Energy reduction in wireless systems by dynamic adaptation of the fixed-point specification

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Abstract—One of the most important applications of Digital Signal Processing is wireless communication. This kind of application requires low power implementation of DSP, which generally uses fixed-point arithmetic. The fixed-point architectures should be developed to maintain the energy consumption power at a reasonable level. In this paper, an approach which adapts the fixed-point specification according to the input receiver SNR (Signal-to-Noise Ratio) is proposed. To underline our approach interest, two applications are examined, one for QPSK receiver and other for WCDMA receiver. An architecture used for this dynamic precision scaling concept is also detailed.

I. INTRODUCTION

Wireless communication is one of the most important sector for Digital Signal Processing (DSP) applications. In 2007, 74% of Digital Signal Processors sold was used for wireless applications [1]. Most of wireless terminals are nomadic and are supplied by battery. The design of low power terminals is one of the key challenges in this domain. New services are provided (image, video, Internet access) and require high data rate. Consequently, the complexity of the baseband digital part is growing. However, the energy consumption can not be increased due to the limited battery lifetime. Thus, new strategies to reduce or maintain the energy consumption power at a reasonable level must be proposed.

Efficient implementation of DSP applications in embedded systems requires the use of fixed-point arithmetic. Thus, the vast majority of embedded DSP applications are implemented in fixed-point architectures [2], [3], [4], [5]. Indeed, fixed-point architectures are cheaper and more energy efficient than floating-point architectures because in fixed-point architecture, the word-lengths of the data are lower.

The energy consumption of an application depends on the word-length of the manipulated data. The energy consumption can be reduced by decreasing the word-length of the data. Nevertheless, this also reduces the computation accuracy. The unavoidable error due to finite word-length computation increases when the data word-length is reduced. In [6], an LMS (least mean square) adaptive filter has been studied. The energy consumption is divided by a factor of two between two fixed-point specifications having a signal to Quantization Noise Ratio of 90 dB and 30 dB.

The traditional approach to design a fixed-point system is based on the worst-case principle. For a digital communication receiver, the maximal performances and the maximal input dynamic are retained and the more constraint transmission channel is considered. Nevertheless, the noise and the signal levels evolve during time. Now, the data rate depends on the service (video, image, speech) used with the terminal and the required performances (bit error rate) are linked to the service. These different elements underline that the fixed-point specification depends on external elements (noise level, input signal dynamic range, quality of service) and can be adapted during time to reduce the average power consumption.

In [7], different trade-offs between accuracy and energy are explored in the context of Software Defined Radio. New standards like the further extension of WLAN (802.11g) offer multiple configurations according to link noise robustness and data rate. Different modes (modulation scheme and coding rate) are proposed. In [7], for each mode an optimized fixed-point specification is determined and leads to a specific implementation. The selection of a fixed-point specification for each modulation scheme and coding rate makes possible the decrease of the average energy consumption by a factor three. In this approach, the adaptation of the fixed-point specification is only linked to the modulation scheme and coding rate.

In [8], word-length tunable VLSI architecture for an OFDM (Orthogonal Frequency Division Multiplexing) demodulator has been proposed. The data word-length are determined at run time according to the observed error at the system output. For this word-length determination process, word-length search symbols are inserted in the frame. This approach saves 32% and 24% of the power consumption for different transmission channels. This technique requires a specific hardware and energy are wasted for the fixed-point optimization process which is carried-out at run time. Moreover, this technique must modify the transmission packet format and can not be used in standard systems.

In this paper, an approach in which the fixed-point specification is adapted according to the input receiver...
SNR (Signal-to-Noise Ratio) for one modulation scheme and one data rate is proposed. This concept is called Dynamic Precision Scaling (DPS). Our approach interest is underlined through a WCDMA (Wide-band Code Division Multiple Access) receiver example.

