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Mobile WiMAX Handset Front-end: Design Aspects and Challenges

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1. From the Standard and Regulations to Front-End Specifications

WiMAX standards and regulations specify the minimum requirements for RF transceivers. They are expressed by parameters, which can take values that impose more or less stringent constraints on the RF system blocks such as the Power Amplifier (PA), synthesizer, filters and Low Noise Amplifiers (LNA). Among the important parameters are:

Signal parameters:

- Peak to Average Power Ratio (PAPR) of waveforms
- Peak transmitted power or output power level
- Transmit Power Control (TPC): precision and range
- Signal bandwidth

Performance criteria for transceivers:

- Noise: Noise Factor (NF), phase noise, etc.
- Linearity: AM-AM and AM-PM characteristics of power amplifiers, compression points at 1dB, interception point, noise power ratio etc.
- Efficiency: drain efficiency, power added efficiency, global efficiency
- Signal to Noise plus Distorsion Ratio (SNDR)

System performance:

- Bit Error Rate (BER) and Error Vector Magnitude (EVM)
- Adjacent Channel Interference (ACI): Adjacent Channel Leakage Ratio (ACLR) or Adjacent Channel Power Ratio (ACPR)
- Spectrum mask

System parameters or modes:

- Frequency bands and channel frequency step
- Channel bandwidth and data rate
- Duplexing mode: in time TDD, frequency FDD or half duplex FDD HFDD
- Multi-antenna MIMO and smart antenna possibility

Peak to Average Power Ratio (PAPR) of Waveforms

OFDM modulation has many well-known interests for mobile communications: robustness to multi-path, simple equalization and good spectral efficiency. Unfortunately, OFDM waveforms are characterized by high PAPR. It is sometimes noted Peak to Mean Power Ratio (PMPR). This power ratio measures signal amplitude fluctuation before the PA. It can be defined on sampled or continuous signals and estimated on RF or baseband signals with a 3 dB difference between both values. For an OFDM signal, its value depends on the number of carriers (N), constellation size (M), the shaping filter and the oversampling ratio (L). For $L=1$ and $M=2^{2n}$, the baseband PAPR is:

$$PAPR = 10\log(N) + 10\log\left(3\frac{\sqrt{M}-1}{\sqrt{M}+1}\right). \quad (1)$$

For $N=1024$ carriers this value is greater than 30 dB. But this maximum is not used in practice for transmitter design. For a random data signal, the PAPR, calculated on each OFDM symbol, is a random value and its value depends on the statistical distribution of signal amplitudes. It can be analyzed with its Complementary Cumulative Distribution Function (CCDF) that describes the probability that the PAPR of an OFDM symbol exceeds a given threshold. The PAPR value that is used in practice is an Effective Power Ratio (EPR) associated with a given probability α (typically $\alpha=10^{-2}$ or 10^{-3}). EPR is the threshold that is exceeded with the probability α :

$$CCDF(EPR) = \Pr(PAPR \geq EPR) = \alpha. \quad (2)$$

Many expressions have been proposed to approximate the probability distribution of the PAPR of OFDM signals. For example Zou's expressions:

$$\Pr(PAPR > x) \approx \sqrt{\frac{\pi}{3}} N \sqrt{x} e^{-x}. \quad (3)$$

It can be observed in Figure 1 that the probability that the power ratio is greater than 12 dB is approximately 10^{-3} for $N=1024$ carriers. For a full OFDM WiMAX signal with a 1024 FFT and a 16-QAM mapping, the PAPR (or effective power ratio with a probability of 10^{-3}) is approximately 12 dB. For the transmitter design this value of 12 dB is used instead of the theoretical maximum of 30 dB that has a very low probability of occurrence.

Fig. 1. Distribution for $N=64$ and $N=1024$.

Peak Transmitted Power or Output Power Level and Transmit Power Control (TPC): Range and Precision

The Peak transmit power is typically 23 dBm for WiMAX subscriber stations.

The power control of the transmitted signal compensates for variations in signal strength (due to distance variation for example). For WiMAX the TPC must be monotonic and able to cover a range of at least 45 dB by steps of 1 dB with a relative accuracy of 0.5 dB.

Error Vector Magnitude (EVM)

The Error Vector Magnitude or EVM represents the average deformation of a constellation after compensation of distortions due to rotation, translation and gain. The relative constellation error and ratio of error magnitude (or power) on constellation points magnitude (or power), can be expressed in percentage or in dB.

The WiMAX 802.16e standard specifies the EVM requirements. The relative constellation error is given for the different QAM mapping and coding rate: QPSK(1/2), QPSK(3/4), 16-QAM(1/2), 16-QAM(3/4), 64-QAM(1/2), 64-QAM(2/3), 64-QAM(3/4) with respective EVM values in dB equal to -15, -18, -20.5, -24, -26, -28 and -30.

For larger constellations such as 64-QAM, the distance between the constellation points is reduced for a given average power. Therefore the constraints on EVM are more severe in order to maintain an acceptable BER.

For the transmitter, the EVM value is one of the parameters (with ACLR) that specify the linearity requirements of the PA. It measures the inband distortion generated by PA non-linearity and it is more or less difficult to fulfil depending on the signal PAPR.

For the receiver, the EVM value specification has consequences on the acceptable ratio of the noise floor to peak signal. For an EVM of -30 dB, this ratio should be greater than 42 dB if we consider a PAPR of 12dB. It is also necessary to add some margin to take blocking signals, for example, into account.

Adjacent Channel Leakage Ratio (ACLR) or Adjacent Channel Power Ratio (ACPR)

The ACLR (or equivalently ACPR) parameter is the ratio of the power in the signal channel to the power in an adjacent channel. The precise definition can vary from one standard to another. The bandwidths in which the powers are measured can be different for the signal channel and for the adjacent channel.

The ACLR impacts the allowed channel spacing and adjacent channel interference ACI performance. For WiMAX, the ACLR requirements are given by national regulatory bodies (such as FCC, ETSI, TTA) and network service providers. It depends on regions, frequency bands, and channel bandwidths.

The ACLR and EVM requirements of WiMAX combined with a PAPR value of 12 dB for the signals, are very challenging specifications for linearity and efficiency of the transmitter. To fulfil linearity requirements, it is possible to use the PA with a high level of back-off. However, high PA back-off has two drawbacks. The first is the necessary over-sizing of the PA: to transmit a WiMAX signal with a maximal power of 23 dBm using a PA with a back-off of 8 dB, the saturation power of the PA must be superior to 31 dBm. The second drawback is the resulting poor efficiency, since the linear class power amplifiers have a much better efficiency for large values of mean input power than for low values. The common efficiency obtained with a class AB PA for a WiMAX signal is smaller than 20%, which is much smaller than what could be achieved for constant envelope signals such as GSM signals.

Frequency Bands

As indicated in Table 1, there are several possible frequency bands whose bandwidths are wide: 100 MHz or 200 MHz. The number and wideness of the frequency bands have consequences on the RF filtering (spurious signals are poorly filtered) and on the necessary tuning range of the RF blocks: broadband PA, LNA and synthesizer. For mobile terminals and ETSI specifications, the level of spurious signals should be below -30 dBm for a measurement bandwidth equal to 10 kHz, 100 kHz or 1 MHz depending on the frequency range spacing from the carrier.

Channel Bandwidth and Channel Frequency Step

The mobile terminal must be able to deal with the different possible channel bandwidths (see Table 1).

For Mobile WiMAX channel profiles, frequency synthesizers must be able to generate frequencies with steps of half the frequency raster (see Chapter 3).

MIMO and smart antenna possibility

The WiMAX standard includes the possibility to use MIMO (Multiple-Input Multiple-Output) technology and specifies the support for multi-antenna transceivers. MIMO technology takes advantage of diversity and spatial multiplexing in order to improve the quality of the communication and to increase the data rate in multi-path environments. Implementing MIMO technology in a mobile WiMAX terminal is a challenge because of the size of antennas in the considered frequency bands. A WiMAX transceiver can contain several transmit and receive channels: typically 1 transmit and 2 receive channels.

1.1 Regulations Characteristics

The WiMAX standard includes two sub-standards. The first standard is 802.16-2004 or Fixed WiMAX, which is designated for fixed Line-Of-Sight (LOS) point-to-multipoint wideband communications within the radio band of 10–66 GHz. The Fixed WiMAX employs a Wireless MAN Single Carrier (SC) transmission with QPSK, 16-QAM or 64-QAM modulation schemes. Either the Time Division Duplexing (TDD) or the Frequency Division Duplexing (FDD) method can be used.

