DYNAMIC TRANSCEIVERS: ADAPTIVITY AND RECONFIGURABILITY AT THE SIGNAL-DESIGN LEVEL

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ABSTRACT

This paper explores concepts related to radio flexibility with applications to an OFDM-based, short-range, indoor radio transceiver design. The concept of dynamic signal design cultivated in the Wind Flex[1] system is introduced. Based on this approach, two algorithms are proposed that minimize the transmission power while satisfying constraints for a given Quality of Service (QoS) level (as defined at the PHY layer). Simulation results demonstrate the performance gains achieved by the proposed algorithms.

1. INTRODUCTION

We describe here the essential elements of a Flexible, Adaptive and Reconfigurable (FAR) radio-modem framework for slowly varying environments, in order to accommodate various new demands placed on such radios by an assortment of agents. Flexibility in the broadest context refers to the ability to respond to various changes in the requirements or the specifications, either present or future. These can be service or user requirements and their related quality of service (QoS) attributes (for example, data and bit error rates, delays, etc.), environmental conditions (for example, changes in the channel due to mobility, interference from other users or systems, etc.), or system conditions (say, operating frequency band). The users, the operator and the channel are the three agents that may affect the general system operation in a mutually independent manner. Flexibility is thus the toolbox that enables the accommodation of any such circumstances, and therefore comprises a set of techniques in the service of the desirable systemic properties of effectiveness (for example, spectral efficiency), reliability, robustness, scalability, component reusability, spatial coverage, power and cost efficiency, adjustment to channel conditions, etc. As new software (SW) and digital signal processing (DSP) tools are being continuously developed and improved, their impact on the design philosophy of such radios needs to be assessed and their empowerment incorporated. To do so effectively, to create designs that harness this power, and to conceptually harmonize the miscellaneous approaches working in parallel to that effect, have provided the key motivation for the present work. It was the motivating thought behind the Wireless Indoor Flexible High Bitrate Modem Architecture (WF for short) project [1]. Similar concepts have also been explored in the Stingray EU Project, a system for outdoor fixed wireless access.

The main principle distinguishing flexible modem design in the present context from other proposed software define radio (SDR) architectures is the built-in intelligence within the modem, which handles all FAR-related features. Thus, this intelligence directs the run-time adaptation and reconfiguration of the transceiver chain according to the physical-layer QoS requirements and the channel state, and does so with novel algorithms, some of which are described below. The physical-layer design approach followed in WF may be named Dynamic Signal Design (DSD). It is related closely to the class of adaptive signal design techniques already discussed extensively in the literature [2],[3] combined in addition with various reconfigurability aspects of modern transceivers.

The paper is organized as follows: the general conceptual framework for FAR is provided in the next section. In section 3, the DSD concept is described. An example of a DSD approach, followed in the WF system, is given in Section 4, where two algorithms are proposed. Finally, simulation results are presented in section 5 which confirm the performance gain achieved with the proposed algorithms. In the Appendix, a review of the analytic performance evaluation of a turbo coded system is given, along with an extension of this analysis to a system with a static, known, frequency-selective channel. This analysis forms the backbone of the optimization routines (algorithms) mentioned previously in the context of DSD.

2. THE FAR CONCEPT

One of the basic efforts in the WF project were (a) the definition, clarification and elaboration of the FAR related concepts in the context of the indoor application, and (b) a subsequent precise determination of how these concepts can be adopted, specialized and demonstrated in a WF modem. The central notion of flexibility is defined as an umbrella

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concept, encompassing a set of independently occurring design features, such as adaptivity, reconfigurability, modularity, etc., such that the presence of a subset of those would suffice to attribute the qualifying term flexible to any particular system. These features are termed independent in the sense that the occurrence of any particular one does not predicate or force the occurrence of any other. For example, an adaptive system may or may not be reconfigurable, and so on. Thus, and following common practice, a system is called *adaptive* if it can respond to changes by properly altering the numerical value of a set of parameters. It is *reconfigurable* if it can be rearranged, at a structural or architectural level, by a non-quantifiable change in its configuration. Here, "non-quantifiable" means that it cannot be represented by a numerical change in a parametric set. For example, the structural change of going from a serially concatenated turbo code to a parallel concatenated turbo code cannot be represented by a change in a numerical quantity; similarly, the architectural change of replacing the hardware implementation of the IF stage by digitization and software-controlled processing, as in pure SDR, cannot also be represented by a numeric change. Clearly, certain potential changes may fall in a gray area between definitions. For instance, changing the number of sub-carriers (SC) in OFDM may appear as an adaptive change since it is quantifiable, but because it has structural implications at the Fast Fourier transform (FFT) and other levels, it may also be considered a structural - reconfiguration type of change.