The paper is organized as follows. In Section 2, the principles of Dynamic Precision Scaling is detailed and the target architectures used for implementing this concept are presented. A simple example corresponding to a QPSK (Quadrature Phase Shift Keying) receiver is analyzed in Section 3. Then, a WCDMA receiver is studied in Section 4. The path search module and the rake receiver are presented. A simple example corresponding to a QPSK target architectures used for implementing this concept are introduced. In [10], a multiplier able to perform operations on 9, 11, 14 and 16 bits is proposed. The other way is to use operator executing only one operation per cycle but able to manipulate data with a word-length. Thus the global energy consumption is reduced by diminishing the processing execution time. But, the execution time of the processing depends of the operand word-length. The aim is to reduce the energy when the word-length is lower than the supported maximal value. Two kinds of approaches are available to minimize the energy consumption through word-length flexibility. One way is to have operators supporting Sub-Word Parallelism (SWP) operations. The operator processes several operations in parallel on operands of smaller word-length. An operator (multiplier, adder, shifter) with a word-length of N is split to execute k operations in parallel on sub-words of N/k word-length as illustrated in Figure 2. This technique can accelerate the code execution time up to a factor k. Thus, the energy consumes at each cycle is constant and independent of the operand word-length. But, the execution time of the processing depends of the operand word-length. Thus the global energy consumption is reduced by diminishing the processing execution time. The other way is to use operator executing only one operation per cycle but able to manipulate data with different word-lengths. In [10], a multiplier able to perform operations on 9, 11, 14 and 16 bits is proposed.

II. DYNAMIC PRECISION SCALING (DPS)

A. Principle

In the Dynamic Precision Scaling (DPS) approach, to reduce the energy consumption, the fixed-point specification is adapted according to the external environment parameters. During time, the system switches between different fixed-point specifications when the external environment parameters are modified. In our approach, different fixed-point specifications are available. They are determined at the system design level. Let $S_{fp}$, be the set of all the fixed-point specifications which can be used.

To adapt the fixed-point specification to the external environment parameters, a metric $p$ describing the external conditions is used. This metric is determined inside the digital system from the measurement of the input signal and/or the output signal. The fixed-point specification is selected according to this metric value as illustrated in Figure 1. Let $f_{fp}$, be the function defining the fixed-point specification to used according to the $p$ metric value:

$$f_{fp} : \mathbb{R} \rightarrow S_{fp}$$
$$p \mapsto f_{fp}(p)$$

In the examples presented in this paper, the metric $p$ is the signal to noise ratio (SNR) at the input of the receiver. Different techniques can be used to estimate this SNR [9]. For the WCDMA receiver, data-aided techniques can be used. Indeed, a pilot sequence is available in the control symbol frame (DPCCH) of the WCDMA norm in the context of UMTS/3G wireless communications. Otherwise for the QPSK receiver, SNR estimators can be used with an estimate of the transmitted symbols from the receiver decisions. The selection of the estimator is a trade-off between the estimation quality and the estimator complexity. Indeed, the supplementary energy consumption due to adaptation part must be minimized to not wreck the energy gain due to fixed-point specification adaptation.

B. Architecture for DPS

To adapt the fixed-point specification during time, the architecture must be programmable or reconfigurable. For processors, a specific code (function) is associated to each fixed-point specification. The processor switches between the part of code when the metric $p$ is modified. For reconfigurable architectures, a configuration is associated to each fixed-point specification. The architecture is reconfigured when the metric $p$ is modified.

To adapt the fixed-point specification during time the processing unit must be flexible in terms of supported word-length. The aim is to reduce the energy when the word-length is lower than the supported maximal value. Two kinds of approaches are available to minimize the energy consumption through word-length flexibility. One way is to have operators supporting Sub-Word Parallelism (SWP) operations. The operator processes several operations in parallel on operands of smaller word-length. An operator (multiplier, adder, shifter) with a word-length of $N$ is split to execute $k$ operations in parallel on sub-words of $N/k$ word-length as illustrated in Figure 2. This technique can accelerate the code execution time up to a factor k. Thus, the energy consumes at each cycle is constant and independent of the operand word-length. But, the execution time of the processing depends of the operand word-length. Thus the global energy consumption is reduced by diminishing the processing execution time. The other way is to use operator executing only one operation per cycle but able to manipulate data with different word-lengths. In [10], a multiplier able to perform operations on 9, 11, 14 and 16 bits is proposed.

III. SYMBOL DETECTION

A. Introduction

In this section, to analyse deeply the quantization noise effects, a simple example is considered. It consists of a transmitter and a receiver using QPSK modulation. The transmission channel is additive white gaussian noise (AWGN) with $E_b/N_0$ varying between 0 dB and 10 dB. No channel coding is used. The input signal of the receiver and the bit error rate are observed. Throughout this paper,
$E_b/N_0$ is used in the different simulations. In general, $E_b/N_0$ is equal to SNR divided by spectral efficiency. In the case of QPSK modulation, $E_b/N_0$ is equal to $\text{SNR}/\log_2 4$.