The second standard is 802.16e-2005 or Mobile WiMAX, which operates in multiple licensed bands within 2-6(11) GHz and enables full mobile broadband access in a cellular network (for fixed and mobile Non-LOS (NLOS) applications). As in the Fixed WiMAX standard, both duplexing methods (TDD and FDD) can be used. Air-interface designations of the Mobile WiMAX are following:

- Wireless MAN-SCa
- Wireless MAN-OFDM
- Wireless MAN-OFDMA

Single carrier based Wireless MAN-SCa employs Time Division Multiple Access (TDMA) using modulation techniques such as BPSK, QPSK, 16-QAM, 64-QAM and 256-QAM.

Orthogonal Frequency Division Multiplexing (OFDM) and OFDM Access (OFDMA) based WMAN use 256 OFDM or a scalable OFDM with 128, 512, 1024 or 2048 carriers modulated with BPSK, QPSK, 16-QAM or 64-QAM. The channel bandwidth is variable in all of the above mentioned modes and it can take different values between 1 MHz and 20 MHz.

Due to a very large variety of multiple variables that can be employed, the WiMAX Forum organisation has established certification profiles, in order to promote the compatibility and interoperability of wireless communication products. The basic radio channel features are summarized in Table 1 and will be considered in the RF design.

Data flow rates depend on the modulation, access mode and the channel bandwidth. For example, in WMAN-OFDM and OFDMA, it is possible to achieve a theoretical throughput of 73.19 Mbit/s for a band of 20 MHz and 256 carriers using a modulation type 64-QAM. As for power requirements, the minimum power is: – 50 dBm for WMAN-SCa and –45 dBm for WMAN-OFDM and OFDMA. Maximum power at emission must comply with the classification as follows (transmit power for QPSK), (WiMAX Forum, 2008):

- Class 1: $20 \leq P(\text{dBm}) \leq 23$
- Class 2: $23 \leq P(\text{dBm}) \leq 27$
- Class 3: $27 \leq P(\text{dBm}) \leq 30$
- Class 4: $30 \leq P(\text{dBm})$

The signal integrity can be characterized by the EVM. For example, WiMAX OFDMA requires an EVM less than 3.16% (for 64-QAM(3/4)) (IEEE, 2005).

Frequency Range (GHz)	Channel Frequency Step (kHz)	Channel Bandwidth (MHz)	FFT Size	Duplexing Mode
2.3-2.4	250	5	512	TDD
		10	1024	
		8.75	1024	
2.305-2.320 2.345-2.360	250	3.5	512	TDD
		5	512	
		10	1024	
2.496-2.69	250 (200)	5	512	TDD
		10	1024	
3.3-3.4 3.4-3.8 3.4-3.6 3.6-3.8	250	5	512	TDD
		7	1024	
		10	1024	

Table 1. Mobile WiMAX profiles defined by the WiMAX Forum (WiMAX Forum, 2008).

2. Front-End Architecture Challenges

2.1 Architectures for WiMAX

Wireless communications require highly efficient and compact transceivers, whatever the signal characteristics are. In this section we focus only on the transmitter where design challenges are more critical in terms of power, size and consumption. A WiMAX transmitter architecture must meet design constraints of: providing high efficiency and linearity for a wideband OFDM (the bandwidth can be up to 100 MHz) and high PAPR signal (or high dynamic range) in the typical range of 20 dB (29 dB theoretical maximum). The linearization of the transmitter is mandatory because power amplification of the WiMAX signal introduces Non-Linearities (NLs) in amplitude and phase, as illustrated in Figure 2.

Fig. 2. Non Linearity effects of compression and conversion in high power amplification.

Identifying a type of architecture for WiMAX requires a careful study of linearization techniques and their performance with wideband and high dynamic range signals. There are several linearization techniques depending on the PAPR of the signal, the added

complexity and the increase in size and consumption of the system that can be accepted by the designers (Villegas et al., 2007). Many criteria characterize the linearization techniques such as static/dynamic processing, adaptability, frequency (digital, baseband, IF or RF), memory effect correction, complexity, stability, resulting efficiency, size increase etc. Herein, for WiMAX application, we basically classify these techniques in three groups: (i) correction techniques, (ii) anticipation of NL and (iii) those based on a decomposition and recombination of the signal, often dedicated to wideband signals.

Examples of correction techniques are feed-back (A), feed-forward (B) and the anticipation technique principle of pre-distortion (C) (see Figure 3). Their common point is to add a modification/correction to the modulated signal (before or after) at the PA stage. The architecture considerations here do not include the modulator nor the baseband signal processing. This needs a carefully derived model of the PA non-linear effects (Volterra series, Wiener or Saleh model etc.). Adaptability to the signal amplitude can be introduced in order to compensate for the lack of accuracy in the NL effects model and memory effects of the PA (also temperature drift compensation can be considered) (Baudoin et al., 2007).

Fig. 3. Principles of feed-back (A), feed-forward (B) and pre-distortion (C).

Each structure contains a major defect. Feed-back (A) reduces the gain of the amplification and introduces a bandwidth limitation due to the transfer function of the loop (stability and dynamic response). The feed-back can be realised on the amplitude (Polar feed-back) or on I and Q quadrature components of the signal (Cartesian feed-back) and both are dedicated to narrowband signal linearization. Feed-forward (B) requires a significant increase in signal processing and RF blocks in the transmitter, with the hypothesis of a precise matching between NLs and reconstructed transfer functions. The improvement in linearity will be expensively paid for in terms of consumption and size (integration). Advantages are the stability and possibility to process wideband signals. The most interesting is pre-distortion (C) because of its flexibility: the anticipation can be done in the digital part and so provide adaptability of the technique, but this needs a feed-back loop. The digital pre-distortion

represents a non-negligible additional consumption of a Digital Signal Processor (DSP) and often requires a look up table. The signal is widened in frequency because of the expensive non-linear law of the pre-distorter (as for IP_x theory on a modulated signal spectrum), requiring baseband and RF parts to be wideband designed. Interesting improvements of pre-distortion have been made with OFDM signals in (Baudoin et al., 2007).

Others techniques presented are based on a vectorial decomposition of the signal in order to drive high efficiency switched mode RF PAs with constant envelope (constant power) signals, avoiding AM/AM and AM/PM (Raab et al., 2003); (Diet et al., 2003-2004). These techniques are dedicated to correcting strong NL effects. We consider the problem of linearization in the communication chain from the digital part to the emission. This drives for a complete modification of the architecture and its elements' specifications in baseband, RF and power RF. After amplification of constant envelope parts of the signals, the difficulty is to reintroduce the variable envelope information with lower NL than in a direct amplification case, while maintaining the efficiency of the architecture. Basic examples of these techniques are the LInearization with Non Linear Components (LINC) and the Envelope Elimination and Restoration (EER) methods (and theirs recent evolutions) (Cox, 1974); (Kahn, 1952); (Diet et al., 2004).

The LINC principle relies on a decomposition of the modulated signal into two constant envelope signals as is shown in Figure 4. The decomposition can be computed by a DSP or by combining two Voltage Controlled Oscillators (VCOs) in quadrature locked-loop configuration (CALLUM). CALLUM is an interesting architecture but presents a possibility of instability and additional costs of realisation. The amplification of these two constant envelope signals drives for the design of two identical High Power Amplifiers (HPAs) at the RF frequency, and often causes signal distortion due to imbalance mismatch. Also the HPA has to be wideband because the signal decomposition is a non-linear process, and the phase modulation ratio is increased.

Fig. 4. LINC decomposition and recombination at RF power amplification.

Whatever the decomposition technique is (LINC/CALLUM), the default is that efficiency is directly determined by the recombination sum operation. It is very difficult to avoid losses at high frequency while designing an RF power combiner.

Fig. 5. Principle of the EER technique (Kahn, 1952).

