System-Level Time-Domain Behavioral Modeling for A Mobile WiMax Transceiver

Jie He, Jun Seok Yang, Yongsup Kim, and Austin S. Kim

HIDS Lab, Telecommunication R&D Center, Samsung Electronics
jie1.he@samung.com, jyang@samsung.com, yongsup2.kim@samsung.com, austin.s.kim@samsung.com

ABSTRACT

The time-domain baseband equivalent model for a mobile WiMax, IEEE 802.16e RF transceiver has been addressed in this paper. The baseband equivalent model is built up based on complex signals and spectrum. Nonlinearity up to the fifth order harmonics is considered, and noise characteristics are modeled in time domain for amplifiers and mixers. The time-domain phase noise is also implemented as part of PLL-based local oscillators. The non-ideal parameters, DNL and INL, are defined to describe non-linearity of ADC and DAC. With this behavioral models implemented on Matlab/Simulink environment, the performance of a complete mobile WiMax transceiver can be evaluated and predicted before an actual hardware design.

1. INTRODUCTION

Wireless communication systems are getting more complicated, especially for whole integration of a RF transceiver, ADC/DAC and digital baseband [1]. Top-down design methodology from system-level to transistor-level is required for successful design and verification in modern complex systems [2]. Full transistor level verification consumes huge amount of time and efforts, which should be avoided until the end of the tape-out. Behavioral modeling is necessary for the system level simulation and could connect seamlessly the system level to the transistor level.

In case of top-down methodology, the complete system level simulation with behavioral models must be performed to determine the parameters of RF/Analog blocks. System performance specification such as constellation error or error vector magnitude (EVM) can only be derived from the time-domain simulation, not from the frequency domain. The time-domain modeling concerning the noise and nonlinearity is addressed in this paper.

For the RF behavioral modeling, there are two kinds of modeling techniques - passband model and baseband complex equivalent model. RF passband model includes all nonlinear effects and harmonics. This technique has higher accuracy, but consumes huge simulation time when the carrier frequency goes up. On the other hand, RF baseband equivalent model [3] makes the functional simulation much faster than passband model by ignoring the carrier frequency and the even order harmonics.

This paper is organized as following. Firstly, the key factors of baseband equivalent behavioral modeling such as nonlinearity and noise are introduced. Mathematical expressions for baseband behavioral modeling are shown by complex signals and spectrum. Secondly, the baseband equivalent modeling for each building block is discussed in detail, including amplifier nonlinearity with fifth order harmonics, time-domain phase noise model and ADC and DAC nonlinearity model. The third part is the verification of the models and the simulation result for a mobile WiMax [4] system.

2. MODELING FACTORS

2.1. Passband and Baseband modeling

In a RF system, all signals are real. However, behavioral models of the RF transceivers can be represented as complex format [5]. Fig.1 shows a typical RF direct-conversion transmitter including mixers, a band pass filter (BPF) and a power amplifier (PA). Here if the linear system is assumed, the transmitter output can be presented as

$$\begin{aligned} rf(t) &= I(t)\cos(\omega_c t) + Q(t)\sin(\omega_c t) \\ &= \mathrm{Re}\{[I(t) + jQ(t)] \times [\cos(\omega_c t) - j\sin(\omega_c t)]\} \\ &= \mathrm{Re}\{[I(t) + jQ(t)]e^{-j\omega_c t}\} \\ &= \frac{1}{2}\{[I(t) + jQ(t)]e^{-j\omega_c t} + [I(t) - jQ(t)]e^{j\omega_c t}\} \end{aligned} \tag{1}$$

Where, $I(t)+jQ(t)$ is the complex baseband input, and $cos(\omega_c t)-jsin(\omega_c t)=e^{-j\omega_c t}$ is the negative carrier. In the complex spectrum as shown in Fig.1, the baseband signal is modulated to the negative complex frequency, and the real output can be represented as a sum of the modulated complex signal and its conjugated part.

Fig.2 shows the direct-conversion receiver with the complex signal spectrum. The received real signal is demodulated by positive complex frequency $e^{j\omega_c t}$ and filtered by LPF. The mathematic expression is as following

$$\begin{aligned} &\{[I(t) + jQ(t)]e^{-j\omega_c t} + [I(t) - jQ(t)]e^{j\omega_c t}\} \times e^{j\omega_c t} \\ &= [I(t) + jQ(t)] + [I(t) - jQ(t)]e^{j2\omega_c t} \end{aligned} \tag{2}$$

After low pass filtering, the complex baseband signal can be recovered.