In case of spread spectrum transmission with a spreading factor $SF$, $E_b/N_0$ is equal to $\text{SNR} \times SF$. For convenience, the term SNR is used in the text to represent $E_b/N_0$.

The aim of the following analyses is to determine the minimal number of bits required for the integer and fractional parts. The fixed-point specification must guarantee no overflow and maintains the bit error rate performances.

### B. Range estimation

The aim of this part is to determine the minimal value of the integer part word-length which guarantees no overflow. In case of an AWGN channel, the input signal $y(k)$ of the receiver is the sum of emitted symbols and white Gaussian noise:

$$ y(k) = s(k) + c(k) \tag{2} $$

where $s(k)$ is the set of modulated symbols:

$$ s(k) \in \{\pm 1 \pm i\} \tag{3} $$

Equations (2) and (3) show that the higher the noise level is, the larger the dynamic range of received signal. For example, the dynamic range at 0 dB should be as twice as the one for very low noise level.

The dynamic range obtained for different SNR levels is presented in Figure 3. The range of input signal decreases when the signal/noise ratio decreases. It decreases from 3.2 (at 0 dB) to 1.7 (at 10 dB). Between these two SNR values, one bit at the receiver input can be saved.

### C. Precision analysis

The aim of this part is to determine the minimal value of the fractional part word-length which leads to a Bit Error Rate (BER) close to the BER obtained with infinite precision (floating point simulation in our case) computation.

The channel is assumed to be a white gaussian $c(k)$ with noise variance $\sigma_c^2 = \frac{N_0}{2E_b}$. The expression of the probability density function (pdf) $f_c(x)$ of $c(k)$ is thus as follows:

$$ f_c(x) = \frac{1}{\sigma_c\sqrt{2\pi}} \exp\left(-\frac{x^2}{2\sigma_c^2}\right) \tag{4} $$

The signal $y(k)$ is then quantized with quantization step size $q = 2^{-m}$. The expression of the probability density function $f_q(x)$ of the quantization noise is then as follows:

$$ f_q(x) = \frac{1}{q} \frac{\text{Id}_{[-q/2, q/2]}}{\sqrt{2\pi}} \tag{5} $$

Thus the noise after quantization corresponding to the sum of the channel noise and the quantization noise has the following probability density function:

$$ f_n(x) = f_c * f_q(x) = \int_{-\infty}^{\infty} f_c(t)f_q(x-t)dt \tag{6} $$

$$ = \frac{1}{2q} \left(\text{erf}\left(\frac{x+\frac{q}{2}}{\sqrt{N_0}}\right) - \text{erf}\left(\frac{x-\frac{q}{2}}{\sqrt{N_0}}\right)\right) \tag{7} $$

and the following probability distribution:

$$ F_n(x) = \frac{1}{2q} \left(\text{erf}\left(\frac{x+\frac{q}{2}}{\sqrt{N_0}}\right) - \text{erf}\left(\frac{x-\frac{q}{2}}{\sqrt{N_0}}\right)\right) \tag{8} $$

In case of BPSK and QPSK, the probability of bit error for single quantization is:

$$ \text{BER}(m, \text{SNR}) = 1 - F_n(1) \tag{9} $$

Figure 4 shows the results of the analytical expression (9) for $m \in \{0, 1, 2, 3\}$ and for infinite precision. From this analysis, the theoretical data fraction word length of a QPSK modulation can be deduced and is presented Figure 5 for different SNR values. The fraction word length increases with the SNR and up to three bits at the receiver input can be saved depending on the channel conditions.

In order to verify this theoretical analysis, simulations of the BER for a QPSK modulation in different quantization and channel conditions have been performed and are presented in Figure 6. As the emitted signal is binary (for real part and imaginary part), a fractional word length of zero, one, two and three bits will be used. Again, the more the SNR increases, the longer the fractional word length required to meet the precision criteria (less than 10% of
error between the BER in infinite and finite precision). Figure 6 confirms that the demodulation processing needs one bit at 0 dB, two bits at 4 dB and more than three bits at 10 dB. After the approximation to finite word-length precision, the quantization error must be negligible compared to channel noise SNR. Thus, a higher number of bits is needed for fractional part when the noise level is lower.

With a dynamic adaptation of the fixed-point specification according to the SNR, a reduction of the word-length can be achieved compared to a classical approach. For instance, in this case, a classical design needs a word length of 2 + 5 = 7 bits. With the proposed dynamic precision scaling, 2 + 1 = 3 bits at 0 dB and 1 + 5 = 6 bits at 10 dB. Between these two SNR values, three bits at the receiver input can be saved.

