1}{2}(\cos(\phi_1 + \phi_2) + \cos(\phi_1 - \phi_2)). \quad (22)$$

Now (19) is rewritten as

$$x(t) = a(\alpha \cos \omega_1 t + \beta \cos \omega_2 t) + \quad (23)$$

$$c\left(\frac{\alpha^3}{4}(3 \cos \omega_1 t + \cos 3\omega_1 t) + \frac{3\alpha^2\beta}{2}(1 + \cos 2\omega_1 t) \cos \omega_2 t + \frac{3\alpha\beta^2}{2} \cos \omega_1 t (1 + \cos 2\omega_2 t) + \frac{\beta^3}{4}(3 \cos \omega_2 t + \cos 3\omega_2 t)\right),$$

$$x(t) = a(\alpha \cos \omega_1 t + \beta \cos \omega_2 t) + \quad (24)$$

$$c\left(\frac{\alpha^3}{4}(3 \cos \omega_1 t + \cos 3\omega_1 t) + \frac{\beta^3}{4}(3 \cos \omega_2 t + \cos 3\omega_2 t) + \frac{3\alpha^2\beta}{2}\left(\cos \omega_2 t + \frac{1}{2}(\cos(2\omega_1 t + \omega_2 t) + \cos(2\omega_1 t - \omega_2 t))\right) + \frac{3\alpha\beta^2}{2}\left(\cos \omega_1 t + \frac{1}{2}(\cos(2\omega_2 t + \omega_1 t) + \cos(2\omega_2 t - \omega_1 t))\right)\right),$$

$$x(t) = \left(a\alpha + \frac{3c\alpha^3}{4} + \frac{3c\alpha\beta^2}{2}\right) \cos \omega_1 t + \frac{c\alpha^3}{4} \cos 3\omega_1 t + \quad (25)$$

$$\left(a\beta + \frac{3c\alpha^2\beta}{2} + \frac{3c\beta^3}{4}\right) \cos \omega_2 t + \frac{c\beta^3}{4} \cos 3\omega_2 t +$$

$$\frac{3c\alpha^2\beta}{4}(\cos(2\omega_1 t + \omega_2 t) + \cos(2\omega_1 t - \omega_2 t)) +$$

$$\frac{3c\alpha\beta^2}{4}(\cos(2\omega_2t + \omega_1t) + \cos(2\omega_2t - \omega_1t)).$$

Define $p_{o1} = (a\alpha)^2$ to be the output power at ω_1 , $p_{o2} = (a\beta)^2$ to be the output power at ω_2 , and $p_{o12} = (3c\alpha^2\beta/4)^2$ to be the output power at $|2\omega_1 \pm \omega_2|$. Now apply (1) with $n = 3$ to compute oIP_3 .

$$oIP_3 = \text{dB}_{10}(p_{o1}) + \frac{\text{dB}_{10}(p_{o2}) - \text{dB}_{10}(p_{o12})}{2}, \quad (26)$$

$$oIP_3 = \text{dB}_{10}\left(p_{o1} \sqrt{\frac{p_{o2}}{p_{o12}}}\right), \quad (27)$$

$$oIP_3 = \text{dB}_{10}\left((a\alpha)^2 \sqrt{\frac{(a\beta)^2}{(3c\alpha^2\beta/4)^2}}\right), \quad (28)$$

$$oIP_3 = \text{dB}_{20}\left(\sqrt{\frac{4a^3}{3c}}\right). \quad (29)$$

To instead compute iIP_3 use (1) with $p_{i1} = \alpha^2$ as the input power at ω_1 , then

$$iIP_3 = \text{dB}_{20}\left(\sqrt{\frac{4a}{3c}}\right). \quad (30)$$

As before, notice that both oIP_3 and iIP_3 are completely independent of α and β , the amplitudes of the two test tones. So again, there is no requirement that the two input tones be of the same size.