WF and Stinrgay have explored the multi-faceted nature of FAR by addressing various aspects of the concept. Adaptive modulation and coding, adaptive space time frequency (STF) coding, weak sub-carrier excision (WSCE), adaptive equalization and frequency-offset/phase-noise compensation, are few such examples in the adaptivity dimension. They comprise an integral part of the optimization procedure, performed in a dedicated engine called *the supervisor*. Such an optimization, which involves intricate QoS negotiations with the higher layers, would be rendered difficult without the help from such efficient adaptive techniques.

3. DYNAMIC SIGNAL DESIGN AT THE PHY LAYER

The increasing demand for better spectral utilization and higher QoS requirements motivate the design of "smarter" communication systems, which are adaptive and adjust (in real-time) the transmission parameters based on the instantaneous link quality, for the ultimate goal of reaching, to the degree possible, the capacity limits of the underlying channel. Typically referred to as "Adaptive Modulation", many algorithms have been proposed and their performance limits assessed [2],[3]. These algorithms comprise the "brains" of controllers or "supervisor" modules in modern transceivers [4], responsible among other things for real-time transmission parameter selection. The purpose of DSD applied to transceivers is to provide the user with a system that can dynamically ("on the fly") find the best possible compromise between a number of contradictory design goals, such as minimum power consumption, robustness against reception errors due to channel variations and interference, spectral efficiency, system capacity and so forth. We note that a system is called ***dynamic*** if it is either *adaptive* or *reconfigurable* (or both) in a real-time sense, based on run-time measurements and resulting actions.

To proceed with a proper formalism, and following the general definition given in [5] for adaptive transmission systems, we will briefly describe the general framework for DSD. Let *Tx* be the transmitter (Tx) that operates in a mode *md*, chosen from a set of available transmission modes *MD* (*md* Є *MD*) by a law L(.), selected periodically at time instants *t=KTa*, where *Ta* is a positive integer termed the "dynamic adjustment period". The following observations apply:

- The set of transmission modes *MD* can be either discrete or continuous and defines the solution space.
- The quantity 1/*Ta* is called "adjustment rate" and determines the frequency (with respect to the symbol rate) at which the system is allowed to adjust.
- L(.) is the law by which *Ta* selects a new mode of operation based on some input parameters, e.g., the channel state estimation, battery level, etc.

Given a *Tx*, the objective of the DSD-defined system is to minimize a selected cost function (based on the optimization criterion), subject to a set of constraints. In order to design a dynamic system, one must take the following steps:

- Define the *Tx* to be used
- Design the environment-state estimator/predictor
- Find the mode selection law L(.)

"Define the *Tx"* means identifying the transmission parameters as well as the parameter-adaptation rules that characterize the system and will be used in the design procedure. Classic examples of basic parameters along with affiliated rules are transmission power (fixed versus variable, discrete versus continuous), modulation schemes (fixed versus variable for every symbol or frame), code (fixed versus variable rate for a target throughput), and so on. "Design the environment-state estimator/predictor" means to define the method for estimating the key environmental parameters that affect system performance. For example, if the key environment is the underlying channel, a channel estimator must be designed or chosen. The quantification of the quality of estimation is also crucial for the design. This also involves the selection of the notification procedure (if the estimation or the decisions take place at the Tx and must be communicated to the Tx). Finally, in order to specify the law, the basic system optimization criterion must be chosen: e.g., maximum capacity, maximum throughput, minimum Tx power, minimum bit error rate (BER), minimum BB processing power, and so on. Then, the definitions of the optimization constraints must follow, for example fixed versus variable power, average versus instant BER, average versus instant throughput, etc. The constraints must be chosen so as to satisfy the target application requirements as well as implementation feasibility. These considerations are most critical since the implementation feasibility aspects are usually ignored. All the above define an optimization problem, the solution to which is the desired DSD Law.

4. EXAMPLE: (WIND-FLEX PLATFORM)

In both systems (WF and Stingray) the basic signal modulation scheme was OFDM [6], along with a powerful turbo coded scheme. Both experienced a fairly static channel that lead to designs that adapt transmission parameters on a frame-time basis.