Another decomposition technique was proposed by Kahn in 1952 and is basically an amplitude and phase separation (EER). This method was first proposed for AM signals as represented in Figure 5. The advantage of EER is to drive the RF PA with a constant envelope modulated signal (carrying the phase information), enabling the use of a switched and high efficiency amplifier (Raab et al., 2003); (Sokal & Sokal, 1975); (Diet et al., 2005-2008). The difficulty is to reintroduce the amplitude information using the variation of the PA supply voltage. This implies a power amplification of the envelope signal, at the symbol rate frequency. The recombination can be done with high efficiency switched (saturated) class PA because their output voltage is linearly dependant on the voltage supply. Synchronisation between the phase and the amplitude information and linear amplification of the amplitude before the recombination are the two major difficulties of such a linearization technique as reported in (Diet, 2003-2004), where a maximum delay of 3 nanoseconds during the recombination of a 20 MHz OFDM 802.11a signal causes spectral re-growth of more than 40 dBc (standard limit) at 30 MHz from the carrier frequency (5 GHz). Recently, a lot of work has been done on the EER based architectures, often classified as polar architectures (Nielsen & Larsen, 2007); (Choi et al., 2007); (Suarez et al., 2008); (Diet et al., 2008-2009). The generation of the amplitude and phase components can be expected to be done numerically thanks to the power of DSPs, as is shown in Figure 6. As was previously discussed in (Diet et al., 2003), the bandwidths of the envelope and phase signals are widened and make it necessary to design the circuit for three to four times the symbol rate. For a WiMAX signal, such a technique requires bandwidth on the phase and amplitude baseband paths in the range of a hundred MHz (as for LINC technique and any other non-linear decomposition method). Since a clipping in frequency and on the envelope is possible, they are suited for new high data rate standards such as WiMAX, where efficiency of the emitter and linearization are mandatory. Also, the multi-standards and multi-radio concepts evolved the polar architectures in multiple ways (Diet et al., 2008). For example, the recombination on the drive signal of the PA is possible because the amplitude information modulates the phase RF signal and is restored by the band-pass shape function of the following blocks: PA + emission filter + antenna. The emitted spectrum is the criterion of quality to be considered carefully, because the Pulse-Width Modulation (PWM) or $\Sigma\Delta$ envelope coding are the source of useless and crippling spectral re-growth. The efficiency is also penalized by the power amplification of such useless components but this is balanced by the advantage of high flexibility of this architecture (Robert et al., 2009). Actual work is focused on the front end design to provide the highest efficiency possible with the PA

driven signal composed by the phase and amplitude coded information. The digitally controlled PA and the mixed-mode digital to RF converter performance are key parameters in the evolution of polar architectures (Suarez et al., 2008); (Robert et al., 2009); (Diet et al., 2008).

Fig. 6. Recent improvements of EER/polar architectures for wideband OFDM signals.

To summarize, WiMAX architecture is expected to be wideband and high efficiency due to the signal characteristics. Polar based architectures seem to be the best candidate at the moment and corresponds to the actual focus in radio-communication research topics. The high performance of digital adaptive pre-distortion techniques on OFDM signals should be noted. It is later of interest in a polar/pre-distorted mixed linearized technique.

2.2 PAPR Reduction

As mentioned above, the OFDM signals suffer from the high envelope dynamics, characterized by the PAPR. The PAPR reduction methods can be categorized into two general groups - the methods introducing signal distortion and the distortionless methods. The simplest method from the former group is the clipping. In its basic form, this method simply clips the signal magnitude to the desired maximal level. In this case the PAPR is reduced at the expense of signal distortions resulting in out of band emissions and BER degradation. In order to suppress these effects, the research was directed towards more advanced methods like repeated clipping and filtering etc.

Many techniques that do not introduce the distortions have been proposed in the past. The Selective Mapping (SLM) is based on the creation of several alternative signal representations of the original input data, bearing the same information (LIM et al., 2005). The one with the lowest PAPR is selected for transmission. In the Partial Transmit Sequences (PTS) (Müller & Huber, 1997) method, the IFFT input symbols are divided into several frequency disjoint sub-blocks. The output of each sub-block is multiplied by the complex rotation factor. These factors are optimized in order to find the variant with the lowest PAPR. Moreover, methods like the tone reservation or tone injection, belonging to the group of so-called additive methods (Tellado, 2000), can be used as well.

The potential of PAPR improvement can be illustrated on the example resulting from the use of PTS (4 sub-blocks, rotation factors being either 1 or -1) and SLM (6 alternatives) methods shown in Figure 7 for the case of FFT size equal to 256.

Fig. 7. The PAPR improvement due to use of PAPR reduction for FFT size equal to 256

2.3 Integration Technologies: System-in-Package (SiP) and System-on-Chip (SoC)

Due to the increasing complexity of integrated circuits and systems, new packaging technologies are developed. Their choices shape the design of elementary circuits.

Three constraints are essential to take into account in the design and integration of architecture transceiver:

- Improving performances
- Circuit size and costs
- Energy consumption

To tackle these challenges, the silicon integration and the packaging must be taken into account when designing systems.

SoC have different digital and analog functions live on the same chip. It is a full integration on silicon. SoC technology combines collectively processed components on the same chip, and using compatible processes. It is more complex to design but can achieve more competitive prices for major production.

The overall reliability is also improved. But today, some functions such as filtering and power amplification can not employ the SoC technology, particularly for WiMAX systems.

The system must be divided into functions. This technique requires co-design. The chips are then assembled using a specific connection technology. The design constraints are released but the unit cost is higher. The reliability is lower and the operating speed of circuits may decrease depending on the chosen connection technology.

3. Frequency Synthesis for Mobile WiMAX Radios

Unlike 2G and 3G wireless communication systems with a fixed channel bandwidth, the Mobile WiMAX standard enables a variable channel size and multiple frequency allocation, and thereby offers very flexible deployment. As a result of multi-wideband operation, a high performance RF front-end has to be employed. One of the most challenging RF blocks

to design in a radio front-end is undoubtedly the frequency synthesizer. The frequency synthesizer acts as a Local Oscillator (LO) generation unit that is used to translate baseband and RF signals by means of mixing and it frequently determines the overall performance of a radio communication system. Evaluation of the frequency synthesizer's complexity and requirements relates to the WiMAX RF architecture. WiMAX considers three RF architectures, TDD, FDD and Half FDD (HFDD). Since the TDD mode utilizes a single frequency band for the uplink and downlink direction, only one local oscillator is required. Contrary to the TDD, the FDD mode requires two separate synthesizers for RX and TX due to the full duplex nature of the architecture and hence, switching requirements can be relaxed. However, the overall complexity of the FDD front-end leads to larger RFIC due to additional circuitry, higher power consumption and most importantly, to higher costs of implementation. Therefore, the TDD architecture is more suitable for the mobile version of the WiMAX standard.

Multi-carrier OFDM scheme, which is employed in WiMAX, has become very popular in wideband communication systems as it offers very high spectral efficiency and can efficiently combat selective fading resulting from multipath propagation (Cimini, 1985). High spectral efficiency is achieved by parallel low-rate narrowband modulation of orthogonal sub-carriers equally spaced by $\Delta f = 1/T_u$, where T_u is the useful symbol length (length of the FT interval). In the ideal case, all sub-carriers are orthogonal, in other words, they don't interfere with each other. Therefore, an inter-carrier guard interval commonly used in frequency-division multiplexing is not required. However, as the sub-carriers are very close together, the OFDM signal becomes less resilient to frequency errors, namely to inaccuracies such as Doppler shift and LO phase noise (Muschallik, 1995); (Zou et al., 2007). The phase noise introduced by the frequency synthesizer can be interpreted as a parasitic phase modulation of the LO carrier that is subsequently superimposed on individual OFDM sub-carriers as a result of the frequency translation (up and down conversions). This parasitic phase modulation causes significant degradation of the OFDM signal and may lead to the loss of orthogonality as a result of Inter-Carrier Interference (ICI). The effect becomes even more harmful as the number of sub-carriers increases in a given bandwidth, because the length of the OFDM symbol T_u gets longer, making the system more sensitive to the LO phase noise. However, longer symbols increase spectral efficiency of an OFDM system, and therefore, these parameters have to be compromised.

There exist two different effects of the LO phase noise on the OFDM signal: Common Phase Error (CPE) resulting from the LO noise contribution present within the sub-channel. CPE affects all sub-carriers in the same manner and can be observed as a common rotation of all constellation points in the I/Q plane (Muschallik, 1995). Other phenomenon is ICI, which results from the adjacent LO noise contribution.

Correlation between the integrated phase noise (phase jitter) and the BER degradation has been reported in (Muschallik, 1995); (Herzel et al., 2005); (Armada, 2001). In order to eliminate this impact, phase noise of the OFDM synthesizer has to be optimized according to the maximum allowed phase noise or phase jitter (for a given BER).