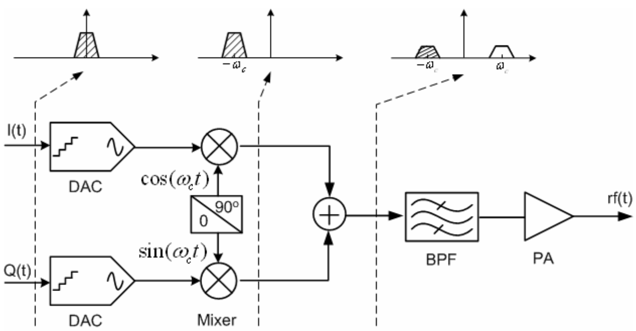

Fig. 1. Direct-conversion transmitter diagram and signal complex spectrum.

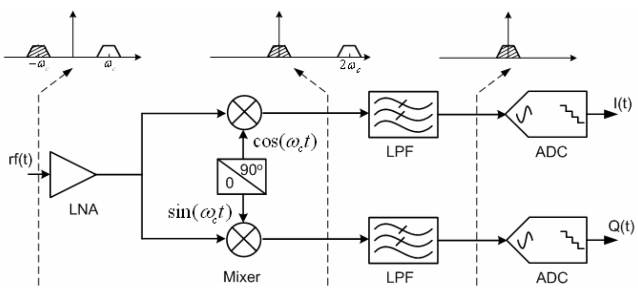

Fig. 2. Direct-conversion receiver diagram and signal complex spectrum.

The RF transceiver simulation usually consumes a lot of time and computing power. Instead of the circuit level, the behavioral level simulation is widely used to evaluate the system performance. Basically, the carrier frequency should be considered to make the simulation environment closer to the reality. The behavioral models with carrier frequency are known as passband models. Because the carrier frequency is usually high, i.e. 2.3GHz for Mobile WiMax IEEE 802.16e, and the simulation in time-domain will be quite slow due to the small time step. Observed from the complex spectrum, the desired complex signal is shifted from baseband to negative complex frequency in Tx path, and then shifted back to baseband in Rx path. By ignoring the carrier frequency, all signal processing can be performed equivalently in a baseband and complex format. The above modeling technique known as complex baseband (BB) equivalent modeling can be adapted to save the time-domain simulation time.

2.2. Nonlinearity

RF systems are nonlinear in reality. Nonlinearity should be concerned in the modeling. For the passband model, the nonlinearity can be expressed as polynomial

$$y(t) = k_1 x(t) + k_2 x^2(t) + k_3 x^3(t) + k_5 x^5(t)\,, \tag{3}$$

where x(t) is the passband signal and k_i (i=1,2,3,5) is the polynomial coefficient. Here, the 5th order harmonic is considered to describe the strong distortion.

As shown in Equation (1), passband signal can be represented as $x(t) = C(t)e^{-j\omega_c t} + C^*(t)e^{j\omega_c t}$, $C(t)$ is the complex baseband equivalent signal. The passband fundamental output can be derived from (3) as

$$y(t)|_{\omega_c} = [k_1 + \frac{3}{4}k_3 \left|C(t)\right|^2 + \frac{5}{8}k_5 \left|C(t)\right|^4]\cdot[C(t)e^{-j\omega_c t} + C^*(t)e^{j\omega_c t}] \tag{4}$$

Then, nonlinearity for baseband equivalent model can be expressed as

$$y_{BB}(t) = [k_1 + \frac{3}{4}k_3 \left|C(t)\right|^2 + \frac{5}{8}k_5 \left|C(t)\right|^4]\cdot C(t)\,, \tag{5}$$

where only odd order coefficients have the distortion effect on the fundamental. The above equation is the complex extension of [6]. Equations (3) and (5) reveal the coefficient relationship between passband and baseband models. If the passband model coefficients are given from simulation or measurement of physical blocks, the baseband model coefficient should be modified to achieve the correct characteristics.

2.3. Noise

Noise is another key factor to be concerned which will deteriorate the receiver performance. For the SPICE like simulators, noise is modeled in the frequency domain. The time-domain noise simulation is required to evaluate the communication system performance such as EVM or RMS constellation error.

