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IP2 and IP3 Nonlinearity Specifications for 3G/WCDMA Receivers

By Chris W. Liu and Morten Damgaard
Broadcom Corporation

Here is a thorough summary of linearity requirements for WCDMA receivers, with special attention to the way they apply to the direct conversion architecture

In 3G/WCDMA mobile handsets, the direct conversion receiver (DCR) architecture is commonly used due to its simplicity and low cost, where interstage filters are completely eliminated. However, with less filtering, the linearity requirement for a direct conversion receiver is very critical. It is essential to define system linearity requirements properly for the receiver to meet the performance requirements.

In FDD mode, the transmitter and receiver are continuously working at the same time, and the transmitted signal leaks into the receiver due to the limited TX-to-RX isolation of the duplexer. The transmitter is possibly the strongest interferer for the companion receiver in the handset, and poses the most stringent linearity requirement for the receiver.

I. 3GPP Test Cases

A number of test cases are specified in 3GPP for a WCDMA receiver, and each test case has different test conditions. Therefore nonlinearity performance requirements are needed of the receiver in each case. The related test cases are summarized in Tables A-F and used throughout the following discussion.

These test cases require the receiver to meet certain BER performance with these defined interferers. For RF system design, the requirements have to be interpreted into RF parameters such as noise figure (NF), compression point (P_{1dB}), second order intercept point (IP2), third order intercept point (IP3), and so forth.

	Test Conditions	
Band	DPCH_Ec	<REFÎor>
I	-117	-106.7
II	-115	-104.7
III	-114	-103.7
V	-115	-104.7
VIII	-114	-103.7
TX power: 24 dBm		

Table A · Sensitivity requirement.

	Test Conditions	
Parameter	**Case 1**	**Case 2**
DPCH_Ec	<REFSENS> + 14 dB	<REFSENS> + 41 dB
Îor	<REFÎor> + 14 dB	<REFÎor> + 41 dB
I_{oac}	-52	-25
F_{uw}	+5 or -5	+5 or -5
Tx power: 20 dBm		

Table B · Adjacent channel blocking.

II. Second Order Nonlinearity

The second order nonlinearity of the receiver will square the modulated blocker signal, such as the TX leakage signal, producing DC and low frequency components which fall into the receive band of the direct conversion receiver. The AM (amplitude modulated) signal is demodulated into the RX channel with twice bandwidth of the original interferers. Moreover, a strong blocking signal will also intermodulate due to second order nonlinearity with the TX leakage signal to create a TX image which can fall into the band.

Mathematic Formula of IIP2

In general, the output signal of a nonlinear system can be described as follows:

$$V_o = a_1 V_i + a_2 V_i^2 + a_3 V_i^3 + \ldots$$

where V_o is the output voltage and V_i is the

Wideband		Narrow band	
Parameter	Test Conditions	Band II, V	Band III, VIII
DPCH_Ec	<REFSENS> + 3 dB	<REFSENS> + 10 dB	
Îor	<REFÎor> + 3 dB	<REFÎor> + 10 dB	
I_{ouw1} (CW)	-46	-44	-43
I_{ouw2} mean power (modulated)	-46	-44	-43
F_{uw1} (offset)	± 10	± 3.5	± 3.6
F_{uw2} (offset)	± 20	± 5.9	± 6.0
Tx power: 20 dBm			

Table C · Intermodulation (wideband and narrowband).

Parameter	Unit	Test Conditions	
		Band II, V	Band III, VIII
DPCH_Ec	dBm/3.84	<REFSENS> + 10 dB	
Îor	dBm/3.84	<REFÎor> + 10 dB	
$I_{blocking}$	dBm	-57	-56
F_{uw} offset	MHz	± 2.7	± 2.8
Tx power: 20 dBm			

Table E · Narrow band blocking (NBB).

Parameter	Unit	Freq range 1	Freq range 2	Freq range 3
DPCH_Ec	dBm/3.84	<REFSENS>+3 dB	<REFSENS>+3 dB	<REFSENS>+3 dB
$\hat{I}_{or}$	dBm/3.84	<REF$\hat{I}_{or}$> + 3 dB	<REF$\hat{I}_{or}$> + 3 dB	<REF$\hat{I}_{or}$> + 3 dB
$I_{blocking}$ (CW)	dBm	-44	-30	-15
F_{uw} (Band I)	MHz	2050<f <2095 2185<f <2230	2025 <f ≤2050 2230 ≤f <2255	1< f ≤2025 2255≤f<12750
Tx power: 20 dBm				

Table D · Out of band blocker (OOB).

Parameter	Unit	Test Conditions	
DPCH_Ec	dBm/3.84	<REFSENS> + 3 dB	
Îor	dBm/3.84	<REFÎor> + 3 dB	
$I_{blocking}$	dBm	-56	-44
F_{uw} offset	MHz	± 10	± 15
Tx power: 20 dBm			

Table F · In band blocking (IBB).

input voltage.