3.1 Frequency Synthesizer Requirements

The frequency synthesizer has to provide the range of all necessary frequencies with proper channel spacing that corresponds to the channel raster. WiMAX channel profiles are summarized in Table 1 (WiMAX Forum, 2008). The channel frequency step given by the

WiMAX standard is 250 kHz, however, the LO has to deliver the required frequency with a resolution of 125 kHz as a result of raster overlapping for different channel bandwidths (WiMAX Forum, 2008). Another directive parameter for design is the tuning range and the frequency settling time. Frequency switching time between RX and TX in the HFDD mode has to be performed agilely, with respect to settling time requirements of the standard, which is to be less than 50 μ s. Moreover, the local frequency synthesizer has to fulfil the tightest signal purity requirements that can be expressed in terms of the integrated phase noise and the spurious output. The integrated phase noise is to be less than 1 deg rms within an integration frequency of 1/20 of the tone spacing (modulated carrier spacing) to 1/2 the channel bandwidth (Eline et al., 2004). Thus for smaller channel bandwidths the integration of the phase noise can start from as low as a few hundred Hertz.

To satisfy very high requirements imposed by the WiMAX whilst at the same time benefiting from the ease of integration and flexibility, fractional- N Phase Locked Loop (PLL) based architecture becomes the appropriate design option. Fractional- N PLL can achieve very small frequency resolution equal to the fractional portion of the reference frequency and hence improve the phase noise performance by a factor of $20\log(N)$ compared to the conventional integer- N PLL (Keliu & Sanchez-Sinencio, 2005).

3.2 High-Speed Frequency Synthesizer Design Example

Fractional- N synthesizers have become very popular and widely used in a range of RF applications because they allow the comparison Phase Frequency Detector (PFD) frequency f_{PFD} to be significantly higher than the required frequency resolution (Keliu & Sanchez-Sinencio, 2005). Higher PFD frequency automatically leads to wider loop bandwidth (as the loop bandwidth is to be $< 0.1f_{PFD}$) and therefore significantly improves settling time performance compared to the integer- N PLL. Moreover, since the Mobile WiMAX standard defines the channel raster resolution as 125 kHz, the PFD frequency of an integer- N synthesizer would have to be as low as 125 kHz and hence, the division factor N would reach up to 30,400 in order to generate the highest frequency of 3.8 GHz (Valenta et al., 2008). This would, in turn, deteriorate the in-band phase noise contribution by up to 48 dB, compared to the fractional- N synthesizer with a reference frequency of 32 MHz (where $N \approx 119$, assuming the same phase noise performance of both reference clocks).

In the next example, we consider a fractional- N Charge Pump (CP) frequency synthesizer as an application for the 3.4 – 3.8 GHz channel profile. A simplified model is depicted in Figure 8. This model includes a tri-state PFD that produces output *up* and *down* signals, proportional to the phase and frequency difference between the reference and the feedback signal. PFD employs two positive edge-triggered resettable FF (flip-flop) to detect the phase and frequency difference and one AND gate to monitor the *up* and *down* signals. The upper FF is clocked by f_{ref} , the lower by f_{div} . Signals *up* and *down* are used to switch the current sources in the CP. These CP current pulses change the voltage drop on the loop impedance and tune the VCO with tuning gain of 125 MHz/V and tuning range of 3.4-3.88 GHz. The fractional division is achieved by altering the division value between two integer values N and $N+1$, hence the average division becomes a fraction. However, the periodic switching between two division values leads to a sawtooth phase error, which can create spurious fractional tones. This problem is solved with help of $\Delta\Sigma$ modulator, which randomize the switching between N and $N+1$, but on the other hand, it introduces quantization noise into the loop.

Fig. 8. General model of the fractional- N charge pump synthesizer.

The loop filter is the key component of the PLL and it characterizes the dynamic performance of the synthesizer. Loop filter design involves choosing the proper loop filter topology, loop filter order, phase margin and loop bandwidth. Due to low phase noise requirements set by the Mobile WiMAX standard, a passive filter has been chosen. The trade-off between the minimum loop bandwidth and the settling time has to be taken into account for optimal loop filter design. A simplified trade-off presented by (Crawford, 1994) is $B_{PLL} = 4/t_{lock}$, where t_{lock} is the settling time. Applied to WiMAX, the PLL would call for at least 200 kHz loop bandwidth in order to settle within 20 μ s. However, such a wide PLL bandwidth would result in a very high in-band phase noise integration and phase jitter deterioration.

A significant settling time improvement can be achieved by means of loop filter switching (Crowley, 1979). The loop filter is switched to the fast wideband mode during the frequency transition and then, after a certain programmable period, is shifted back to the normal narrowband value. To understand the switching principle, let us have a look at the PLL control theory and the PLL linearized model. The effect of a closed feedback loop on the input reference signal φ_{in} can be described by the closed loop transfer function $T(s)$ as:

$$T(s) = \frac{\varphi_{out}(s)}{\varphi_{in}(s)} = \frac{G(s)}{1 + G(s)H} = \frac{\frac{K_d K_{vco}}{s} F(s)}{1 + \frac{K_d K_{vco}}{s} F(s) \frac{1}{N}} \quad (4)$$

where the $G(s)$ represents the open loop transfer function and H corresponds to the division factor $1/N$. K_d is the gain of the CP/PFD detector and equals to $I_{cp}/2\pi$, K_{vco} is the VCO gain in MHz/V and $F(s)$ refers to the transimpedance of the second order loop filter (see Figure 8.).

$$F(s) = \frac{1 + sC_2R_2}{s(C_1 + C_2)\left(1 + s\frac{C_1C_2R_2}{C_1 + C_2}\right)} \quad (5)$$

The angular open loop crossover frequency ω_c and the phase margin θ_c are defined at the point where the magnitude of the loop gain reaches unity. This can be expressed as $\|G(s)H\|=1$ (0 dB), where

$$G(s)H = \frac{K_dK_{vco}F(s)}{sN} = \frac{I_{cp}K_{vco}F(s)}{2\pi sN}, \quad (6)$$

$$G(s)H|_{s=j\omega_c} = -\frac{I_{cp}K_{vco}}{2\pi\omega_c^2N} \frac{1 + j\omega_cT_2}{1 + j\omega_cT_1} \frac{1}{C_1 + C_2}. \quad (7)$$

And then, the open loop phase margin θ_c at the crossover frequency ω_c reads

$$\theta_c[\text{rad}] = \pi + \arctan(\omega_cT_2) - \arctan(\omega_cT_1). \quad (8)$$

T_2 and T_1 correspond to time constants of zero and the pole in the loop filter transfer function respectively ($T_2=C_2R_2$, $T_1=C_1C_2R_2/(C_1+C_2)$).

Let us consider a situation, where the crossover frequency ω_c is increased by factor a in order to increase the loop bandwidth and thereby decrease the settling time. This adjustment is applied only during the frequency transition. To ensure the loop stability at $a \cdot \omega_c$, the phase margin θ_c defined by the equation (8) has to remain constant. This can be done by means of reducing the value of T_2 and T_1 by the factor a with help of a parallel resistor $R_s = R_2/(a-1)$, as displayed in Figure 8. Moreover, the product of all elements in (7) has to be increased by factor of a^2 as the angular frequency ω_c in (7) is in the power of two. This can be done by

Fig. 9. Phase noise performance (a) and the open loop gain with the phase (b) at 3.59 GHz. The blue line corresponds to behaviour during the frequency transition (wideband mode).

means of increasing the charge pump current I_{cp} by factor a^2 . In this example, we consider $a=4$ and therefore, to activate the fast wideband mode, the CP current is increased by a

factor of 16 ($I_{cp} \rightarrow 16 \cdot I_{cp}$), while reducing the dumping resistance by a factor of 4 (by using a parallel resistor R_s). The PLL open-loop crossover frequency, the zero and pole frequency ($1/R_2C_2$ and $1/[R_2C_1C_2/(C_1+C_2)]$) are all increased by a factor of 4 while the loop stability remains unaffected (see the constant phase margin in Figure 9 b). Figure 9 a) shows the phase noise performance and the resulting phase jitter of the dual bandwidth adjustment. Phase jitter of both loop configurations has been calculated as follows:

$$\sigma_{rms} [^\circ] = \frac{180}{\pi} \sqrt{2 \int_{f_1}^{f_2} S(f) df} , \quad (9)$$

where $S(f)$ is the total phase noise at the PLL output. f_1 and f_2 correspond to the integration borders (from 1/20 of the carrier spacing to half the bandwidth). Figure 10 presents the settling time improvement that has been achieved by means of loop filter switching. The settling time has dropped from 88 μs to 32 μs (settling time within 100 Hz).