The noise figure (NF) is used for RF circuits to describe the noise performance. The noise figure is defined in the frequency domain. However, the noise RMS power has to be derived from the noise figure for the time domain simulation. Fortunately, it is not difficult to translate between two domains by using below equation.

$$P_n = \sigma_n{}^2 = 4kRT \cdot f_s \cdot (10^{NF/10} + 1)\,, \tag{6}$$

where k is the Boltzmann constant, T is the noise temperature, R is the source impedance and f_s is the sampling frequency. This equation is correct for the white noise. For the color noise such as flicker noise, the filter is required to generate the correct frequency pattern. This will be discussed in the next section. To generate the white noise, a constant sampling frequency higher than the Nyquist frequency should be used to get the random signal and to keep the white noise [3]. It is also a challenge of passband model simulation. This problem can also be alleviated with the baseband equivalent model. In the RF system, noise sources can be classified as flicker noise, thermal noise, phase noise and quantization noise. LNA and PA suffer from the thermal noise. Low frequency noise at baseband amplifiers and LPFs is dominated by flicker noise. Phase noise is the key noise source in VCO and clock signal. ADC and DAC have unique source of quantization noise. Most of noise sources will be touched in this paper.

3. BUIDING BLOCK MODELING

3.1. Amplifier

Both LNA and PA are amplifiers; but the parameter sets are different due to design application. LNA usually amplifies the weak received signal with noise. As a result, both amplification and noise parameters should be concerned. On the other hand, PA amplifies large signal and the distortion is the key factor so that not only the amplitude but also the phase distortion parameters should be concerned. According to the large signal level, the noise parameter can be neglected for PA.

At first, the relationship between parameters and odd order polynomial coefficients in (3) is investigated. The fundamental coefficient is the linear gain as (7).

$$k_1 = Gain \tag{7}$$

Before deriving the high order coefficients, we should convert IP3 and P1dB from power quantity to voltage quantity, v_{IP3} *and* v_{P1dB} as following,

$$v_{IP3} = \sqrt{2 \cdot 10^{(IP3-30)/10} \cdot R} \quad , \tag{8}$$

$$v_{P1dB} = \sqrt{2 \cdot 10^{(P1dB-30)/10} \cdot R} \quad , \tag{9}$$

where R is system impedance. Because the average effective power is assumed, the factor two should be multiplied. The second and third order coefficients can be expressed as [7].

$$k_3 = -\frac{4k_1}{3v_{IP3}^2}. \tag{10}$$

According to the 1dB compression point definition [7] and the fundamental distortion from third and fifth order [6], the relationship between 1dB compression and polynomial coefficients is

$$\frac{k_1 v_{P1dB} + \frac{3}{4} k_3 v_{P1dB}^3 + \frac{5}{8} v_{P1dB}^5}{k_1 v_{P1dB}} = 0.89125, \tag{11}$$

then the fifth order coefficient k_5 can be expressed as

$$k_5 = -(0.10875 \cdot k_1 + \frac{3k_3 v_{P1dB}^2}{4}) \cdot \frac{8}{5 \cdot v_{P1dB}^4} \tag{12}$$

The polynomial model is good at low and medium input power, and has some limitation at high power. To get it over, a hard saturation is applied at high power. The reasonable estimation of saturation level is 2/3 of v_{IP3} if not specified in specification or measurement. The complex baseband equivalent model can be modified from the polynomial passband model according to equation (5) with the same k_i. Note that the second order coefficient k_2 is ignored in the baseband equivalent model.

In case of noise, the passband model has to use sampling frequency not less than $2\omega_c+BW$ to generate the white noise, and use a passband filter to guarantee that noise is band-limited around the carrier frequency. This will significantly decrease the simulation speed. For the baseband equivalent model, the required sampling frequency is not simply less than bandwidth BW, which results in dramatic enhancement on simulation speed. In the baseband equivalent model, noise power should be decomposed into complex in-phase I_n and quadrature Q_n with equal power $0.5P_n$. Because the noise parameter is based on the linear model, the noise is added at the output to avoid the nonlinear effect. The output noise should be magnified by the linear gain factor, k_1. The LNA baseband equivalent model is shown in Fig.3. PA amplifies the large signal and exhibits more nonlinearity than LNA. Strong nonlinearity can generate the phase shift proportional to the signal amplitude, which is modeled as amplitude modulation (AM) to phase modulation (PM) conversion gain. Due to the large signal, the noise model is unnecessary for the PA behavioral model. Fig.4 depicts the PA baseband equivalent behavioral model.