If using a traditional two tone signal as input V_i, we have:

$$V_i = A \cdot \cos(\omega_1 t) + A \cdot \cos(\omega_2 t)$$

The second order component of the output is:

$$\begin{aligned} V_o &= a_2 V_i^2 \\ &= a_2 A^2 \left(\cos^2(\omega_1 t) + \cos^2(\omega_2 t) + 2\cos(\omega_1 t)\cos(\omega_2 t)\right) \\ &= a_2 A^2 \left(1 + \frac{1}{2}\cos(2\omega_1 t) + \frac{1}{2}\cos(2\omega_2 t) + \cos((\omega_1+\omega_2)t) + \cos((\omega_1-\omega_2)t)\right) \end{aligned}$$

It shows that the second order IM2 products are created at three frequencies: DC, $f1 + f2$ and $f1 – f2$. In terms of power level, IM2 products are distributed against total IM2 power as:

- 50% (–3 dB) at DC
- 25% (–6 dB) at $f1 + f2$
- 25% (–6 dB) at $f1 – f2$

The IM2 of low frequency is considered only in this paper since it is the one falling into band. The power level of the IM2 product at $f1 – f2$ is 25% of the total IM2 power which is 6 dB below the total IM2 power. So the power level of IM2 at low frequency ($f2 – f1$) can be expressed as:

$$P_{im2}(\text{dBm}) = 2 \cdot P_{in} - \text{IIP2} - 6 \text{ dB} \quad (1)$$

IIP2 With WCDMA TX Uplink Leakage

If the signal is AM modulated, such as TX leakage signal, the 2 tone formula above can not be fully applied. The difference between the formula and 2 tone signal is called correction factor [4]. The following section derives the correction factor for a WCDMA signal using ADS simulation (Fig. 1).

The two tone test case and the case of single tone plus WCDMA uplink signal test case are simulated. The coefficients of the second order product in the model is set to 0.1 for simplicity since only the difference between using the two tone measurement and the TX uplink modulated signal is needed. The two tone simulation results are shown in Figure 2 and the simulation results with TX uplink signal are shown in Figure 3.

With the TX uplink signal, since it is modulated, the low frequency IM2 products are measured by integrating the power from 1 kHz to 2.0 MHz in the frequency domain. The difference of IM2 between the two tone and modulated signal is 9.72 dB. The IP2 two-tone formula for TX leakage case with correction factor is as follows:

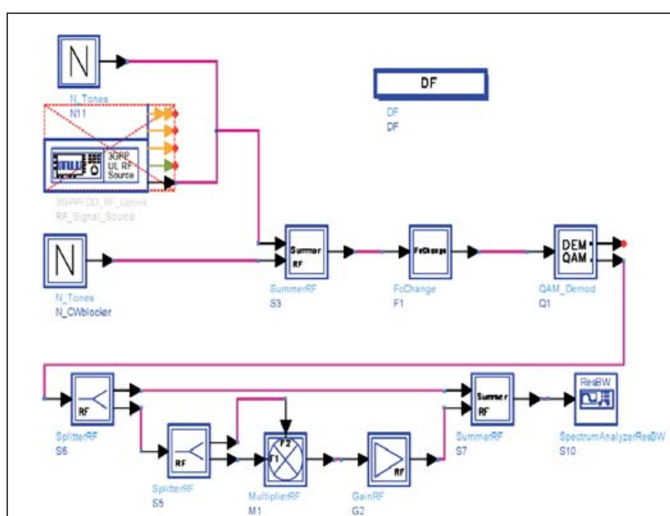

Figure 1 · Advanced Design System (ADS) bench for IP2 simulation.

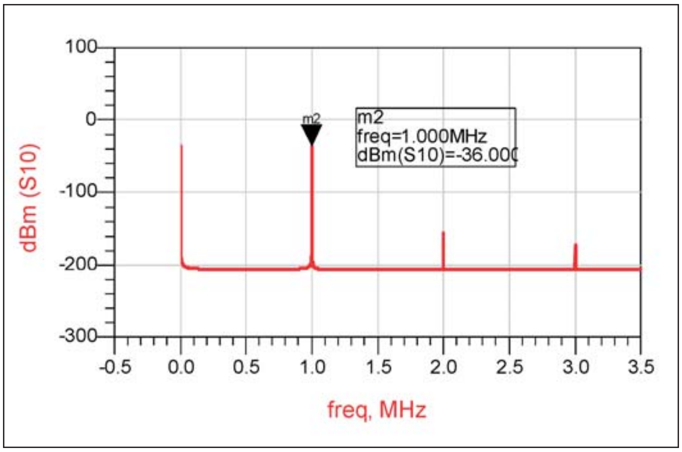

Figure 2 · Low frequency output of two tone simulation for IP2.

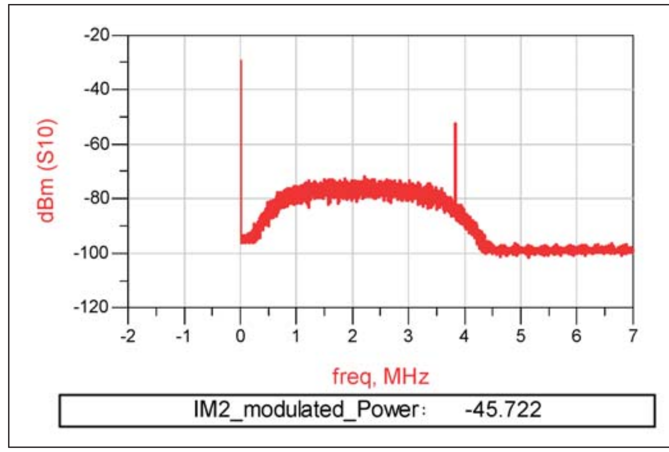

Figure 3 · Low frequency output of WCDMA uplink simulation for IP2.