D. Precision analysis for multiple quantization errors at the receiver

In the general case, the receiver includes some processing and therefore generates multiple rounding quantization noises \( q_1, q_2, ..., q_K \). Due to the central limit theorem, the sum of these noises is then considered gaussian and has the following pdf:

\[
f_Q(x) = \frac{1}{\sigma_Q \sqrt{2\pi}} \exp\left(-\frac{x^2}{2\sigma_Q^2}\right) \tag{10}\]

where

\[
\sigma_Q^2 = \sum_{i=1}^{K} \sigma_{q_i}^2 = \sum_{i=1}^{K} \frac{q_i^2}{12} \tag{11}\]

Thus, the total noise including the processing and quantization errors at the receiver has the following probability density function \( f_N(x) \):

\[
f_N(x) = \frac{1}{\sqrt{\sigma_c^2 + \sigma_Q^2}} \exp\left(-\frac{x^2}{2(\sigma_c^2 + \sigma_Q^2)}\right) \tag{12}\]

and the following probability distribution \( F_N(x) \):

\[
F_N(x) = \frac{1}{2} \left(1 + \text{erf} \left( \frac{x}{\sqrt{\sigma_c^2 + \sigma_Q^2}} \right) \right) \tag{13}\]

In case of BPSK and QPSK, the probability of bit error for multiple quantizations is:

\[
\text{BER}(\sigma_Q, \text{SNR}) = 1 - F_N(1) = \frac{1}{2} \text{erfc} \left( \frac{1}{\sqrt{\sigma_c^2 + \sigma_Q^2}} \right) = Q\left( \frac{1}{\sqrt{\sigma_c^2 + \sigma_Q^2}} \right) \tag{14}\]

The case of the WCDMA receiver presented in the following section corresponds to this theoretical BER.
IV. WCDMA RECEIVER

A. Presentation

WCDMA is a standard for the third-generation of cellular network which is based on DS-CDMA (Direct Spread CDMA) technology. In these systems, a rake receiver is used to counter with the effects of multi-path fading. A finger is allocated to each path to decode the symbol associated with the path. One important component associated with the rake receiver is the path searcher. A path searcher finds the delay of different paths, which is then used to synchronize the input signal with the code generated in the receiver and thus to obtain an optimal combination of received energy.

In WCDMA, there are two layers of spreading codes [11]: channelization code and scrambling code. The channelization code $C_{ch}$ is used to achieve orthogonality between channels when time-shift is equal to 0. The length of $C_{ch}$ is SF, the spreading factor. The scrambling codes used in uplink are Gold codes $S_G$. The input data $d_i$ is multiplied with the spreading codes, and the transmitted signal $T_x$ is:

$$T_x = d_iC_{ch}S_G$$

In a multi-path (time dispersive) Rayleigh channel, the global received signal $R_x$ is the sum of elementary signals $R_{x_{k,t−τ_k}}$ for different channel paths:

$$R_{x_{k,t−τ_k}} = a_kT_x + n_k + i_k$$

where $τ_k$ is the delay of $k^{th}$ path in the channel, $a_k$ is the attenuation, $n_k$ and $i_k$ are additive white gaussian noise and channel interference respectively.

At the receiver, the delay $τ_k$ of each path is estimated by the path searcher module. Then, the symbol are decoded with a rake receiver made-up of different fingers, one for each estimated path. Each path is processed by a finger and is despreaded synchronously by multiplying with conjugated spreading codes. Then, the finger sums up the despreaded signal on a symbol duration. Due to the code properties, the useful signal is amplified and the noise and interferences become negligible. SF is then the processing gain of spreading spectrum. In our case, a spreading factor of SF = 16 is used for DPDCH, which is the data symbol frame of the WCDMA norm in the context of UMTS/3G wireless communications at the speed of 240 Kbps without channel coding. The processing gain is thus equal to 12 dB.

A Rayleigh channel model respecting the 3GPP channel case 3 [12] without Doppler effect is used, corresponding to a multi-path fading with four path components (gain, delay): (0 dB, 0 ns); (-3 dB, 261 ns); (-6 dB, 521 ns); (-9 dB, 781 ns).