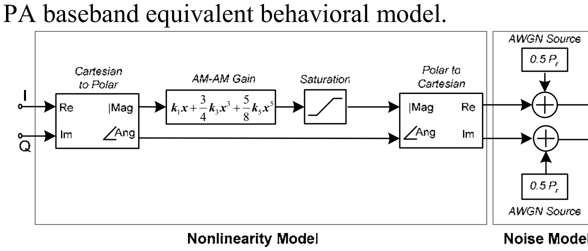

Fig.3. LNA complex baseband equivalent model

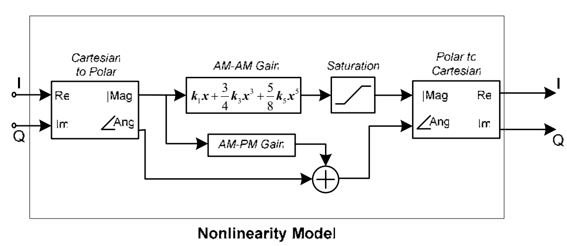

Fig.4. PA complex baseband equivalent model

3.2. Mixer

Unlike the amplifier models, a mixer has two input ports - RF and LO. The nonlinearity is usually equivalent to the RF input port and the noise is equivalent to the output. The baseband model is shown in Fig.5, where the nonlinearity model and the noise model are the same as those in LNA.

The only difference is the addition of an ideal multiplier after the nonlinearity model.

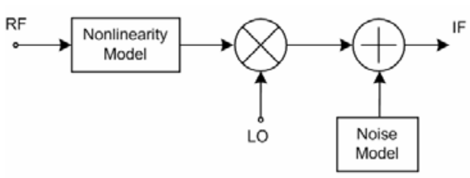

Fig.5. Mixer baseband equivalent model

3.3. Local Oscillator

Local oscillator (LO) can be represented in the complex passband format $e^{j(\omega_c t+\varphi_n)}$, where ω_c is the carrier frequency and φ_n is the phase noise. By ignoring the carrier frequency, the complex expression becomes to be baseband equivalent format $e^{j\varphi_n}$, where the phase noise is the key factor for the baseband equivalent modeling. The phase noise can be decomposed into three parts – $1/f^3$, $1/f^2$, and noise floor N_f as shown in Fig.6.

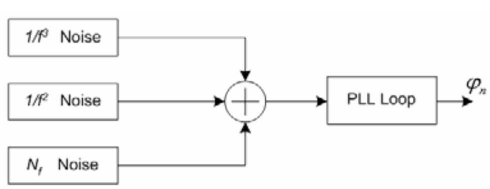

Fig.6. Phase noise model

The noise floor N_f is white noise generated from average white Gaussian noise (AWGN) noise source. The $1/f^2$ noise can be generated when the white noise goes through an integrator. Its discrete-time frequency response is

$$H(z)_{1/f^2} = \frac{1}{1-z^{-1}} . \tag{13}$$

$1/f^3$ requires the frequency response

$$H(z)_{1/f^3} = \frac{1}{(1-z^{-1})^{3/2}} , \tag{14}$$

which can be approximately implemented by FIR or IIR filter[8]. Ref. [8] also demonstrated that IIR approximation is more efficient than the FIR version in case of memory. The IIR approximation can be expressed as

$$H(z)_{1/f^3} = \frac{1}{a_0 + a_1 z^{-1} + a_2 z^{-2} + ...} \tag{15}$$

$$a_0 = 1$$

$$a_k = (k - \frac{5}{2})\frac{a_{k-1}}{k} \quad . \tag{16}$$

The accuracy of approximation depends on the length of the series a_k.

Oscillator is usually PLL based and the PLL effect should also be concerned. PLL behaves as high pass filter to the VCO phase noise and the second order HPF is modeled as PLL loop effect.

3.4. Filter

In case of the baseband equivalent model, filter effect can also be equivalent to baseband – low pass filter (LPF). In the previous discussion, a sampling frequency is utilized to implement the time-domain noise. When the sampling frequency is the channel bandwidth, the filter is assumed to be ideal and able to omit. When the sampling frequency is higher than the channel bandwidth, i.e., in case of 4 times channel bandwidth, the filter can be modeled as ideal one, or modeled as LPF.