In Stingray, which is a multiple input multiple output (MIMO) OFDM system, the resulting channel (after STFC decoding) at the receiver (Rx) has small frequency selectivity, basically due to the diversity gain achieved by the Space Time Frequency (STF) coding scheme. The main goal in Stingray is to properly change the STF code in order to maximize the system throughput. Due to the increased flatness of the resulting channel (as opposed to a system with single antennas), the average SNR at the demodulator was the basic parameter for choosing the Tx mode on every frame (=78 OFDM symbols). A reconfigurable STF coding scheme along with the classic adaptive modulation for flat fading channel has been the final design approach. On the other hand, WF system is meant to work in the 17GHz band, and has been measured to experience large frequency selectivity within the 50-MHz channelization. The result is strong performance degradation due to few sub-carriers (SC) experiencing deep spectral nulls where, even with a powerful coding scheme as turbo codes, performance degradation is unacceptable. In order to keep system implementation complexity at a minimum, and also minimize the required channel feedback traffic, two design constraints have been adopted: same constellation size for all SC's, as well as same power for all within a OFDM symbol, although both these parameters are adjustable (adaptive). These constraints are taken into account in the present solutions; see Table 1 for a summary of the relevant choices and parameters.

Parameter space	Parameter Adaptation Rules
Transmission power	Same power over all sub-carriers / variable power per OFDM frame per user slot
Modulation Schemes: BPSK, QPSK, 16QAM, 64QAM	Fixed constellation per OFDM frame per user slot plus Weak Sub-Carrier Excision (on-of bit-loading)
Codes: ½, 2/3 and ¾ - rate punctured turbo code, (13,15) octal	Puncture Rate variability per OFDM frame per user slot

Table 1 Parameter space and adaptation rules

4.1 Algorithms for transmit power minimization

As shown in the Appendix, the BER performance of a coded system is approximated by a non-linear function of the BER performance of the corresponding uncoded part. This fact will be used in the DSD herein. Assume that the system employs four different code-puncturing rates and three different code-block sizes. This is equivalent to having 12 different "codes" available, each with a different performance. A non-adaptive system would have pre-stored the required average SNR needed for these codes in order to achieve the target BER for the channel model of interest. These SNR values correspond to an *average* performance for this channel model. However, the actual needed SNR for each channel realization deviates significantly from the average value. Since the channel dynamics are slow, it makes sense to attempt adaptive parameter selection, based on each channel realization. We can see from equation (11) in the Appendix, that the average demodulation BER can be evaluated in order to approximately predict the coded performance for a given average SNR. The needed average SNR can then be iteratively evaluated for a specific code, constellation, and channel realization to achieve the target coded BER. The first adaptation algorithm proposed has low complexity and limited feedback information requirements:

Algorithm #1

1. Select the code rate, constellation size based on the target bit rate.
2. Read the required uncoded BER (from a look up table) based on the target BER.
3. Find the average SNR needed in order to reach the required uncoded BER
4. Compute the power needed in order to achieve the required average SNR, based on the current average SNR.
5. If required power > maximum available power, re-negotiate QoS (lower the requirements) and go to step 1; else output the power/constellation size/code rate

The average SNR needed, for a given channel realization, in order to achieve an uncoded BER of interest is a solution to the non-linear equation:

$$\sum_{i=1}^{N} e^{-\left((E_s / N_o) H_i^2\right)} = C * BER_{un} \quad (1)$$

where C is a constant depending on the BER approximation and the number of SC's, and BER_{un} is the needed uncoded BER. An iterative method can be used in order to find the solution, such as the bisection algorithm. The main computational requirement is the sum of N exponentials. This computation must take place in each iteration.

The second algorithm proposed employs "hard" (or on-off) bit loading by excluding from transmission the SCs with the smaller channel gains. We call this the WSCE method. The significance of WSCE is the ability to choose between different code rates for the same target rate, a feature absent from the first version above. Let us assume that we order the different pairs ={code rate,constellation}, based on their required SNR, needed to achieve a certain BER performance. It is obvious that this ordering also applies to the throughput of each pair (one will not use pairs that need more power to give lower throughput). For each of these pairs, the fixed percentage of excised carriers is computed, so that they all have the same final (target) throughput. For example, suppose that the choice that achieves the target rate is to use 1/3 code rate and 4-QAM constellation. The neighboring code rates with higher throughput are those with code rates 1/2, 2/3, and 3/4. The percentage of carriers to excise, in order that these codes achieve the *same* target rate, are 34%, 50%, and 65%, respectively. The algorithm must now compute the power requirements for each case, in order to find the choice that meets the target BER with the minimum power.