B. Symbol detection

1) Range estimation: The flow graph of one finger of a rake receiver is presented in Figure 7. In the finger, the correlation between the input signal and the codes is used to amplify useful signal to detect the transmitted symbol.

The correlation process increases the range of useful signal, but not that of noise. An approach is then proposed to determine more accurately the data dynamic range. Before the correlation process, the whole useful signal plus noise is considered. After this process, only the useful signal is taken into account when calculating the dynamic range.

In equation (16), the received signal is:

$$R_x = \sum_{k} R_{x_{k,t−τ_k}} = \sum_{k} a_kT_{x_{t−τ_k}} + n_i$$

where $n_i$ is interference plus noise, which can be considered gaussian with variance $σ^2$. Assuming that a user has one DPDCH channel, from (15), $d_iC_{ch}$ take values in $\{±1 \pm i\}$. Thus $T_x = d_iC_{ch}S_G \in \{±2, ±2i\}$. In our simulations, $T_x$ is normalized into $\{±1, ±i\}$.

The increase of users in a cell rises the interferences. These interferences are processed as noises and thus increase the SNR. To take account of the different cases, a few number of communications in a cell with good transmission conditions and a great number of users with bad transmission conditions, a great range of SNR is considered (0 to 25 dB).

The input $s(n)$, corresponding to the received signal $R_x$, is normalized to have the maximum amplitude of both real and imaginary parts one. Because a gaussian noise $σ^2$ has 99.7% of its values in $[-3σ, 3σ]$, assuming the real and imaginary parts of $\sum a_kT_{x_{t−τ_k}}$ are in $[-1, 1]$, the input is considered in $[-1−3σ, 1+3σ]$. The normalization process is thus implemented by dividing the input by $1+3σ$ then cut off by 1.

Multiplication with complex scrambling code $c_{ch}(n)$ results in doubling the amplitude. Each real and imaginary part is then multiplied with OVSF code, which does not change absolute value. Averaging in 256 chips results in an important attenuation of noise, which leads the output to signal-only. Thus, to evaluate the dynamic range after the correlation, only the useful signal is taken into account. Both real and imaginary parts of $s_5(k)$ are in $[−\frac{2}{1+3σ}, \frac{2}{1+3σ}]$. As a result, the channel estimation coefficient $α_i$ is in $[−\frac{2}{1+3σ}, \frac{2}{1+3σ}]$. 

![Fig. 7. Data flow graph of a symbol decoder in the WCDMA receiver.](image-url)
The complex multiplication of $\hat{s}_i$ and $s_1(n)$ results in $[-\frac{4}{1+3\sigma}, \frac{4}{1+3\sigma}]$. Multiplication with real OVSF code does not change dynamic range. Then, accumulation results – with the same explication as above – in $[-\frac{1}{1+3\sigma} \cdot SF, \frac{1}{1+3\sigma} \cdot SF]$ for the real part and in $[-\frac{4 \cdot 256}{(1+3\sigma)^2}, \frac{4 \cdot 256}{(1+3\sigma)^2}]$ for the imaginary part.

Estimated and simulation based values are presented in Figure 8. The $acc_I$ and $acc_Q$ data are studied as they have the largest dynamic range and the largest variation when SNR changes. In both simulation and estimation, there are a difference of 4 bits between 0 dB and 15 dB, 7 bits between 0 dB and 25 dB.

It is noticed that there are a difference from one to two bits between estimated and simulated results. This difference is explained by the channel model used in the simulation. If a single path channel model, for example, is used, the difference is less than 1 bit. Moreover, the analytical estimations are more pessimistic.

2) Precision analysis: Similar to (11), the total quantization noise in a symbol decoder can be presented by the sum of quantization noises in each step. Suppose the same bits $N$ are used in every quantization, some calculations show that:

$$
\sigma_{Q,I}^2 = \frac{2^{-2N}}{12} \times 9 \cdot SF \tag{18}
$$

$$
\sigma_{Q,Q}^2 = \frac{2^{-2N}}{12} \times 9 \cdot 256 \tag{19}
$$

The normalization of the signal has an impact on noise: $\sigma^2 = \sigma^2 \times \frac{1}{(1+3\sigma)^2}$. The total noise $\sigma_{Q,I}^2 + \sigma_{Q,Q}^2$ has then less dependence on $\sigma$ (thus, SNR). In fact, it is showed in Figure 9 that about 10 bits are sufficient to approximate floating-point precision when $E_b/N_0$ varies between 0 dB and 10 dB.