Fig. 10. Transient responses of the fast PLL for two cases: wideband mode enabled/disabled (blue/red line respectively). Plot b) displays the absolute frequency error to 3.6 GHz.

3.3 Hybrid PLL Approach for Frequency Synthesis

The product of all elements in equation (7) can be changed not only by increasing the I_{cp} , but also by simultaneous altering of the feedback division factor N and the I_{cp} . However, by altering the division factor in the feedback path, the output frequency will shift as well. To keep the output frequency constant either during the transition, or once in the stable mode, the hybrid fractional/integer PLL approach has to be considered (Memmler et al., 2000); (Kyoungcho et al., 2008). The fractional- N wideband mode is enabled along with the loop switching during the frequency transition and then, after settling, the integer- N narrowband mode is turned on. This approach brings a new degree of flexibility and alleviates CP requirements in terms of the I_{cp} current range by reducing the division ratio N . In a very particular situation, the bandwidth can be switched only by altering the feedback division factor and the dumping resistance while keeping the I_{cp} constant. In the fast wideband fractional- N mode, the additional dividers (x -labelled blocks in Figure 11) are not employed and the reference frequency as well as the feedback signal is applied directly to the PFD. After settling, the integer- N mode is enabled by switching the loop bandwidth to the normal

narrowband value and by activating two additional dividers. The additional division ratio x is chosen such that the PFD frequency corresponds to the required frequency resolution. Applied to the previous example where a 32 MHz reference is used, the division x has to be 256, in order to achieve 125 kHz resolution. The wideband fractional mode is then enabled by switching the resistance R_s and by disabling both dividers. This adjustment results in boosting the loop bandwidth by factor 16 while keeping the I_{cp} constant. The stability is not affected since $R_s = R_2 / (a - 1)$, where $a = \sqrt{256}$ and at the same time the ratio I_{cp} / N is increased by factor a^2 ; $\{I_{cp} / (N / a^2)\}$.

The hybrid approach inherits the speed performance from the fractional- N PLL and at the same time the design simplicity of the integer- N PLL (the $\Delta\Sigma$ modulator can be replaced by an accumulator because the fractional spur reduction is not needed during the transient period). Moreover, a higher loop bandwidth enlargement is possible compared to the conventional bandwidth switching (where only I_{cp} and R are altered). On the other hand, the main drawback of the hybrid architecture is evident: the phase noise performance of such a synthesizer corresponds to the performance of an integer- N synthesizer, which has inherently worse phase noise performance compared to the fractional- N architecture.

Due to the high degree of flexibility and integrability, the hybrid PLL approach is a very promising choice for multi-standard and multi-band transceivers, where different standards impose different requirements in terms of phase noise, settling time or channel raster (Valenta et al., 2009).

Fig. 11. Functional scheme of the hybrid PLL synthesizer.

4. High Performance Power Amplifier Design for Mobile WiMAX Radios

4.1 Transistor Technology

The technologies available to realize integrated circuits in the WiMAX bands are as follows. Gallium Arsenide (GaAs), which allows high operating frequency and high output power to be achieved. This technology is generally used for power amplifiers and switches. The manufacturing cost is higher compared to silicon. Using this technology requires a SiP packaging.

Silicon technologies are numerous: silicon bipolar transistor, CMOS, BiCMOS etc. Under certain conditions, they can achieve good performances, up to tens of GHz. A final interesting technology is the SiGe that may eventually replace the GaAs for some functions.

Figures of merit are classically used to describe RF and microwave transistors characteristics. They are used by the designer to compare the performance of technologies. Two frequency characteristics are used: the cutoff frequency f_t , and the maximum oscillation frequency f_{max} . These frequencies define the upper limits of component operation. They are particularly important for low noise or power devices.

The cut-off frequency f_t is defined when the amplitude of the current gain (H_{21}) is equal to 1 (0 dB) and the output is short circuited. The maximum oscillation frequency f_{max} is defined when the unilateral power gain is equal to 1 (0 dB).

For RF or microwave integrated circuit design, f_t and f_{max} must be as high as possible. A rule of thumb is to give an operating frequency equal to 10% of the maximum f_t .

There are other electrical characteristics used depending on the application. The minimum noise factor, N_{fmin} , given in dB at the working frequency, it is important for all low-noise components. The output power P_{out} , given in dBm or W, is important for power components. For such power applications, it is also important to define the output power density, expressed in watts per millimetre gate development in the case of field-effect transistors, and emitter surface in the case of bipolar transistors.

4.2 Continuous Wave (CW) Classes

Power Amplifiers in RF applications such as WiMAX are designed to linearly amplify high dynamic signals with high efficiency, due to the consideration of the battery lifetime. They are loaded by the emission filter and/or the antenna and deliver a RF power that can reach 27 to 30 dBm. In that context, the performance evaluation criteria are the spectral re-growths as a form of measure of the linearity (linked ACPR and to the ACLR), the factor of use Fu as a technological aspect for the design, see (10), and the efficiency of operation; see (11) for drain/collector η and (12) for added efficiency η_{AE} . For simplicity, PA topologies are presented here for common emitter RF MESFETs and can be easily adapted to BJT, HBT, (LD)MOS or x-HEMT transistors.

$$Fu = \frac{V_{max} \times I_{max}}{P_{RF_out}} \quad (10)$$

$$\eta = \frac{P_{RF_out}}{P_{DC}} \quad (11)$$

$$\eta_{AE} = \frac{P_{RF_out} - P_{RF_in}}{P_{DC}} \quad (12)$$

PAs are grouped in several classes of operation. A class is determined by (i) the polarisation of the transistor, (ii) the wanted shape of the output signal (voltage and current) and (iii) the hypothesis on transistor saturation (current source or switch behaviour). There are two families of PA classes: the switched class (SW), discussed in the following section, and the CW or biased class. A CW class is characterized by the polarisation point of the transistor, implying the conduction angle (θ) of the PA, as illustrated in Figure 12.

Fig. 12. Angle of conduction and polarization of a CW MESFET PA.

The PA classes with $\theta \geq \pi$, are class A (2π), AB ($2\pi > \theta > \pi$) and B (π). Their main characteristics are that the PA dissipates power whatever the amplitude of the signal, and therefore the maximum efficiency is obtained at maximum linear amplitude of the output voltage. For class C ($< \pi$) the dissipated power is non-linearly proportional to the input signal amplitude and increases with the reduction of θ . Equations of the maximum efficiency as a function of the conduction angle $\theta = \theta_c$ are based on the analysis of (Krauss et al., 1980) and are summarised by (13), (14), (15) and in Figure 12.

$$V_{RF} = \frac{V_p}{\cos\left(\frac{\theta_c}{2}\right) - 1} \quad (13)$$

$$\eta = \frac{P_{RF}}{P_{DC}} = \frac{\theta_c - \sin(\theta_c)}{4 \left(\sin\left(\frac{\theta_c}{2}\right) - \frac{\theta_c}{2} \cdot \cos\left(\frac{\theta_c}{2}\right) \right)} \quad (14)$$

$$Fu = \frac{V_{\max} \cdot i_{RF} \left(1 - \cos\left(\frac{\theta_c}{2}\right) \right)}{P_{RF}} = \frac{8\pi \left(1 - \cos\left(\frac{\theta_c}{2}\right) \right)}{\theta_c - \sin(\theta_c)} \quad (15)$$

It is important to consider that the efficiency is given as a peak value for CW classes. For an AM signal, the average efficiency will rely on the amplitude information statistic, if there is no saturation. An improvement in efficiency is gained if saturation/clipping on the peak values is introduced in order to increase the average power of the output signal for the same power dissipated by the amplifier. The amplifier gain depends on the load line and so also on θ . Gain and linearity reduce while the efficiency increases. The WiMAX signal presents such a high PAPR that the amplification by a CW class PA requires a low efficiency back off or a linearization technique to reduce the non-linear effects of compression and conversion introduced by the saturation. Techniques interesting for wideband and high PAPR signal are those using the highest efficiency PAs. A way to improve CW efficiency is to saturate

class B or class C PAs as described in (Krauss et al., 1980) as the mixed mode class C/C*. In fact, the saturation of the PA corresponds to the use of the transistor as a switch and not as a classical current source. The amplification of amplitude information by the driving input signal becomes impossible in this mode. SW class PAs are based on this principle.