4.3 CP_1

As a side effect of the above analysis, it is possible to come up with a quick estimate of CP_1 . To do so, start by setting $\beta = 0$ and rewriting (25).

$$x(t) = \left(a\alpha + \frac{3c\alpha^3}{4}\right)\cos\omega_1t + \frac{3c\alpha^3}{4}\cos 3\omega_1t \quad (31)$$

Consider only the fundamental terms. At low power levels (small values of α) the $a\alpha$ term dominates. As the input power increases, the $3c\alpha^3/4$ term becomes more significant and the amplifier starts to compress. When the sum of the two terms is 1 dB below the first term alone, we have reached the 1 dB compression point (assuming that the effect of higher-order terms is not significant). Thus, $iCP_{1\text{dB}}$ is the value of α , referred to as α_{CP} , that satisfies

$$10^{-1/20}a\alpha_{\text{CP}} = a\alpha_{\text{CP}} + \frac{3c\alpha_{\text{CP}}^3}{4}, \quad (32)$$

$$\alpha_{\text{CP}} = \sqrt{\frac{4a}{3c}}\sqrt{1 - 10^{-1/20}}. \quad (33)$$

From (30)

$$iCP_{1\text{dB}} = iIP_3 + \text{dB}_{20}(\sqrt{1 - 10^{-1/20}}) \quad (34)$$

From this it can be seen that in the case where compression point is determined largely by 3rd order distortion, which is usually the case,

$$iCP_{1dB} = iP_3 - 9.6 \text{ dB} . \quad (35)$$

Thus, if you know iP_3 , and a rapid way of computing iP_3 will be given below, then you can easily estimate iCP_{1dB} . This estimate ignores the effect of higher-order terms, and so can be a bit on the high side.

5.0 Measuring IP_n With a Simulator

5.1 Using SPICE

With a circuit simulator, such as SPICE, it is difficult to measure IP_n for several reasons. First, a Fourier analysis is needed to resolve the various components in the output, which requires the output signal to be periodic. For this reason, both input tones must share a common period. Generally, the tones are very close in frequency, implying that the common period will be quite long relative to the period of either tone. This implies that a transient simulation will be expensive. Second, the accuracy of a SPICE Fourier analysis degrades at higher harmonics. This is problematic because the two input tones are closely spaced and so the harmonics of interest in the output will have large harmonic numbers.

With SPICE, there is no alternative. The best you can do is to choose $\Delta f = |f_1 - f_2|$ as large as possible, taking care to assure that the output intermodulation products remain in band. In this case, it is best to set the input power in both tones to be the same because it costs nothing and because it maximizes the size of the intermodulation tones, and so reduces error.

5.2 Using an RF Simulator's Large-Signal Analysis

RF simulators, such as a harmonic balance simulator or Spectre®RF¹, provide large signal quasiperiodic steady-state analyses that do not suffer from the two problems listed above. However, the simulations can still be expensive due to the large number of tones involved, particularly if you are trying to compute IP_n of circuits like mixers that require a third large tone, the LO, to operate properly.

With this case it is again best to set the input power in both tones to be the same, and for the same reason.

5.3 Using an RF Simulator's Small-Signal Analysis

RF simulators also generally provide a way of combining a small signal analysis, such as a periodic or quasiperiodic AC analysis, PAC or QPAC [2, 3], and a large signal periodic or quasiperiodic steady-state analysis, PSS or QPSS. In this case, it is possible to exploit the fact that the tones need not be the same size. The power of one of the tones can be reduced to the point where it is considered small, and so can be handled by the

1. Spectre is a registered trademark of Cadence Design Systems.

separate small-signal analysis. In this way the number of large tones that need be included in the large signal analysis can be reduced by one, which considerably speeds up the simulation and improves accuracy. The additional cost of the small-signal analysis is more than made up for by the time savings in the large signal analysis. It is more accurate because many of the error mechanisms that are present in the large-signal analysis are missing from the small-signal analysis. Another benefit is that unlike with a purely transient or PSS based approach, the frequency of the small tone becomes unconstrained. In particular, it is not necessary to choose the frequencies so that they are commensurate, and the time required for the simulation does not suffer with a poor choice of frequencies.