3.5. ADC & DAC

ADC and DAC are the interface blocks between the digital baseband and analog RF. Quantization noise occurs between analog and digital conversion. Digital baseband signal with high resolution is converted into low resolution DAC, and the analog signal is quantized into the digital signal. Both cases will deteriorate the signal-to-noise ratio (SNR). Nonlinearity is another factor deteriorating the SNR. It comes from the non-uniform quantization both in ADC and DAC, which is usually defined as Integral Non-Linearity (INL) and Differential Non-Linearity (DNL). INL and DNL can be modeled into the non-uniform quantization level $v_{q,k}$ (k=1,2,…2^n, n is the number of bits). Firstly, generate the kth DNL - $v_{DNL,k}$ by Gaussian random number with zero mean and $DNL_{max}/3$ standard deviation, and then calculate the kth INL – $v_{INL,k}$ by integrating all previous DNL,

$$v_{INL,k} = \sum_{i=1}^{k} v_{DNL,i} \tag{17}$$

if $v_{INL,k} > INL_{max}$, then re-generating $v_{DNL,k}$ until $v_{INL,k} <= INL_{max}$. Finally, the non-uniform quantization level can be determined by

$$v_{q,k} = v_{INL,k} + \frac{k}{2^n} v_{ref} + v_{offset} \tag{18}$$

where v_{ref} is the analog reference voltage or full-scale voltage, and v_{offset} is the analog offset voltage, i.e. to minimize the quantization noise in typical ADC, v_{offset} is set to $v_{ref}/2^{n+1}$ which is equal to 0.5 LSB.

4. SIMMULATION RESULTS

4.1. Nonlinearity

Firstly, nonlinearity models are applied to the LNA and PA. Fig. 7 shows the input and output waveform of LNA. The parameters for LNA are: Gain=10dB, IP3=0dBm, P1dB= -10dBm and the system impedance R=50 Ω. The input signal is real with 0.15V amplitude. The output shows the nonlinear distortion. Fig.6 depicts the input and output waveform of PA. The parameters for PA are: Gain=10dB, IP3= 20dBm, P1dB= 10dBm, R=50 Ω and AM-to-PM-Gain = 1deg./dB. The input amplitude is 1V and the output shows more distortion than LNA output. The strong distortion in PA is resulted from the large signal AM-to-PM gain, in spite of the 20 dB higher IP3 and P1dB than LNA.

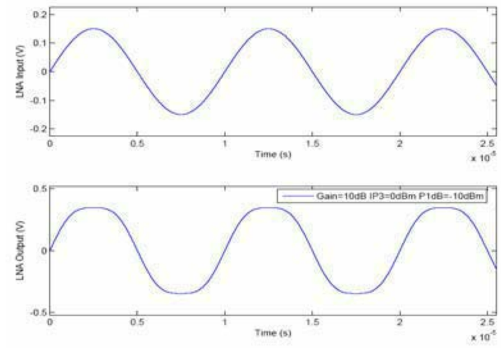

Fig.7. LNA nonlinearity simulation

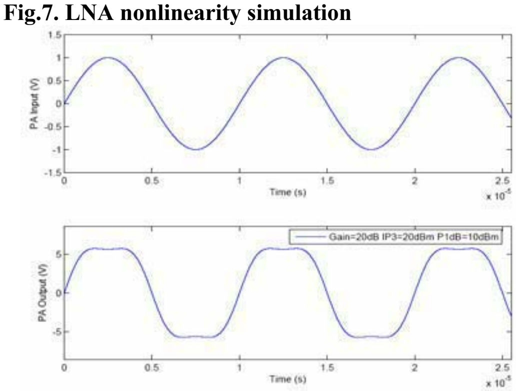

Fig.8. PA nonlinearity simulation

4.2. Phase noise

The phase noise for a local oscillator is simulated to verify the accuracy of time-domain phase noise model. The phase noise parameters for oscillator are specified as -50dBc/Hz @10kHz, -80dBc/Hz @100kHz and -120 dBc/Hz @Noise Floor. As shown in Fig.9, the simulated power spectral density of phase noise is quite close to above specification, which means the implemented model is accurate for simulation. Fig.10 shows the simulated phase noise with PLL function enabled, and the loop bandwidth is 100 kHz. The power spectral density (PSD) is analyzed by averaged FFT with the Hann window.