Fig. 13. Efficiency and Factor of use of CW classes.

It is important to consider that the efficiency is given as a peak value for CW classes. For an AM signal, the average efficiency will rely on the amplitude information statistic, if there is no saturation. An improvement in efficiency is gained if saturation/clipping on the peak values is introduced in order to increase the average power of the output signal for the same power dissipated by the amplifier. The amplifier gain depends on the load line and so also on θ . Gain and linearity reduce while the efficiency increases. The WiMAX signal presents such a high PAPR that the amplification by a CW class PA requires a low efficiency back off or a linearization technique to reduce the non-linear effects of compression and conversion introduced by the saturation. Techniques interesting for wideband and high PAPR signal are those using the highest efficiency PAs. A way to improve CW efficiency is to saturate class B or class C PAs as described in (Krauss et al., 1980) as the mixed mode class C/C*. In fact, the saturation of the PA corresponds to the use of the transistor as a switch and not as a classical current source. The amplification of amplitude information by the driving input signal becomes impossible in this mode. SW class PAs are based on this principle.

4.3 SW Classes

SW classes are based on the hypothesis that the transistor switches perfectly. This latter is supposed to act as a current source and as a voltage source alternately with non-overlapping, which would result in dissipated power (Krauss et al., 1980); (Raab et al., 2003), see Figure 14. The filtering is mandatory in RF applications, except if the SW PA is dedicated to the amplification of a square signal, which is class D. When the square signal is low-pass filtered in order to recover a PWM or other pulse-coded information ($\Sigma\Delta$), the amplifier is often named a class S amplifier. The load line of every SW PA class tends to be that of an ideal switch, which is impossible in practice due to the real characteristics of a transistor. These imperfections are mainly represented by resistive and capacitive channel effects,

which induce current and voltage overlapping and reduce the ideal 100% efficiency. This is illustrated in Figure 14. Although the switching cannot be perfect, the SW class presents higher efficiency than the CW class at peak amplitude signals. Average efficiency is higher for AM signals if the AM information is recombined. The problem is to reintroduce the amplitude information of AM signals efficiently. One of the solutions in the case of SW PAs is that the transistor has an output voltage linearly dependant on the drain (for a MESFET) voltage due to the saturation, and this variation does not theoretically affect the high efficiency. This enables a supply modulation possibility in order to amplify non-constant modulated signals.

Fig. 14. Principle of SW classes (left) and Load line for CW and SW PA Classes (right).

Some SW classes can be considered for the emission of RF signals: class F and E. A class F PA is a saturated transistor, switched at the centre frequency, with additional resonating tanks in order to produce a sum of odd harmonics on the current and even harmonics on the voltage. This results in the generation of RF power only at the fundamental frequency, restoring the switching signal. It mathematically provides a null dissipated power (100% efficiency) but in practice the class F performance is limited by the non infinite number of resonating tanks. Another limitation of this PA class is the large number of shunt and series reactive elements necessary, all optimised for one frequency of operation. The tolerances of the inductors and capacitors limit the performance of the PA due to its increase in sensitivity to the values of the reactive elements.

Class E PA design is different from class F. The number of shunt and series reactive elements (2 to 3) of the output network is limited and the calculus of optimum values is far more complex because it results from several hypotheses on the drain voltage and drain current shapes. These are: (i) perfect switching of the transistor at the centre frequency (duty cycle 50%) implying no overlapping of the drain voltage and drain current, (ii) perfect filtering of the fundamental frequency signal delivered to the load, (iii) null drain current in the transistor off state and null voltage in the transistor on state and (iv) null value of the drain voltage derivate when the transistor is switched on. Computations of these hypotheses discussed in (Sokal & Sokal, 1975); (Krauss et al., 1980); (Raab et al., 2003); (Diet et al., 2004-2008); (Robert et al., 2009) are summarized in Figure 15, where $\theta = \omega t$.

Theoretical equations determine particular shapes of the drain voltage and drain current for the class E operation. This is the typical reference for the PA designer. The large value of the factor Fu , particular to the class E, results in the PA being large, which is costly when integrating. Despite these design difficulties, a class E PA is very attractive because it provides very high efficiency regarding the number of reactive elements necessary for the output network. Some parasitic effects due to the transistor technology such as the drain capacitance (NL) and the drain resistance can be introduced in the determination of the (inductive) optimal impedance of the output network $Z_{in}(\omega)$. Works like those of (Diet et al., 2008) show that it is possible to determine some different output networks providing the same $Z_{in}(\omega)$. This was discussed for the first parallel inductance class E topology in (Grebennikov, 2002). Two class E topologies were compared in (Diet et al., 2008) for their

Fig. 15. Class E basic design equations (Sokal & Sokal, 1975); (Kraus and al., 1980).

potential use in high efficiency RF polar architecture, showing the keen interest in class E PAs for high data rate applications such as WiMAX.

To conclude the section on PAs for RF applications, one has to consider that the resulting efficiency in a given architecture is the criterion in the choice of a PA class. In the context of WiMAX transmitters, the linearization is mandatory whichever the class. High efficiency (SW) classes are favoured due to their high efficiency but the challenge shifts to the linearity of the amplified signal, expressed by the EVM, the ACPR and whether the emitted signal spectrum respects the standard mask. Class E topology in polar architecture is currently the most popular solution.

4.4 Example of a Design for Mobile WiMAX

This sub-chapter presents and explains an example of a Mobile WiMAX transmitter. In the past years many transmitter architectures have been conceived (Masse, 2004); (Liu et al., 2005); (Pozsgay et al., 2008); (Yamazaki et al., 2008). All these transmitters are based on the direct conversion principle that offers a high degree of integration. Therefore, these architectures are preferred for Mobile WiMAX applications. Most of the time they successfully fill specifications of the standard (noise, linearity and emitted power), but have reduced performance in terms of global efficiency. As explained in (Lloyd, 2006), WiMAX PAs are mainly in CW classes. They operate with 7 or 8 dB back-off to guarantee good linearity. The efficiency of such amplifiers used in Mobile WiMAX applications is about 18%

(Eline et al., 2004). To address solutions to this issue, polar architecture associated with a class E amplifier seems to be a good trade-off between linearity and efficiency. The example presented here was extracted from (Robert et al., 2009). This transmitter is designed for WiMAX applications at 3.7 GHz. The first step is to conceive the Class E power amplifier, able to work at such a frequency. Figure 16 shows a serial inductor Class E topology (Sokal & Sokal, 1975). It is possible to calculate the optimum value of the Class E output network elements. In this case, the output of the transistor is supposed to be a short or an open circuit and the quality factor (Q) of C_0 - L_0 is set to $Q=5$ for good switching.

Fig. 16. Class E serial inductor topology and corresponding equations (Sokal & Sokal, 1975); (Robert et al., 2009).

The transistor that has been used is a GaAs E-PHEMT, Avago ATF50189. The model introduces a drain to source capacitance C_{DS} . C_1 has to be calculated considering it absorbs the value of C_{DS} . An input matching network is designed in such a way that it should not filter the wide spectrum signal provided by the architecture (see Figure 17), to preserve a constant envelope property. The amplifier offers drain efficiency (η) of 86.4% and 10.1 dB gain when the input is an 8 dBm 1-tone signal. As the amplifier will work at 3.7 GHz, a WiMAX signal is then injected at the polar architecture input (Suarez et al., 2008). Simulations parameters are fixed as: 10 MHz channel, 841 used subcarriers (1024 FFT size). The raw symbol rate is calculated using 64-QAM modulation to achieve the highest PAPR. Through the architecture, the signal provided to the amplifier is an 8 dBm constant envelope signal. First we observe the filtering behaviour of the amplifier (see Figure 17). The overall gain is 3.5 dB whereas at the section with a 20 MHz bandwidth around the carrier the gain is 8.3 dB. The signal has been amplified more at frequencies close to the carrier frequency than in the rest of the spectrum, due to the Class-E output network.

The amplifier drain efficiency obtained using a high PAPR WiMAX (10 MHz) signal through a polar $\Sigma\Delta$ architecture at 3.7 GHz is 46.2%. The overall efficiency reaches 36.8%. This example stops at the frontier between amplifier and filter. It can be observed in Figure 17 that filtering is still necessary after the amplifier. As the class E output network is dependant on the impedance presented to it (by filter and antenna), a co-design between theses three elements should be done to guaranty best performance (Diet et al., 2008).