$$P_{\mathrm{im2\_txleakage}} = 2 \cdot P_{\mathrm{in}} - \mathrm{IIP2} - 15.722\ \mathrm{dB} \tag{2}$$

IIP2 With In-Band Blocking

One of 3GPP test cases is adjacent blocker test, in which the modulated downlink blocking signal is injected into the receiver with offset frequency at ±5 MHz. By using method similar to the one discussed above, with an adjacent blocker signal (Test Model 1), the corrected formula for IIP2 with a WCDMA downlink signal is:

$$P_{\mathrm{im2\_adj}} = 2 \cdot P_{2T} - \mathrm{IIP2} - 3.87\ \mathrm{dB} \tag{3}$$

IIP2 With Out of Band Blocker (OOB)

In OOB test cases, depending on the blocker frequency, the IM2 products of the OOB blocking signal and TX leakage signal may fall into the RX band. The power of IM2 can be calculated using the formula in Equation 4.

$$P_{\mathrm{im2\_OOB}} = P_{\mathrm{CW}} + P_{tx} - \mathrm{IIP2} \tag{4}$$

III. Third Order Nonlinearity

For third order nonlinearity, the 3GPP intermodulation test case defines the IP3 requirements of the receiver. However the blockers such as adjacent channel blockers, narrow band blockers, out of band blockers, etc either leak into the RX channel, or cross-modulate, or intermodulate with the TX leakage signal, the distortion products falling into wanted channel. The receiver needs to have good linearity performance under all blocker conditions.

IIP3 With Two Tone Test

Assuming the input signal as V_i with two tone signals,

$$V_i = A1 \cdot \cos(\omega_1 t) + A2 \cdot \cos(\omega_2 t)$$

The output signal $y(t)$ can be expressed as:

$$y(t) = a_1 \cdot V_i(t) + a_2 \cdot V_i^2(t) + a_3 \cdot V_i^3(t)$$

The third order intermodulation products are:

$$y(t)_{3\mathrm{rdorder}} = 3a_3A_1^2A_2\cos(\omega_1)^2 t\cos(\omega_2)t +$$
$$3a_3A_1A_2^2\cos(\omega_1)t\cos(\omega_2)^2 t$$
$$= \frac{3a_3A_1^2A_2}{4}\cos(2\omega_1 - \omega_2)t + \frac{3a_3A_2^2A_1}{4}\cos(2\omega_2 - \omega_1)t + \ldots$$

The third order IM3 products are created at: $2f1 + f2$, $2f1 - f2$, $2f2 - f1$ and $2f2 + f1$. In a down conversion receiver, only the low frequency products are interesting. The power of IM3 at low frequency is expressed in following classical formula:

1. IM3 at $2f2 - f1$:

$$P_{\mathrm{IIM3}}(\mathrm{dBm}) = 2 \cdot P_2 + P_1 - 2 * \mathrm{IIP3} \tag{5}$$

2. IM3 at $2f1 - f2$:

$$P_{\mathrm{IIM3}}(\mathrm{dBm}) = 2 \cdot P_1 + P_2 - 2 * \mathrm{IIP3} \tag{6}$$

If two tones are equal, the formula becomes:

$$P_{\mathrm{IIM3}}(\mathrm{dBm}) = 3 \cdot P_{\mathrm{in}} - 2 * \mathrm{IIP3} \tag{7}$$

IIP3 With Cross Modulation

The TX signal leakage signal can be cross modulated by a strong blocker such as a narrow band blocker, adjacent channel blocker, with the cross modulated signal falling into the receive channel.

An example is illustrated in Figures 4 and 5. Figure 4 shows the output of a nonlinear circuit with two tone signal. Replacing one tone by a TX uplink leakage signal in

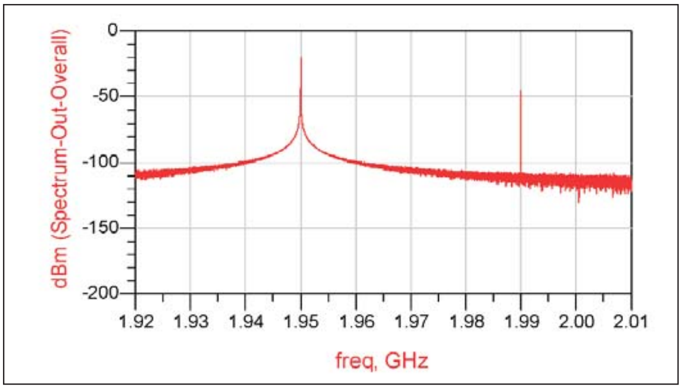

Figure 4 · Two tone simulation.