In conclusion for the rake receiver, the optimised word-length of the output is equal to 19 bits at 25 dB and 12 bits at 0 dB. Thus, between the two SNR values 7 bits can be saved.

C. Path searcher

A path searcher (PS) with power delay profile (PDP) algorithm is now studied. This module analyses in temporal windows of a chip length the correlation between the input signal and the code generated inside the receiver. This PS module achieves the coarse-grain delay synchronization and then the fine-grain synchronization will be carried-out by the Delay Locked Loop (DLL) inside each finger. The processing is focused on the control channel and the unknown complex amplitude $a_k$ is removed by computing the real and imaginary part and then taking the module.

The signal flow graph is presented in Figure 10. First, the PS module computes the correlation between the input signal and the code associated with the control channel. Then, the square module of the correlation is computed and this value is compared to an adaptive threshold $t(t)$. A path is detected if this value is greater than the threshold. The adaptive threshold is proportional to the average of all the correlation values.

1) Range estimation: Firstly, the filtered received data RX is normalized into $[-1, 1]$ with the same method that the one presented in IV-B.1. It is then multiplied with complex conjugate of spreading code $C_{ch} S_n^*$ and results
in \([-2, 2]\) for each real and imaginary part. Accumulation along with \(N_W\) symbols \((N_W:\ \text{correlation window size})\) – only the signal is summed up significantly – results in \([-2^{N_W}, 2^{N_W}]\), and then averaging into \([-\frac{2}{1+3\sigma}, \frac{2}{1+3\sigma}]\). All these analyses are summarized by the data flow graph of Figure 10.

The estimated and simulation based dynamic range of each value is presented in Figure 11. Both estimated and simulation based results have a difference of 3 bits in accumulation value between 0 dB and 25 dB, and a difference of 6 bits in power profile.

It is noteworthy that estimated and simulated results differ of 1 or 2 bits. This depends on the channel model and is due to fact that the Nyquist filter is not taken into account, which can slightly modify the values.

2) Precision analysis: In the Path Searcher, quantization noise of each power profile is chi-square distributed with variance of:

\[
\sigma_{Q, PS}^2 = \frac{q^4 N^2}{9} = 2^{-4N} L_c^2
\]  

(20)

where \(L_c = 256\) the length of OVSF code in DPCCH channel.

This PS module is based on the decision theory and classical criterions are used to analyze the performances. The misdetections corresponding to the non-detection of an existing path and the false-alarms corresponding to the detection of a non-existing path are measured.

The results are obtained with a monte-carlo approach. Simulations are performed for different Rayleigh channel models, each having four paths. For each kind of Rayleigh channel, 500 experiments are carried out. In Figure 13, the average number of paths which have been missed are reported. In Figure 12, the average number of non-existing paths which have been detected are reported.

Given that the precision is limited for small word-length, a lot of power values are coded with a value of 0 and thus the average value \(t(l)\) is smaller than in the infinite precision case. Consequently, the number of false-alarms increases when the fractional part word-length decreases. The number of misdetections decreases with the fractional part word-length. This phenomenon appears surprising at first sight, but is due to the threshold reduction when the fractional part word-length is decreased. From these results two conclusions can be drawn. The threshold \(t(l)\) depends on the SNR even in the case of infinite precision. The reduction of the computation accuracy modifies the threshold \(t(l)\) value, and, in our case, the misdetection is too important for low SNR and low computation accuracy. Thus, the parameter \(\alpha\) must be adapted according to the SNR value and the fixed-point specification. This adaptation of the parameter \(\alpha\) according to the SNR will allow the improvement of the misdetection and false-alarm probability.

![Fig. 11. Estimated and simulation based values of range for the Path Searcher.](image1)

![Fig. 12. Number of false-alarms for the Path Searcher.](image2)

![Fig. 13. Average number of misdetections for the Path Searcher.](image3)
V. Conclusion

In this paper, the concept of energy consumption reduction by adapting the fixed-point specification is addressed. The concept and the target architecture were presented. The results show that the global number of bits required to limit the degradation of the BER depends on the SNR at the receiver input. For the rake receiver the difference is around 7 bits between different values of the SNR. For the path searcher, the false-alarm and misdetection probability are used as performance metric. The difference is around 6 bits between the extreme values of SNR. This great difference is due to the square operation which emphasizes the phenomenon. For the future work, the energy consumption associated to each fixed-point specification has to be determined to evaluate the gain in terms of energy consumption of our approach.

References