This example shows that direct conversion architecture is not the only solution to Mobile WiMAX signal emission. A trade-off between integration, cost and performance, and thus efficiency and consumption needs to be considered.

Fig. 17. $\Sigma\Delta$ Polar architecture with class E PA output spectrum.

5. Filter Design Aspects and Solutions for Mobile WiMAX Radios

This section is dedicated to the band-pass filters required in WiMAX transmitters at the end of the transmission chain. The filter’s requirements are related to the WiMAX RF architecture as was the case with the frequency synthesizer and the power amplifier. As presented in section 2.1, architectures based on decomposition and recombination of the signal offer high efficiency and become good candidates for WiMAX transmitters. In particular, polar architectures for wideband OFDM signals require a band-pass filter after the PA to filter the quantization noise introduced by envelope coders and to guarantee that the output signal respects the spectrum power mask (see Figure 18).

5.1 WiMAX Filtering Requirements

Mobile WiMAX profiles (Table 1) are defined in terms of the frequency bands and the TDD mode which is most suitable. The RF filter in a TDD system is not required to attenuate the TX noise as severely as in FDD systems. The TDD mode prevents the TX noise from self jamming the RX since only one is *on* at any time (Eline et al., 2004). Furthermore, TDD mode implies that all the frequencies of the frequency range can be used for transmission at a given time. It is required that one RF band-pass filter be designed for the whole allocated frequency range. Within this range, any channel can be selected (3.5, 5, 7, 8.75 or 10 MHz). According to Table 1 and depending on the frequency range, the bandwidth of the RF filter can be 15 MHz, 100 MHz, 190 MHz or 200 MHz.

The ratio (f) between the filter bandwidth and its central frequency serves to compare filter requirements and is also a useful criterion to choose the filter’s technology. Table 2 summarizes the bandwidth and the fractional bandwidth (f) for each frequency range.

Freq. GHz	Range	2.3-2.4	2.305-2.32 / 2.345-2.36	2.496-2.69	3.3-3.4 / 3.4-3.8 / 3.4-3.6 / 3.6-3.8
Bandwidth-MHz	100	15 / 15	194	100 / - / 200 / 200	
<i>f</i> - %	4.26	0.64	6.96	2.99 / - / 5.71 / 5.40	

Table 2. Bandwidths and fractional bandwidths calculated from Mobile WiMAX profiles

The WiMAX standard establishes the maximum EVM accepted for WiMAX transmitters depending on the modulation and coding rate. The most stringent EVM is -30 dB (3.16%)

corresponding to a 64-QAM (3/4) modulation (IEEE, 2005). The EVM is calculated observing all the imperfections of the transmission chain blocks. Therefore, the maximum in-band ripple and group delay of the filter depend on this EVM value and on the imperfections generated by the other blocks of the architecture.

The filter's out-of-band rejection is determined from the ACLR requirements which are set by national or regional regulatory bodies. This emission mask is often designed as a transmit spectrum mask. ACLR specifications may vary by region, band and channel bandwidth. For the particular case of Europe, the mask proposed by the European Telecommunications Standards Institute (ETSI) for the last WiMAX frequency range (> 3 GHz) is used as reference (ETSI, 2003). Figure 18 presents the masks for a high complexity modulation format (such as 64-QAM) which leads to the most stringent filtering constraints. Since the filtering is carried out after the power amplification, the transmission filter must offer high power handling capability (high power dynamics and maximum power levels up to 23 dBm (Lloyd, 2006)). The transmission filter for the WiMAX transmitter should also

Fig. 18. Power Spectrum mask for a high complexity modulation format (ETSI, 2003).

present low Insertion Losses (IL) to increase the whole architecture power efficiency. Moreover, as size and cost are critical parameters for manufacturers, it is very often required to use a filtering technology that enables integration.

5.2 Filtering Technologies

The most notable RF filtering technologies include LC Filters, Ceramic Filters, Surface Acoustic Wave (SAW) Filters, Bulk Acoustic Wave (BAW) Filters and Low Temperature Co-Fired Ceramic (LTCC) Filters. LC filters can support high frequencies and can be integrated as a SoC. However, their main drawback is that they require too much area and can offer only a limited quality factor (Q). Ceramic filters offer low IL (about 1.5 - 2.5 dB), high out-of-band rejection (> 35 dB) and low cost. On the other hand the large size of ceramic filters significantly penalizes the integration.

SAW filters are smaller than LC and Ceramic filters, but have limitations in the frequency domain (up to 3 GHz). The power limitation of 1 W is acceptable for Mobile WiMAX devices. Typical IL varies between 2.5 and 3 dB and out-of-band rejection can reach up to 30 dB. The main drawback is that SAW filters cannot be IC integrated.

LTCC technology offers integration of high Q passive components along with low IL, high maximal operation frequency and acceptable out-of-band rejection. LTCC filters are smaller than LC and ceramic filters and can be integrated as SOP.

BAW filters use Film Bulk Acoustic Resonators (FBAR) that are characterized by a high quality factor Q . Moreover, they have low IL (1.5 – 3 dB), significant out-of-band rejection (≈ 40 dB) and high maximal operation frequency (up to 15 GHz). BAW filters can also deal with high output power (3 W). They are CMOS compatible and can be integrated “above IC”, which results in size reduction and package simplicity (Lakin, 2004). Besides, by employing wafer scale manufacturing using IC processing, BAW techniques offer high potential for low cost manufacturing.

5.3 Recent WiMAX Filters: BAW filters

BAW technology using Aluminium Nitride (AlN) piezoelectric material allows frequency operation up to 12 GHz. Each BAW filter is implemented with different sizes of FBARs, which are connected in Ladder or Lattice topologies (Lakin, 2004). The frequency response of this particular type of resonator depends on the thickness of a thin piezoelectric film. Ladder filters are the most straight forward to implement because resonators may be individually optimized for reactance and central frequency during the design and

Fig. 19. FBAR resonator – BVD model.

manufacturing process. Ladder filters are single ended, while lattice filters are double ended. The filter output is at the antenna input and therefore, the ladder topology is preferred to the lattice topology due to signal unbalance effects and the ease of tuning. Filters are preceded by unbalanced output of the HPA as well. The Butterworth Van Dyke (BVD) model is an electric circuit model that characterises FBAR resonators. The BVD equivalent circuit of the crystal resonator is shown in Figure 19.

The resonator is in the form of a simple capacitor, having a piezoelectric material as the dielectric layer and suitable top and bottom metal electrodes. The simplified equivalent circuit of the piezoelectric resonator has two arms. C_p is the geometric capacitance of the structure. The R_s , L_s , C_s branch of the circuit is called "motional arm," which arises from mechanical vibrations of the crystal. The series elements R_s , L_s , C_s are controlled by the acoustic properties of the device and they model the motional loss, the inertia and the elasticity respectively. These parameters can be calculated from equations presented in Figure 19. ϵ_r is the material's relative permittivity (10.59 for the AlN), k_t^2 is the electromechanical coupling constant (6% for the AlN), V_a is the acoustic material velocity (10937 for the AlN), A is the surface area of the electrodes, d is the thickness of the piezoelectric material, and Q is the quality factor. w_s and w_p correspond to 2π multiple of the resonance (f_s) and anti-resonance (f_p) frequencies of the resonator.

Thickness of series and shunt resonators can be different; d_1 and d_2 refer to the thickness of the series and the shunt resonator respectively. Only two parameters need to be optimized in order to design a band-pass filter. These parameters are the area A (expressed as $l \times l$) and

Fig. 20. $[S_{21}]$ parameter of a WiMAX RF filter using BAW technology (7th order ladder).

the resonator thickness (d_1 and d_2). The series and the shunt resonator form a stage. In order to achieve the required frequency response and out-of-band rejection, particular stages are put together to build cascades. Each additional stage (two series-shunt resonators) increases the filter order by one. Therefore, a six resonator ladder filter is a third order filter. An example of a WiMAX filter in the 3.6 – 3.8 GHz frequency band was proposed in (Suarez et al., 2008). The emission filter in this case has a bandwidth of 200 MHz. Figure 18 shows that the out-of-band rejection must be 50 dB at twice the channel bandwidth (20 MHz from the edge for a 10 MHz channel). Resonator manufacturing constraints impose the thickness to be between 1 μm and 6.5 μm and surface areas between 10 $\mu m \times 10 \mu m$ and 600 $\mu m \times 600 \mu m$. Thickness of the series and shunt resonators is 2.79 μm and 2.88 μm respectively. Figure 20 presents the frequency response of the WiMAX filter along with dimensions, IS and out-of-band rejection.