Algorithm #2

1. Select the competitive triplets of:{code rate, constellation , WSCE%}, based on the target rate.
2. Read the required uncoded BER (from a LUT) for each of the choices.
3. Find the average SNR needed in order to get the required uncoded BER for each choice.
4. Compute the power in order to achieve the required average SNR based on the current average SNR for each choice.
5. If required power > max available power for all the triplets, then re-negotiate QoS and go to step 1; else, output the triplet with the min power requirement.

The extra computation load is mainly due to the channel tap sorting. Proper exploitation of the channel correlation in frequency (coherence bandwidth) can reduce this complexity overhead. Instead of sorting all the channel taps, one can sort groups of highly correlated taps.

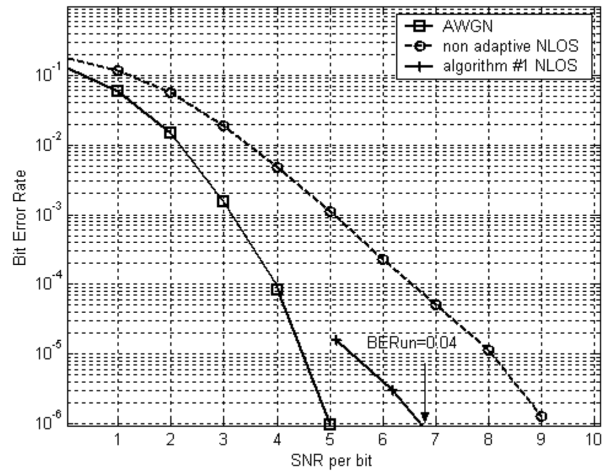

Fig. 1. Simulation results using algorithm #1: Max-Log Map, 4 iterations, S-random-int,NLOS,BPSK,**N**=50,rate=1/2

There are many sorting algorithms in the literature with different performance versus complexity characteristics that can be used, depending on implementation limitations.

It is worth noting that, in the WF project, the turbo decoding module demands over 50% of the overall BB processing power at the Rx. This means that code reconfigurability (that is, changing from one coding scheme to another , say, from turbo to convolutional) may be a worthy goal. It can be shown that code reconfigurability can be accommodated with small adjustments on the second algorithm.

4.2 Simulations

The main simulation system parameters are based on the WF Platform. It has 128 sub-carriers in a 50M-Hz bandwidth (100 active for tx), and the channel model is fully described in [7]. There are two channel scenarios, one line of sight (LOS) with coverage about 100m, and one with no line of sight (NLOS) with coverage about 10m. Both experience deep fades over the 50MHz band, and can be considered static over a frame period (1 frame equals 178 OFDM symbols). The adopted constellations schemes are BPSK, 4-QAM, 16-QAM and 64-QAM, adaptively chosen based on the target throughput requirements. It uses a parallel- concatenated turbo coding scheme with variable rate via three puncture patterns (1/2,2/3,3/4) [8]. The recursive systematic code polynomial used is $(13,15)_{oct}$. Perfect channel estimation and zero phase noise are also assumed herein. Simulation results using algorithm #1 for adaptive transmission power minimization are presented in Figs. 3. The performance gain of the proposed algorithm is shown for BPSK, the code rate equals ½, and the code information block length is 50 bits (**N** = 100bits).

Performance is plotted for no adaptation, as well as algorithm #1 for target BER below 2*10^{-5}, for the NLOS scenario. The performance over a flat (AWGN) channel is also shown for comparison reasons, since it represents the coded performance limit (given that these codes are designed to work for AWGN channels).

The average SNR reduction is more than 2dB. In addition to the transmission power gain, the adaptive schemes practically guarantee the desired QoS for every channel realization. Note that in the absence of adaptation, users experiencing "bad" channel conditions will never get the requested QoS, whereas users with good channel would correspondingly end up spending too much power versus what would be needed for the requested QoS. By adopting these algorithms, one computes (for every channel realization) the exact needed power for the requested QoS, and can thus either transmit with minimum power or negotiate for a lower QoS when channel conditions don't allow transmission. An average 2dB additional gain is achieved by using the second algorithm versus the first one.