This approach is not completely general. It is only capable of computing intercept points for those intermodulation products that fall at the frequencies $(n-1)f_1 \pm f_2$, a restriction that in practice makes this approach useful only for computing IP_2 and IP_3 .

The large tone (f_1) drives the circuit into a nonlinear region. The small tone (f_2) is needed only to produce an in-band response by mixing with the distortion products generated by the first tone. To compute IP_2 or IP_3 of an amplifier using this approach, apply the first tone perform and perform the large-signal PSS analysis. Identify P in (1) (in dBm). This is either the input or output power at f_1 depending on whether iIP_n or oIP_n is desired. Then, apply the second tone as a small complex exponential at a sideband of the first tone and use a PAC analysis to compute the terms that make up ΔP (in dB), the ratio between the power at the output in the fundamental frequency of the small tone (f_2) and the power at the output in the n^{th} -order intermodulation product. IP_n is then computed with (1).

The intermodulation distortion of a clocked circuit such as mixer is measured in a similar manner, except the both the LO and the first tone are applied together and the response computed with a QPSS analysis. The response to the second tone is computed using a QPAC analysis.

In practice this approach is limited to computing IP_2 and IP_3 , but given that restriction, it can be used in the same situations as the traditional approach and gets the same answer, but is considerably faster and is in general more accurate.

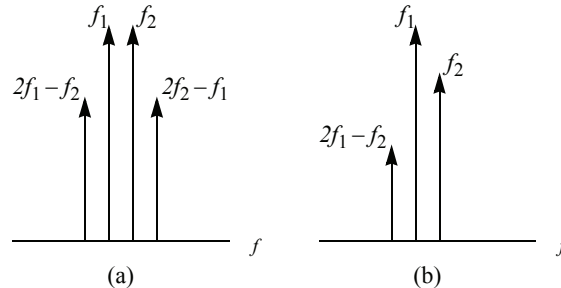
Qualitative Explanation of the Results. In a traditional IP_3 measurement, the response generally takes the form shown in Figure 3a. In this case, large tones are applied at both f_1 and f_2 and the circuit responds with intermodulation tones at $2f_1 - f_2$ and $2f_2 - f_1$. The intermodulation tone at $2f_1 - f_2$ results because the tone at f_1 modulates the characteristics of the circuit, which causes any signal at f_2 to mix down to $2f_1 - f_2$. As long as the tone at f_2 is small enough so that it does not cause significant compression, the mixing process is linear, and so the amplitude ratio between the responses at f_2 and at $2f_1 - f_2$ is constant relative to the amplitude of the tone at f_2 . Since it is the ratio (or the difference when working in dB) that is used in (1), the size of the tone can be reduced, as shown in Figure 3b, without changing the result.

The same arguments can be made for the response at $2f_2 - f_1$ relative to the tone at f_1 .

This explanation rests on the assumption that the large tone is not so large that it causes significant compression. This assumption is justified because IP_n measurements are, by definition, measured where power levels are too small to cause significant compression. If the levels were large enough to cause compression, then you would be outside the asymptotic range for both the fundamentals and the intermodulation terms.

FIGURE 3.

Response of a circuit to a traditional two-tone IP_3 test (a), and to a rapid IP_3 test (b).



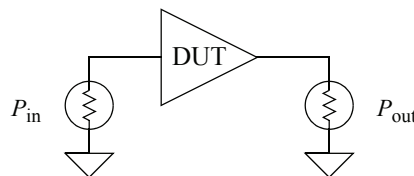
6.0 LNA Example

The process for performing a rapid calculation of IP_3 on an LNA using SpectreRF from within Artist is given below. It assumes that you have performed a traditional compression point analysis at least once and found iCP_{1dB} . This analysis would not need to be repeated unless a change were made to the circuit that significantly changed its compression point.