5.4 Recent WiMAX Filters: Other Examples.

Some other examples of band-pass filters for WiMAX using LTCC technologies have been published recently. For example, (Kim et al., 2008) proposes a Quad band module for Wi-Fi/WiMAX applications. This module provides insertion loss of less than 4.6 dB at 3.5 GHz and the attenuation is more than 26 dB. A LTCC filter for the same frequency band has been proposed by (Accute). It provides bandwidth of approximately 200 MHz, 2.7 dB IL, ripple of less than 1 dB and more than 30 dB of attenuation. A single band WiMAX SiP solution based on the LTCC technology has been recently proposed by (Heyen et al., 2008). It comprises the complete passive and active RF front-end plus the transceiver RFIC for up and down conversion to base band. The RF transmission filter provides harmonic and spurious rejection better than 40 dB.

6. Other Design Aspects

In previous sections the performance of the baseband and radiofrequency parts of a WiMAX transmitter were discussed. These are the parts of the transmitter between the digital

electronics and the antenna. The importance of the choice of architecture has been demonstrated, as have the impacts of key elements such as frequency synthesizers, power amplifiers and emission filters. This section points out the considerations to add for a global design of such a transceiver regarding the performance of Digital to Analogue Converters (DACs) and antennas at these frequencies.

6.1 Digital and Analogue

The DAC enables baseband signal generation after the shaping filter. It should have low distortion, sufficient bandwidth and low consumption. DACs are used in conventional architecture for I and Q paths generation and in polar architecture for phase (I and Q) and envelope paths. In polar architecture there is one DAC more and the required bandwidth is extended due to the non-linear processing when generating the “phase” and the “magnitude/envelope” of the signal. Also, the coding of the envelope is an additional restriction in terms of speed for the $\Sigma\Delta$. As the Signal to Noise Ratio (SNR) of the signal is admitted to grow with the number of bits and bandwidth, these specifications are mandatory limiting factors. Nowadays, some converters work in the range of several bits near a GHz and around 12 to 20 bits near a MHz.

Due to the conclusions of previous sections, the example presented here is the simulation of a polar architecture for an OFDM signal with 64 sub-carriers (typically IEEE802.11a). The symbol rate is 20 MHz and the carrier frequency is 5.2 GHz but can be shifted to 3.7 GHz without altering the observations because the DAC influences are introduced on the baseband processing. Figure 21 presents the emitted spectrum in an ideal polar/EER transmitter simulation with limitation of the bandwidth on the envelope and phase signals. Limits are three times the symbol rate for the envelope and seven times for the phase ones. The mask of the IEEE 802.11a standard is added on the same figure and it is noticeable that the emitted spectrum is not far from the limit. The most limiting parameters are the phase signals.

Fig. 21. Emitted spectrum of a 20 MHz OFDM Hiperlan 2 signal with band width limitations of 60 MHz (envelope signal) and 140 MHz (I and Q phase signals). The EVM rms is 0.2%.

This implies a high bandwidth for the baseband signals generation. The resolution will therefore be strongly impacted because the higher the bandwidth, the lower the resolution (without consideration of power consumption). The second step of our example is to limit the number of bits for the signal representation. Here the envelope is coded in either signed or unsigned format (depending on the specification/complexity of the hardware part of the system) and without a clipping that could have reduced the needed dynamic for the envelope but at the cost of an EVM increase. Results in the classical architecture case and polar one are presented on Figure 22.

EVM / EVM max (% rms)	7 bits	6 bits	5 bits	4 bits	6 bits (envelope unsigned)	8 bits (2's complement)
I and Q paths for OFDM sig.	0.3 / 0.5	0.6 / 0.9	1.2 / 1.8	2.8 / 5	0.9 / 1.8	0.4 / 1.1
I and Q paths OFDM phase sig.	0.1 / 0.2	0.4 / 0.7	0.7 / 1.3	1.6 / 2.5		
Envelope default is unsigned format	0.2 / 0.3	0.4 / 0.5	0.6 / 0.9	1.3 / 2.1		

Fig. 22. Results of resolution limitation for an OFDM Hiperlan 2 transmitter.

The limitation of the resolution with an acceptable EVM of 0.5% rms (without any other architecture imperfection) is at the edge of the actual DACs performance, which is 8 bits with a supposed bandwidth of tens of MHz. To realistically illustrate the influence of both parameters introduced in the simulation, we show in Figure 23 the simulation of the polar/EER architecture with DACs of 8 bits resolution and with the same bandwidth limitations as shown in Figure 21. The emitted spectrum is compared with the same mask and the constellation and EVM are presented. The results show an acceptable EVM below 0.5% rms and the spectrum is, in conclusion, the main criterion for characterizing the DAC impact in the architecture.

Fig. 23. Emitted spectrum and constellation for an OFDM Hiperlan 2 transmitter with bandwidth limitations of 60 MHz (envelope signal) and 140 MHz (I and Q phase signals), and an 8 bit DAC.

6.2 Antennas

Antennas for handsets have to be adapted to the difficult environment of indoor mobility (omni-directivity or wide radiation lobe, polarization) while maintaining a small size and cost. Solutions are, for example, helicoidal antennas, patch or planar antennas with tuned slot; often with a ground reflector in the case of mobile phone application to avoid radiations toward the user and coupling to the circuit (in this case the ground plane is a kind of “shield”). The use of antenna diversity or Multiple Input Multiple Output (MIMO) benefits the receiver and significantly increases its performance, but this is a challenge in terms of power consumption for a battery operated system (additional RF sub-systems). In the case of the integration of multiple wireless systems, it is important to focus on antenna integration and especially multi-band or wideband antennas. Whatever the standards considered, diversity of antennas and antennas for multiple standards are research topics for systems offering mobile communications and connectivity (such as WiMAX). In conclusion, integrated low cost antennas are to be investigated for this type of system with regards to the standards specifications (bandwidths, propagation environment) and with architectural considerations (size, cost, consumption in the case of MIMO).

7. Conclusion

As a result of flexible and multi-band radio operation, the Mobile WiMAX standard presents a challenge for every stage of the RF front-end. Promising techniques and mechanisms for linear and high efficient transmission have been discussed, along with their advantages and limitations. The ultimate goals are high degree of RF integration into cheap CMOS technology and high power efficiency along with linearity. At this point, the polar based architecture seems to offer high performance solutions for high PAPR wideband signals, while providing high efficiency due to switched mode amplification.

It has been shown that the RF filtering, which is required after the power amplifier presents a significant challenge for RF designers. Appropriate filtering technologies have been presented, including current examples of WiMAX filters.

Moreover, signal deterioration resulting from the frequency synthesizer's phase noise contribution has been discussed as well, along with solutions for low noise high speed synthesis.

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WIMAX New Developments

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WiMAX (Worldwide Interoperability for Microwave Access) is a wireless broadband access network named by industry group called the WiMAX forum formed in June 2001. It is Wireless MAN with IEEE 802.16 family standards. Loosely, WiMAX is a standardized wireless version of Ethernet that enables the last mile, intended primarily as an alternative to wire technologies (such as Cable Modems, DSL and T1/E1 links) to provide broadband access to customer premises. Mission of the WiMAX forum is to promote and certify compatibility and interoperability of broadband wireless products. This book touches most of the above issues in form of 22 individuals' papers containing research work in WiMAX domain in particular. WiMAX has two important standards/usage models: a fixed usage model IEEE 802.16-2004 for Fixed Wireless Broadband Access (FWBA) and a portable usage model IEEE 802.16e-2005, which is mainly concentrated on Mobile Wireless Broadband Access (MWBA). Both are released as standards and amendments are available in form of drafts. Higher data rate transmissions (@ 100 Mbps) are achieved in IEEE 802.16-2004 WiMAX through LOS communications which incorporate a stationary transmitter and receiver but IEEE 802.16e supporting NLOS communication is much complicated and little less bit rate is achieved. 2-11 GHz licensed band is the range of frequencies with TDD and FDD supports. The book will provide a wide horizon to visualize the WiMAX technology and its developments leading towards 4G systems. It will provide a good platform to the researchers with clues to the innovative ideas in WiMAX domain. I wish all the best to the authors and readers of this book in their successful research of WiMAX technology.

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