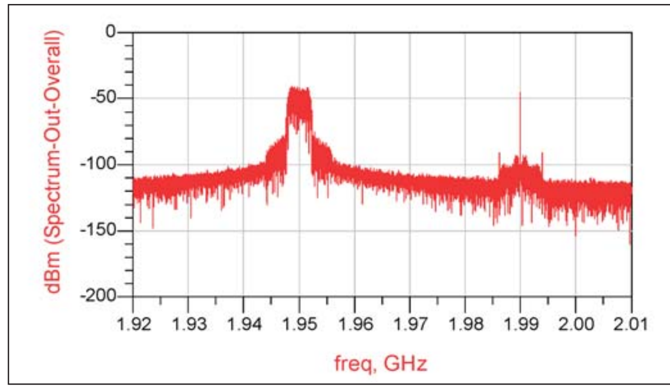

Figure 5 · One tone is replaced by TX leakage signal.

Figure 5, cross modulation products are observed around the single tone at 1.99 GHz.

For mathematical analysis, we could assume that the signal at the LNA input comprises of two interferers besides the wanted signal, namely a CW blocker signal and the TX leakage signal. Note that the TX leakage signal is amplitude modulated, while the blocker is not modulated in this analysis. Assuming the input signal $x(t)$, with blocker and TX leakage signal then we get:

$$x(t) = A_1 \cdot \cos(\omega_1 t) + A_2 \cdot [1 + m(t)] \cdot \cos(\omega_{tx} t)$$

where $m(t)$ is the amplitude modulation having a fundamental frequency at WCDMA chip rate.

When the signal inject into the non-linear RF circuit, receiver in this case, the output signal can be expressed as follows, considering only up to third order nonlinearity:

$$\begin{aligned} y(t) &= a_1 \cdot x(t) + a_2 \cdot x^2(t) + a_3 \cdot x^3(t) \\ &= a_1 A_1 \cdot \cos(\omega_1 t) + a_1 \cdot A_2 \cdot (1 + m(t)) \cos(\omega_1 t) \\ &+ a_2 \cdot A_1^2 \cdot \cos^2(\omega_1 t) + a_2 \cdot A_2^2 \cdot (1 + m(t))^2 \cos^2(\omega_{tx} t) \\ &+ 2a_2 \cdot (1 + m(t)) \cdot A_1 \cdot A_2^2 \cdot \cos(\omega_1 t) \cdot \cos(\omega_{tx} t) + a_3 A_1^3 \cos^3(\omega_1 t) \\ &+ 2a_3 \cdot A_1^2 \cdot A_2 \cdot \cos^2(\omega_1 t) \cdot \cos(\omega_{tx} t) \\ &+ a_3 \cdot A_1^2 \cdot A_2 \cdot (1 + m(t)) \cos^2(\omega_1 t) \cdot \cos(\omega_{tx} t) \\ &+ 3 \cdot a_3 \cdot A_1 \cdot A_2^2 \cdot (1 + m(t))^2 \cdot \cos(\omega_1 t) \cdot \cos^2(\omega_{tx} t) \end{aligned}$$

The term $3 \cdot a_3 \cdot A_1 \cdot A_2^2 \cdot (1 + m(t))^2 \cdot \cos(\omega_1 t) \cdot \cos^2(\omega_{tx} t)$ can be further expanded to:

$$\begin{aligned} &3/2 \cdot a_3 \cdot A_1 \cdot A_2^2 \cdot (1 + m(t))^2 \cdot \cos(\omega_1 t) \\ &+ 3/2 \cdot a_3 \cdot A_1 \cdot A_2^2 \cdot (1 + m(t))^2 \cdot \cos(\omega_1 t) \cos(2\omega_{tx} t) \end{aligned}$$

The term $3/2 \cdot a_3 \cdot A_1 \cdot A_2^2 \cdot (1 + m(t))^2 \cdot \cos(\omega_1 t)$ shows the blocker signal being modulated by the square of the amplitude of the TX leakage.

Cross Modulation With Adjacent Channel Blocker

The straight forward thinking is to use the IIP3 formula with little modification. The quantity of the cross-modulation product can be evaluated with the approximated formula:

$$P_{crossmod} = C_{factor} + 2P_{TX} + P_{adj} - 2\mathrm{IIP3} \quad (8)$$

where C_{factor} is the correction factor that takes into account the difference between using two CW tone measurement and modulated signal measurement. The correction factor has been determined by comparing the difference between two tone case and the TX leakage case in simulation. It is found that C_{factor} is around 7.4 dB. The corrected formula for the adjacent channel blocker test case is:

$$P_{crossmod\_adj} = 2P_{TX} + P_{adj} - 2\mathrm{IIP3} - 7.4 \quad (9)$$

Cross Modulation With Narrow Band Blocker

The narrow band blocker appears at ±2.7 (2.8) MHz, the offsets are much closer to the carrier compared the adjacent channel blocker case. The approximate formula is found to be:

$$P_{crossmod\_nb} = 2P_{TX} + P_{nb} - 2\mathrm{IIP3} - 2.4 \quad (10)$$

Adjacent Channel Leakage—ACLR

The receiver front end nonlinearity can create spectrum re-growth for an adjacent channel blocker, partially falling into the wanted band. The approximate formula to calculate the leakage power is reported in [3] as:

$$P_{aclr} = -20.75 + 2 \cdot (P_{in} - \mathrm{IIP3}) + 1.6 \cdot \mathrm{PAR} \quad (11)$$

where PAR is the Peak to Average ratio of the downlink signal.