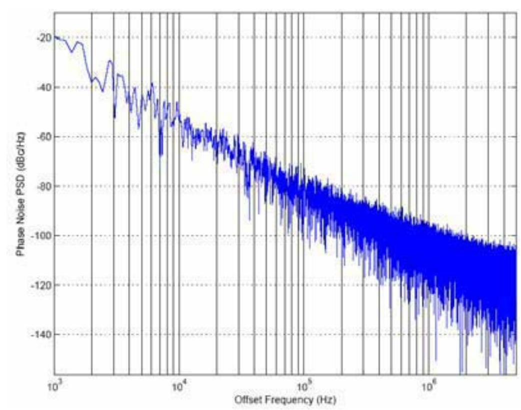

Fig.9. Simulated local oscillator phase noise without PLL

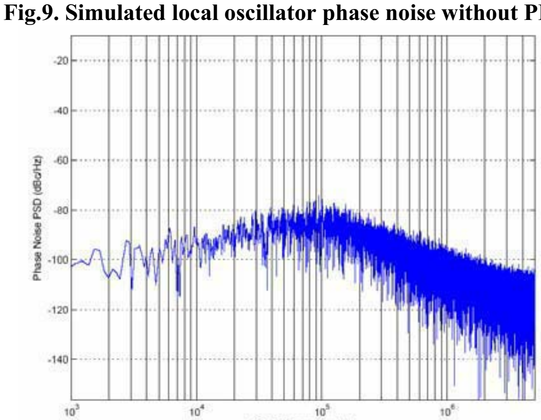

Fig.10. Simulated local oscillator phase noise with PLL

4.3. INL & DNL

Nonlinearity in ADC and DAC is verified by adapting the proposed method in section 3.5. The characteristic of DNL and INL for 10-bit converter is shown in Fig.11, from which we can identify that DNL generation is a high frequency random process and INL generation is a low frequency random process.

4.4. System Performance

The mobile WiMax system behavioral model is implemented on Simulink behavioral modeling as shown in Fig.12. The whole system level simulation consists of the OFDM baseband modulation and demodulation, a RF behavioral transmitter and a receiver, and a channel model. The RF baseband equivalent behavioral models include ADC, DAC and Tx/Rx RF conversion blocks as shown in Fig.12. The signal bandwidth is 8.75MHz and channel bandwidth is 10MHz, which determines that the minimum sampling rate is 10MHz for system simulation. Also in baseband equivalent model, the noise bandwidth is equal to the channel bandwidth 10MHz. To simulate the adjacent channel characteristics, the sampling rate is set up to 4 fold channel bandwidth. Fig.11 shows the 16-QAM constellation and spectrum of idea signal source (Point A in Fig.12), transmitted signal (Point B in Fig.12) and received signal (Point C in Fig.12). From Fig.13 (b), the adjacent channel protection ratio (ACPR) is - 35dB@±5MHz offset. According to Fig.13(c), the received signal noise ratio (SNR)

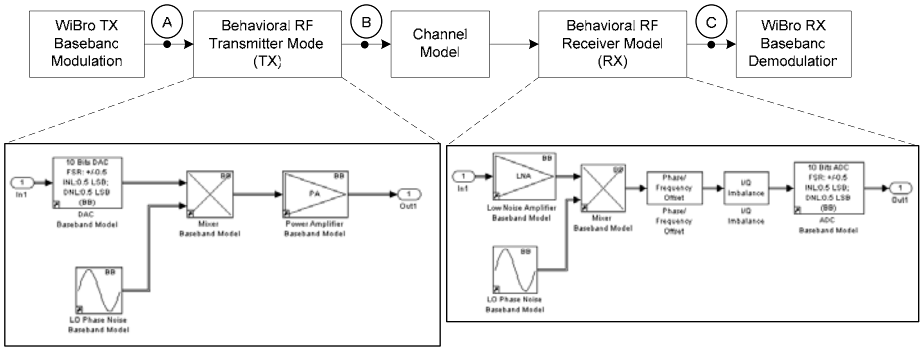

Fig.12. Mobile WiMax system implemented in Simulink with behavioral baseband equivalent model of RF transceiver.