5. CONCLUSIONS

Flexibility concepts can be sorted out and defined in a coherent theoretical framework. Like all scientific theories, it needs a "what, why, how" trilogy to be established and pursued. In general, it is easier to discuss and demonstrate adaptivity (both Tx and Rx) and its benefits, whereas reconfigurability is a newer field of inquiry and technology. First "R" targets are easy to grasp, such as min total power consumption; other metrics are also useful and could be pursued. We note that DSD transceivers are not identical to SDR, in particular due to a heavier emphasis on reconfigurable HW, autonomous (self) optimization and adjustment to the environment (DSD includes notions of "smart" and "cognitive" radio). Such a fully developed theory could serve a great variety of systems under current development

APPENDIX
TURBO COFDM PERFORMANCE EVALUATION

First we give a short description of the COFDM system model. The information bit stream is buffered into blocks before entering the channel codec. The encoded bits are distributed across N sub-carriers, as dictated by the bit-loading algorithm. Each individual group of bits inside the *m*-th OFDM block, associated with the *k*-th sub-carrier, is mapped into an appropriate QAM constellation sub-symbol, denoted by $X(k,m)$. This set of N consecutive QAM symbols constitutes an OFDM symbol. The sub-symbols of the *m*-th OFDM symbol are arranged in parallel and fed to a *N*-point IFFT. The output samples of the IFFT are:

$$x_{i,m} = \frac{1}{N}\sum_{k=0}^{N-1} X_{k,m} e^{j2\pi\frac{ki}{N}} \quad (2)$$

These samples are arranged serially with a cyclic prefix extension, modulated by the Tx front-end, and fed to the channel. At the Rx side, by properly apply DFT on the received sampled sequence (prior to stripping off the prefix) yields:

$$Y_{k,m} = X_{k,m} H_k + \hat{n}_{k,m} \quad (3)$$

where H_k is the k-th bin of the N-point DFT of the channel impulse response, and $\hat{n}_{k,m}$ is the projection of the noise to the *k*-th DFT base vector. The sequence is then passed through the equalizer and soft-output demodulation block, which provides the decoder with the appropriate soft-output bits.

Next, we examine briefly the performance analysis of turbo codes in order to find a parametric description when transmitting through a static, frequency-selective channel. The basis is the performance analysis presented in [9] for a Rayleigh fading channel, up to the point where we can continue the analysis with a deterministic channel tap vector, suitable for the (well) estimated channel in our disposal.

A well-known bound of the probability of a word error for the Maximum Likelihood (ML) decoding of a (**N**,**K**) block code is (**N**=codeword length, **K**=message length):

$$P_{word} \le \sum_{d=1}^{\mathbf{N}} A(d) P_2(d) \quad (4)$$

where $A(d)$ is the number of codewords with Hamming weight d and $P_2(d)$ the probability of incorrectly decoding to a codeword with Hamming weight d. Due to complexity issues in the calculation of $A(d)$ for a fixed interleaver, a method has been proposed for deriving an average upper bound: this is done by averaging over all possible interleavers:

$$\bar{A}(d) = \sum_{i=1}^{\mathbf{K}} \binom{\mathbf{K}}{i} p(d \mid i) \quad (5)$$

where $\binom{\mathbf{K}}{i}$ is the number of input words with Hamming weight i and $p(d \mid i)$ is the probability that an input word with Hamming weight *i* produces a codeword with Hamming weight d. The average bound for word and bit error can then be expressed as:

$$\bar{P}_{word} \le \sum_{d=d_{\min}}^{\mathbf{N}} \bar{A}(d) P_2(d) = \sum_{d=d_{\min}}^{\mathbf{N}} \sum_{i=1}^{\mathbf{K}} \binom{\mathbf{K}}{i} p(d \mid i) P_2(d)$$

$$= \sum_{i=1}^{\mathbf{K}} \binom{\mathbf{K}}{i} E_{d|i}[P_2(d)] \text{ and } \bar{P}_{bit} \le \sum_{i=1}^{\mathbf{K}} \frac{i}{\mathbf{K}} \binom{\mathbf{K}}{i} E_{d|i}[P_2(d) \quad (6), (7)$$

where $E_{d|i}$ is the expectation with respect to the distribution $p(d \mid i)$. Having $p(d \mid i)$ for the code of interest, one can evaluate the performance by finding $P_2(d)$ for the channel of interest. In this case, a fixed vector of coefficients represents the channel $\mathbf{H}$. Let us consider two codewords c_0 and c_1 that differ in d bit positions indexed by $(i_1, i_2, \ldots, i_d)$. Assuming BPSK signaling with $\pm\sqrt{E_s}$ amplitude per SC and an AWGN with zero mean and power spectral density $N_0/2$, a good upper bound for the probability of incorrectly decoding the codeword c_0 into the c_1 is:

$$P(c_0 \to c_1 \mid \mathbf{H}) = Q\left(\sqrt{\frac{2E_s}{N_o}\sum_{k=1}^{d} H_{i_k}^2}\right) \le \frac{1}{2} e^{-\left((E_s/N_o)\sum_{k=1}^{d} H_{i_k}^2\right)} \quad (8)$$

To compute the average word error probability, one must average $P(c_0 \to c_1 \mid \mathbf{H})$ over the channel gains $\mathbf{H}$. As mentioned, we are interested in the case where $\mathbf{H} = [H_1, H_2, \ldots, H_N]$ is a deterministic vector. We should then average over all possible sets of $H_{\{i,d\}} = \{H_{i_1}, H_{i_2}, \ldots, H_{i_d}\}$, randomly selected, *d*-dimensional vector of channel gains:

$$P(c_0 \to c_1) \le \frac{1}{2} \sum_{\text{for all } H_{\{i,d\}}} p_{\mathbf{H}}(H_{\{i,d\}}) e^{-\left((E_s/N_o)\sum_{k=1}^{d} H_{i_k}^2\right)} \quad (9)$$

where $p_{\mathbf{H}}(H_{\{i,d\}}) = p_{\mathbf{H}}(\{H_{i_1}, H_{i_2}, \ldots, H_{i_d}\})$ =the probability of selecting this particular *d*-tap. Assuming a proper mixture of the following conditions: (a) the code block size is much larger than the OFDM symbol length; (b) large channel interleavers are employed; (c) the channel does not change appreciably for a large number of OFDM symbols, we can arrive at the approximation:

$$p_{\mathbf{H}}(\{H_{i_1}, H_{i_2}, \ldots, H_{i_d}\}) \approx \prod_{l=1}^{d} p_{\mathbf{H}}(\{H_{i_l}\}) \quad (10)$$

with $p_{\mathbf{H}}(H_{i_l}) = 1/N$, where N is the number of the channel fading coefficients in one OFDM symbol. Using this approximation, and using (9), we get:

$$P(c_0 \to c_1) \approx \frac{1}{2}\left(\frac{1}{N}\sum_{i=1}^{N} e^{-\left((E_s/N_o)H_i^2\right)}\right)^d \quad (11)$$

The term in the parenthesis of (11) is an approximation (a multiplication with 0.5 gives a closer approximation) to the uncoded (demodulated) performance of the system. Substituting this result in equation (7) for the average bit error probability, we conclude that the coded performance of a system depends strongly on the uncoded performance at the SNR of interest. As a consequence, one can base algorithmic decisions on the uncoded performance for the corresponding SNR per coded bit. We have tested this approximation using the codeword weight coefficients evaluated in [10] for the 4-state recursive systematic 1/3 rate turbo encoder with generators (5,7) for two WF channel realizations, shown in Fig. 2. It is thus verified that the approximation works well in high-SNR regions. When higher constellations are used with Gray coding, the same approach may be employed. Each bit in the constellation belongs to a subchannel whose BER performance is equivalent to that of BPSK properly defined for a specific average SNR per bit [11].

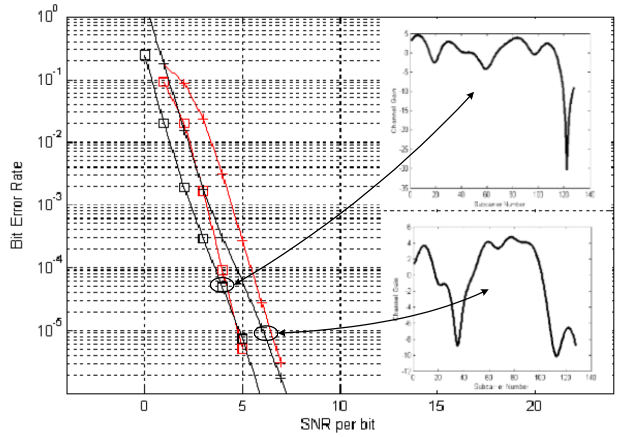

Fig. 2. Simulated and analytical performance for two different WF channel realizations

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