IV. System Specification

In order to determine the IP2 and IP3 requirements, it is essential to consider all impairments including IM2, IM3 products, which must be low enough for the receiver sensitivity degradation to be acceptable. In the following sections, IP2 and IP3 are derived. The following analysis are applied to Band-I typical case.

IP2 at TX Frequency Requirement at Sensitivity Level

Maximum Allowed Noise

First of all, let's look at the maximum allowed noise floor without any interferers. From the 3GPP specification, the specified sensitivity is:

- Total input power: I_{or} = –106.7 dBm/3.84 MHz,
- Dedicated physical channel power: DPCH_EC = P_{sens} = –117 dBm/3.84 MHz.
- The processing gain of 12.2 kps reference channel: G_p = 25 dB

It has been reported that the required E_b/N_t, the ratio of signal energy bit to noise spectra density, shall be better than 5.2 dB to guarantee the Base Band (BB) modem can demodulate the WCDMA signal properly. In this paper, we assume E_b/N_t as 5.2 dB and add 2.2 dB implementation margin. Therefore, the allowed total noise power at antenna input is:

$$P_n = P_{sens} + G_p - \frac{E_b}{N_t} = -117 + 25 + 7.4 = -99.42 \text{ dBm}$$

Typically some margin is needed to cover process variation for production yield. By applying 1 dB margin, the total noise allowed at antenna connector is:

$$N_{\text{tmax\_ant}} = P_n - 1 = -100.4 \text{ dBm/3.84 MHz}$$

At RFIC input, taking front end loss into account, the maximum allowed interfere level is:

$$N_{\text{tmax}} = N_{\text{tmax\_ant}} - L_{\text{front}} = -104.2 \text{ dBm}$$

where L_{front} = 3.8 dB is assumed including switch loss and duplexer RX insertion loss, etc.

The distortion products due to component nonlinearity shall be kept below a certain level so that the receiver sensitivity will not be degraded too much. The interferers are not only from second and third order nonlinearity but also from other sources, which include:

- PA noise at RX band
- RFIC phase noise at RX band
- TX leakage IIP2
- TX leakage reciprocal mixing

Additionally, if blockers are present, other concerns also include:

- Blocker reciprocal mixing
- Cross modulation
- Adjacent blocker—ACLR
- Front end switch IM2/IM3 products

TX Noise at RX Band

First of all, the typical performance of the PA, RFIC and duplexer is studied in this case. Though the components from different manufacturers could perform differently, the typical performances used in this paper are summarized in Tables 1, 2 and 3, which are based on major manufacturers' datasheets.

It should be noted that for the sensitivity test case, the transmitter output power is defined as maximum output power, 24 dBm, at the antenna connector. In all other test cases, the output power at the antenna connector is defined as 20 dBm according 3GPP specifications.

The output power of the PA can be calculated by add loss between the PA output and the antenna connector.

$$PA_{\text{out}} = P_{\max} + FE_{\text{loss}} + Dup_{\text{loss}}$$

where FE_{loss} is the front end loss, including switch, PCB traces and matching loss. Dup_{loss} is the duplexer TX insertion loss.

The RFIC output power can be calculated based on PA_{out} and the PA gain. It is around 0.5 dBm, $P_{\text{rfic}} = 0.5$ dBm, in this case.

The noise present in the RX band at the LNA input due to PA phase noise is

$$P_{\text{pa\_noise}} = PA_{\text{noise}} - \text{ISO}_{\text{tx-rx}} + 10^{*}\text{Log}_{10}(\text{BW}) = -118.15\ \text{dBm}$$

where $\text{ISO}_{\text{tx-rx}}$ is duplexer isolation from TX to RX at RX band, and channel bandwidth BW = 3.84 MHz.

And the TX noise at RX band due to RFIC is:

$$P_{\text{rfic-noise}} = P_{\text{n\_rfic}} + PA_{\text{gain}} + 10^{*}\text{Log}_{10}(\text{BW}) - \text{ISO}_{\text{tx-rx}}$$

$$= -120.15\ \text{dBm}$$

Since the total noise due to TX transmitter is the summation of both, it becomes:

$$P_{\text{tx\_noise}} = 10\log 10\left(10^{\frac{PA\text{tx\_noise}}{10}} + 10^{\frac{\text{Pr fic\_noise}}{10}}\right) = -114.86\ \text{dBm}$$

TX Leakage Reciprocal Mixing

The TX leakage power at the RFIC input is:

$$P_{\text{leakage}} = PA_{\text{out}} - \text{ISO}_{\text{tx\_rx}} = -25.6\ \text{dBm}$$

Assuming the phase noise of the RX LO at the TX frequency is –157 dBc/Hz, then the reciprocal mixing at the RX band is:

	Band	Gain	Rx Noise (dBm/Hz)
PA	1	27	-140
PA	2	27	-140
PA	5	27	-140

Table 1 · PA noise and gain.

	RX-ANT(dB)	TX-ANT(dB)		RX-TX (dB)	
Band	Insertion loss	Insertion loss	RX noise attenuation	TX attenuation	RX attenuation
1	2.3	1.9	45	53	44
2	3.9	3.2	45	53	48
5	2.7	2.3	45	50	45

Table 2 · Duplexer performance.