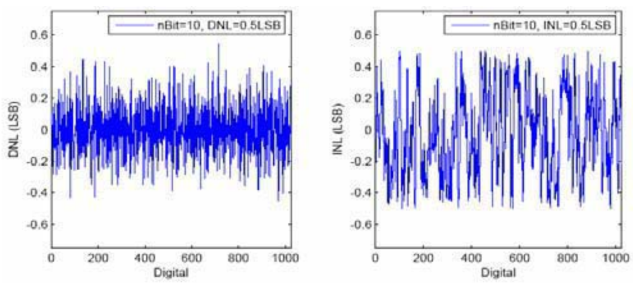

Fig.11. DNL and INL Characteristic for 10-bits converter

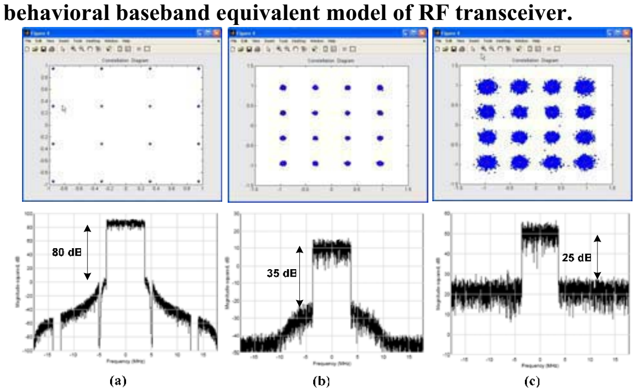

Fig.13. Wibro 16QAM Signal Constellation and Spectrum (a) Ideal signal source (b) Transmitted signal after PA (c) Received signal after ADC.

is 25dB which meets the minimum SNR requirement for $1.0e^{-6}$ BER. As one of important system performance parameters and part of standard specification, RMS constellation error (alternative of EVM) can be simulated from the constellation analysis and various simulation cases are listed in Table.1. The whole simulation for one frame OFDM signals takes about 4 minutes on a Pentium 4 3.0GHz PC with 1GB memory.

Table.1. Simulated system-level constellation performance

RF Behavioral Modeling Examples			RMS_Constellation_Error (dB)	Notes
Case 1	16 QAM	Ideal TX	-51.032	TX
Case 2	16 QAM	Nonideal TX LO	-45.449	TX
Case 3	16 QAM	Nonideal TX Mixer	-33.672	TX
Case 4	16 QAM	Nonideal PA	-21.311	TX
Case 5	64 QAM	Ideal TX	-50.056	TX
Case 6	64 QAM	Ideal TX and RX	-22.179	RX
Case 7	64 QAM	Nonideal RX LO	-20.766	RX
Case 8	64 QAM	Nonideal RX	-19.722	RX

5. CONCLUSION

The baseband equivalent behavioral models for the transceiver are discussed in this paper. Those models could be implemented in Matlab/Simulink basis to analyze overall mobile WiMax system performance. The impact of nonlinearity and noise from RF/Mixed-signal building blocks is addressed by the time-domain system simulation. Besides the mobile WiMax application, the proposed baseband equivalent model technique is able to be applied by wide range of the system-level simulation covering RF nonlinearity and noise effects.

REFERENCES

[1] M. Zargari, "A Single-Chip Dual-Band Tri-Mode CMOS Transceiver for IEEE 802.11a/b/g WLAN", International Solid State Circuits Conference 2004, Feb. 2004.

[2] Kenneth S. Kundert & Olaf Zinke. *The Designer's Guide to Verilog-AMS*. 2004.

[3] Jess. Chen, "Modeling RF Systems", http://www.designers-guide.org, 2005.

[4] IEEE STD 802.16e-2005, IEEE, 2005.

[5] Jan Crols, Michael Steyaert. *CMOS Wireless Transceiver Design*, Kluwer Academic, 1997.

[6] Paul W. Tuinenga, "Models Rush In Where Simulators Fear To Tread: Extending the Baseband-Equivalent Method", 2002 International Behavioral Modeling and Simulation Conference, Santa Rosa, 2002.

[7] John Roger, Calvin Plett. *Integrated Circuit Design*, Artech House, p.31, 2003.

[8] N. Jeremy Kasdin, "Discrete Simulation of Colored Noise and Stochastic Processes and $1/f^{\alpha}$ Power Law Noise Generation", Proceeding of the IEEE, Vol.83, No.5, May, 1995.