This example assumes that you are using the test bench shown in Figure 4. It consists of the device-under-test (DUT), a low-noise amplifier (LNA), that is connected to ports at both its input and its output. A port is a primitive component built into Spectre that combines a voltage source and a constant-valued resistor. It is used in this test bench to provide both the source and the load to the LNA. When acting as a load, the voltage source is disabled. In this application using port at the input is helpful because ports allow their output power to be specified in dBm. However, it is important to realize that the specified amount of power is only delivered to the input of the amplifier if the input match is perfect. If the match is not perfect, then some of the specified input power will reflect back from the input of the amplifier.

FIGURE 4.

Test bench for an LNA.



Procedure. Assume that for our hypothetical LNA the $iCP_{1dB} = -10$ dBm, that $f_1 = 900$ MHz and that $f_2 = 901$ MHz.

1. Specify the input power, the input frequency, and the waveshape of the source port. The waveshape is set to sinusoid with `type=sine` and the frequency is set to f_1 with `freq=900MHz`. Choose an input power that is at between 20-40 dB below iCP_{1dB} . In this case, since $iCP_{1dB} = -10$ dBm, the input power level is set to `dbm=-30dBm`.

2. Set the PAC small-signal amplitude of the source to a non-zero value, the actual value used is not important. Typical values are `pacmag=1` or `pacdbm=-30`. The first sets the small-signal amplitude to 1 Vp, and the other sets it to -30 dBm.
3. Enable the PSS analysis and assure that the fundamental frequency is set to 900 MHz. The accuracy defaults (`errpreset`) should be set to moderate in this case because the accuracy demands of this application are modest.
4. Enable the PAC analysis and configure it to run a single point analysis with the frequency set to $f_2 = 901$ MHz and the maximum number of sidebands set to at least 2. There is no reason to use more than 2 if you are only interested in computing IP_3 .
5. Run the simulation assuring that both the PSS and PAC analyses run without errors.
6. Open the direct plot form for the PAC analysis.
7. Choose *IPN Curves*.
8. Specify that the load as being a port.
9. Specify the single point input power as being the large-signal source power, the same value that was specified in step 1, in this case -30 dBm.
10. For IP_3 , specify the order to be 3rd.
11. Choose whether the input or output-referred intercept point is desired.
12. Specify first and third sidebands, where the first (f_2) is 901 MHz and the third ($2f_1 - f_2$) is 899 MHz.
13. Select the output port to initiate the display of the intercept point and read the resulting value off the display.

7.0 Mixer Example

Here the process for performing a rapid calculation of IP_3 on a down conversion mixer using SpectreRF from within Artist is given. The testbench used is similar to the one in Figure 4, except a mixer is used as the device under test (DUT).

Procedure. Assume that for our hypothetical mixer the $iCP_{1dB} = -10$ dBm, that $f_1 = 840$ MHz and $f_2 = 850$ MHz, and that $f_{LO} = 780$ MHz.

1. Specify the fundamental names on the sources for the LO and f_1 . In particular, specify the fundamental name for the LO to be “lo” and for f_1 to be “fund1”.
2. Specify the input power, the input frequency, and the waveshape of the source port. The waveshape is set to sinusoid with `type=sine` and the frequency is set to f_1 with `freq=840MHz`. Choose an input power that is at between 20-40 dB below iCP_{1dB} . In this case, since $iCP_{1dB} = -10$ dBm, the input power level is set to `dbm=-30dBm`.
3. Set the small-signal amplitude of the source to a non-zero value, the actual value used is not important. Typical values are `pacmag=1` or `pacdbm=-30`. The first sets the small-signal amplitude to 1 Vp, and the other sets it to -30 dBm.
4. Enable the QPSS analysis and assure that the two fundamental frequency appear in the *Fundamental Tones* section of the form (Click *Update From Schematic* if they need updating) as shown in Figure 5. Modify the *lo* tone so that it’s signal size is set to *Large* and assure the *fund1* tone is set to *Moderate*. The accuracy defaults (`errpreset`) should be set to moderate in this case because the accuracy demands of this application are modest. You can specify a *stabilization time* if you need the simula-

tion to progress to a certain point in time in order to operate properly (this is unusual) or as a way of avoiding convergence difficulties in QPSS (more common). Uncheck the *Sweep* box.