	Band	Rx Noise (dBc/Hz)
RFIC	1	-165
RFIC	2	-165
RFIC	5	-170

Table 3 · RFIC phase noise at RX band.

$$P_{\text{tx\_reciprocal}} = -116.75\ \text{dBm}$$

TX Leakage IIP2

First, let's make an assumption that the allowed sensitivity degradation as X dB, and then we will have:

$$10^{\left(\frac{PA\text{tx\_noise}}{10} + \frac{PA\text{tx\_im2}}{10} + \frac{Nt\max}{10} + \frac{P\text{tx\_reciprocal}}{10}\right)} = 10^{\frac{Nt\max + X}{10}}$$

Arranging the equation above and expressing the above equation in dB, we get:

$$10\log 10\left(10^{\frac{PA\text{tx\_noise}}{10}} + 10^{\frac{P\text{tx\_im2}}{10}} + 10^{\frac{P\text{tx\_reciprocal}}{10}}\right)$$

$$= N_{t\max} + 10\log 10\left(10^{\frac{X}{10}} - 1\right)$$

The term

$$10\log 10\left(10^{\frac{X}{10}} - 1\right)$$

is the factor that determines how much the impairments shall be below $N_{t\max}$. For example, if $X = 0.3$ dB, meaning 0.3 dB sensitivity degradation is allowed, the total impairment power level must be 11.4 dB below the maximum allowed noise floor $N_{t\max}$. In general, it is good to keep IM2 due to TX leakage between 13 to 16 dB below $N_{t\max}$ so that the IM2 product degrades the sensitivity by only 0.1 to 0.3 dB. In this paper 16 dB is chosen for better performance. If so, P_{im2} shall be less than –120 dBm that is 16 dB below $N_{t\max}$.

From equation (2), IIP2 at the TX frequency can be derived as:

$$\text{IIP2(dBm)} = 2 \cdot P_{2T}\text{(dBm)} - P_{im2\_tx} - 15.722 \text{ dB} = 48 \text{ dBm}$$

Therefore the IIP2 requirement due to TX leakage for Band I at the RFIC input should be better than 48 dBm at the TX frequency. With IIP2 = 48 dBm, adding all other impairments, the overall sensitivity degradation is less than 0.6 dB. Further improvements in the PA phase noise and duplexer attenuation will continue to minimize the degradation.

Adjacent Channel Blocker Case 1

The adjacent blocker appears at 5 MHz offset with a power level of –52 dBm at antenna connector. Since the input wanted signal is 14 dB higher than sensitivity, it is reasonable to assume that the degradation due to the total interferer power shall be less than 14 dB. However in practice, margin is needed.

Note that with an adjacent channel blocker, the major contributor of in channel distortion is from ACLR, rather than IP2. First of all, we need to determine the power leakage into the channel due to adjacent channel blocker, P_{aclr}. To minimize the impact of ACLR, in this paper, P_{aclr} is kept 16 dB below the wanted signal.

The allowed total noise floor N_t, is 14 dB higher than $N_{t\max}$ in this test case, therefore N_t = –90.2 dBm, The allowed P_{aclr} will be $P_{aclr} = N_t - 16 = -106.2$ dBm. From the equation (10), IIP3 can be calculated,

$$\text{IIP3} = \frac{1}{2}\left(-P_{aclr} - 20.75 + 2 \cdot P_{adj} + 1.6 \cdot PAR\right) = -7.2 \text{ dBm}$$

We also need to check the IP3 requirements due to Cross modulation. Based on the cross modulation formula derived in previous section, equation (8), the cross-modulation product can be calculated:

$$P_{crossmod} = 2P_{TX} + P_{adj} - 2\text{IIP3} - 7.4 = -108.27 \text{ dBm}$$

The cross modulation product with IP3 = –7.2 dB is 18 dB below the N_t, and not a problem.

Then P_{im2_adj} due to adjacent channel blocker can be assumed 16 dB below the N_t as well. Then IIP2 at 5 MHz adjacent channel, IIP2_adj is obtained:

$$\text{IIP2(dBm)} = -P_{IIM2\_adj} + 2 \cdot P_{adj} - 3.87 \text{ dB} = -15 \text{ dBm}$$

Adjacent Channel Case 2

The blocker is present at 5 MHz offset with a power level of –25 dBm and the input wanted signal is 41 dB higher than sensitivity. In this case, N_t = –63.2 dBm, the allowed P_{aclr} is:

$$P_{aclr} = N_t - 16 = -79.2 \text{ dBm}.$$

The IIP3 and cross modulation product are calculated below.

$$\text{IIP3} = \frac{1}{2}\left(-P_{aclr} - 20.75 + 2 \cdot P_{adj} + 1.6 \cdot \text{PAR}\right) = 6.2 \text{ dBm}$$

$$P_{crossmod} = 2P_{TX} + P_{adj} - 2\text{IIP3} - 9.67 = -110.7 \text{ dBm}$$

Similar to the adjacent channel blocker case 1, IIP2 can be derived for the adjacent channel blocking case 2:

$$\text{IIP2(dBm)} = -P_{IIM2\_adj} + 2 \cdot P_{adj} - 3.87 \text{ dB} = 12 \text{ dBm}$$

In-Band Blocker 10 MHz and 15 MHz

The input wanted signal is 3 dB higher than sensitivity and the blocker at 10 MHz offset has a power level at –56 dBm, and –44 dBm at 15 MHz offset. Since the offset frequency is far from the carrier, there is no concern for ACLR in this case.