FIGURE 5.

Example of what the QPSS analysis form might look like when finding the mixer IP_3 .

The screenshot shows a software interface for QPSS analysis. It features a table titled 'Fundamental Tones' with columns for '#', 'Name', 'Expr', 'Value', 'Signal', 'SrcId', and 'Harms'. The table contains two rows: one for 'Fund1' at 840M and one for 'LO' at 780M. Below the table are input fields for 'LO', '780M', '780M', 'Large', and 'VLo', along with buttons for 'Clear/Add', 'Delete', and 'Update From Schematic'. There are also checkboxes for 'View Harmonics', 'Sweep', and 'Save Initial Transient Results (saveinit)', and radio buttons for 'Accuracy Defaults (errpreset)' set to 'moderate'.

#	Name	Expr	Value	Signal	SrcId	Harms
1	Fund1	840M	840M	Moderate	Pin	3
3	LO	780M	780M	Large	VLo	3

5. Enable the QPAC analysis and configure it to run a single point analysis with the frequency set to $f_2 = 850$ MHz and the sidebands set to at least *Maximum Clock Order* = 2. There is no reason to use more than 2 if you are only interested in computing IP_3 .
6. Run the simulation assuring that both the QPSS and QPAC analyses run without errors.
7. Open the direct plot form for the QPAC analysis.
8. Choose *IPN Curves*.
9. Specify that the load as being a port.
10. Specify the single point input power as being the large-signal source power, the same value that was specified in step 1, in this case -30 dBm.
11. For IP_3 , specify the order to be 3rd.
12. Choose whether the input or output-referred intercept point is desired.
13. Specify first and third sidebands, where the first ($|f_{LO} - f_2|$) is 70 MHz and the third ($|f_{LO} - (2f_1 - f_2)|$) is 50 MHz, as shown in Figure 6.
14. Select the output port to initiate the display of the intercept point and read the resulting value off the display.

8.0 If You Have Questions

If you have questions about what you have just read, feel free to post them on the *Forum* section of *The Designer's Guide Community* website. Do so by going to www.designers-guide.org/Forum.

FIGURE 6.

Example of what the QPAC plot form might look like when finding the mixer IP_3 .

Function

Voltage Current
 IPN Curves

Select: Port (fixed R(port))

Single Point Input Power Value (dBm): -30

Input Referred IP3: Order: 3rd

	Freq. (Hz)	LO	Fund1
3rd Order Harmonic	10M	0	-1
	50M	1	-2
	70M	-1	0
	110M	2	-3
	130M	-2	1
	710M	-2	0
1st Order Harmonic	10M	0	-1
	50M	1	-2
	70M	-1	0
	110M	2	-3
	130M	-2	1
	710M	-2	0

Add To Outputs: Replot

I would like to thank Mark Reichelt for pointing out several errors in (25).

9.0 References

- [1] K. S. Kundert. *The Designer's Guide to SPICE and Spectre*. Kluwer Academic Publishers, 1995.
- [2] Ken Kundert. Introduction to RF simulation and its application. *Journal of Solid-State Circuits*, vol. 34, no. 9, September 1999. Also available from www.designers-guide.org/Analysis.
- [3] R. Telichevesky, K. Kundert and J. White. Efficient steady-state analysis based on matrix-free Krylov-subspace methods. *Proceedings of the 32nd Design Automation Conference*, June 1995.
- [4] P. Vizmuller. *RF Design Guide: Systems, Circuits and Equations*. Artech House, 1995.