Using equation (3), keeping the IM2 product 16 dB below N_t, IIP2_adj at 10 MHz and 15 MHz are computed:

$$\text{IIP2}_{10\text{ MHz}} = -P_{IIM2\_10\text{ MHz}} + 2 \cdot P_{10\text{ MHz}} - 3.87 \text{ dB} = -12 \text{ dBm}$$

$$\text{IIP2}_{15\text{ MHz}} = -P_{IIM2\_15\text{ MHz}} + 2 \cdot P_{15\text{ MHz}} - 3.87 \text{ dB} = 12 \text{ dBm}$$

Intermodulation

For band I, the narrow band blocking test case does not apply, but for other bands. It does in the wideband intermodulation test case, one of the blocking signals is a CW tone at –46 dBm, while another blocker is a WCDMA modulated signal with power level of –46 dBm. The wanted signal is 3 dB higher than sensitivity power level.

Keeping the IM3 product 16 dB below the N_t, which is 3 dB above $N_{t\max}$, we can compute the IIP3 using equation (7).

$$\text{IIP3} = 0.5 * (3 \cdot P_{in} - P_{IIM3}) = -15 \text{ dBm}$$

Out of Band Blocker

For out of band (OOB) blocking cases, we need to consider both the second order and third order distortion from the switches that is between the duplexer and the antenna. The switches in the front end module generate second and third order nonlinearity components falling in RX band as well. Bases on the manufacturer's datasheet, it is assumed in this paper that IM2 and IM3 of the front

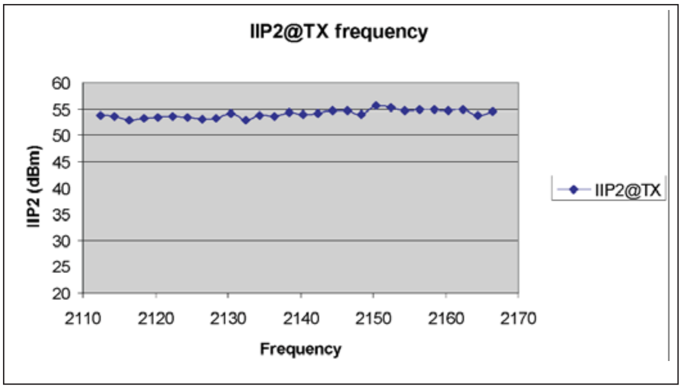

Figure 6 · Measured results of IIP2 at TX frequency.

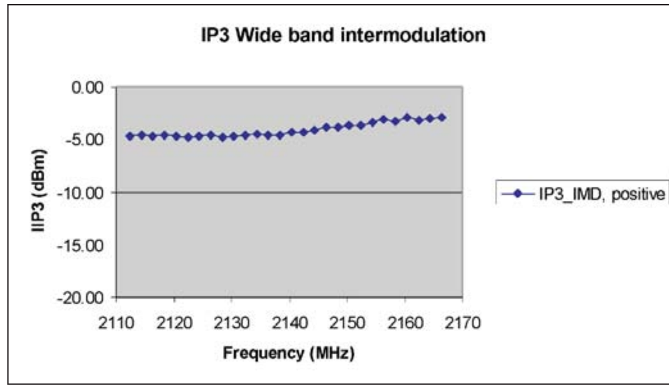

Figure 7 · Measured results of IIP3 wideband intermodulation.

end switch module are:

$P_{sw\_im2} = –110$ dBm
$P_{sw\_im3} = –110$ dBm

With the OOB test case, not all blocker frequencies produce distortion that falls into wanted band. For the second order nonlinearity, we will consider:

- $F_{cw} = F_{rx} – F_{tx}$
- $F_{cw} = F_{rx} + F_{tx}$

For the third order nonlinearity, we will consider:

- $F_{cw} = F_{rx} – \Delta F/2$
- $F_{cw} = F_{rx} – 2^*\Delta F$
- $F_{cw} = 2^*\ F_{tx} + Frx$
- $F_{cw} = \Delta F/2$

where F_{cw} is the blocker frequency, F_{rx} is the RX channel frequency, F_{tx} is the TX frequency related to F_{rx}, and ΔF is the frequency separation between F_{rx} and F_{tx}.

F_{rx} – F_{tx}

The CW blocking signal is at very low frequency, which is the separation between RX and TX frequencies. The second order IM2 results in mixing of F_{tx}+ $(F_{rx} – F_{tx}) = F_{rx}$.

As the CW blocker level is –15 dBm, and the duplexer attenuation at the CW blocker frequency is around 30 dB for Band I, we get $P_{cw} = –45$ dBm at the input of the RFIC and the TX leakage power level $P_{tx} = –29.6$ dBm.

Using equation (4), with P_{im2} 16 dB below N_t, which is 3 dB above noise floor of –104.2 dBm, IIP2 is calculated as:

$$IIP2 = P_{CW} + P_{TX} – P_{IM2} = 38 \text{ dBm}$$

F_{tx} + F_{rx}

In this case, the OOB signal is at high frequency. The mixing of $(F_{tx} + F_{rx}) – F_{tx} = F_{rx}$, falling in the RX band. The blocker level is –15 dBm, and the duplexer attenuation at the blocker frequency is around 30 dB for Band I so that $P_{cw} = –45$ dBm, and $P_{tx} = –29.6$ dBm. Using equation (3), with P_{im2} 16 dB below N_t, we have

$$IIP2 = P_{CW} + P_{TX} – P_{IM2} = 38 \text{ dBm}$$

F_{rx} – ΔF/2

The blocker power level actually varies depending on frequency offset according the 3GPP requirements.

When F_{cw} is between 2050 MHz and 2075 MHz, the blocker level is –44 dBm. As the duplexer attenuation at blocker frequency is around 5 dB, P_{cw} is –49 dBm. IIP3 is calculated as:

$$IIP3 = 0.5(2 \cdot P_{CW} + P_{TX} - P_{im3}) = -7 \text{ dBm}$$

When F_{cw} is located between 2025 to 2050 MHz, the blocker power level is –30 dBm. Assuming duplexer attenuation at blocker frequency is 15 dB, P_{cw} is –45 dBm. IIP3 is calculated as:

$$IIP3 = 0.5(2 \cdot P_{CW} + P_{TX} - P_{im3}) = -3 \text{ dBm}$$

When F_{cw} is between 2015 and 2015 MHz, the blocker power level is –15 dBm. With the duplexer attenuation at the blocker frequency of 30 dB, P_{cw} is –45 dBm. IIP3 is calculated as:

$$IIP3 = 0.5(2 \cdot P_{CW} + P_{TX} - P_{im3}) = -3 \text{ dBm}$$

F_{rx} – 2*ΔF

The CW blocker frequency is located between 1730 and 1780 MHz. To allow P_{im3} = –117 dBm, which is 16 dB below the maximum allowed noise floor N_t, the blocker power level is –15 dBm. Assuming duplexer attenuation at the blocker frequency is 38 dB, P_{cw} is –54 dBm. IIP3 is calculated as:

$$IIP3 = 0.5(P_{CW} + 2P_{TX} - P_{im3}) = 1 \text{ dBm}$$

F_w = 2*F_{tx} + F_{rx}

When the blocker frequency is between 5950 and 6130 MHz, the CW blocker has a level of –15 dBm. As the duplexer attenuation at the blocker frequency is around 30 dB, P_{cw} is –45 dBm. With P_{im3} = –117 dBm, which is 16 dB below noise floor N_t, IIP3 is calculated as:

$$IIP3 = 0.5(P_{CW} + 2P_{TX} - P_{im3}) = 5 \text{ dBm}$$

F_w = ΔF/2

In this case the blocker is defined as –15 dBm CW signal, and the duplexer loss at this frequency is 30 dBm. Then P_{cw} is –45 dBm. Similar to the previous case, to have P_{im3} 16 dB below the maximum allowed noise floor N_t, IIP3 is calculated as:

$$IIP3 = 0.5(2 \cdot P_{CW} + P_{TX} - P_{IM3}) = 5 \text{ dBm}$$

V. Measured Results

The IIP2 and IIP3 have been fully verified on a Broadcom WCDMA transceiver IC. The results of IIP2 at TX frequency and IIP3 wideband intermodulation are shown in Figures 6 and 7, respectively. Both results show that the receiver has excellent linearity.

VI. Conclusion

The system analysis of nonlinearity of a WCDMA receiver is discussed. The formulas for IP2 and IP3 for all test cases are presented and applied. The formulas are useful for system analysis. The assumed margins can be different in each test case. Also, the formulas may be slightly different from other published results due to differences in simulation setup.

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References

1. http://www.3gpp.org
2. http://www.broadcom.com
3. Harald Pretl, Linus Maurer, Werner Schelmbauer, and Robert Weigel, "Linearity Considerations of WCDMA Front-Ends for UMTS," *Microwave Symposium Digest,* IEEE MTT-S, Volume 1, 11-16, June 2000, pp. 433-436.
4. Walid Y. Ali-Ahmad, "Effective IM2 estimation for two-tone and. WCDMA modulated blockers in zero-IF," *RF Design,* April 2004.
5. T. Ranta, J. Ella, and H. Pohjonen, "Antenna switch linearity requirements for GSM/WCDMA mobile phone front-ends," *Wireless Technology,* the European Conference , 3-4 Oct. 2005, pp. 23-26.
6. Qizheng Gu, *RF System Design of Transceivers for Wireless Communications,* Springer.

Author Information

Chris W. Liu received his M.Sc from University of Montreal, Canada and B.Sc from Tianjin University, China. Chris has more than 17 years experience in the wireless communications. He is currently with Broadcom as Principal Systems Engineer. Chris can be reached by email at: chrisl@broadcom.com.

Morten Damgaard received his M.Sc. degree from University of Aalborg, Denmark in 1989. He has 20 years of experience doing cell phone RF design, semiconductor RF system design, and radio architecture design for GSM, EDGE, WCDMA, and multi-mode transceivers and radios. Morten currently works with Broadcom Corporation in Irvine, CA as a Senior Manager of Cellular RF System Engineering. He holds 31 U.S. patents.