BLIND SIGNAL PARAMETER ESTIMATION FOR THE RAPID RADIO FRAMEWORK

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ABSTRACT

There are many instances where it is desirable or even essential to rapidly build a functional radio receiver to recover symbols from an unknown modulated source. The term "Rapid Radio" refers to a demonstrable analysis environment and receiver implementation methodology for rapid deployment. An automated tool for signal analysis requires several stages for the estimation and classification of the RF signal parameters and modulation type. To fulfill the requirements of the first stage in the case of single carrier linear modulations, an estimator of the symbol rate, frequency offset, and roll-off factor based in frequency-domain analysis is proposed. An estimate of the SNR is obtained as a by-product of the method. The technique was validated with experiments over-the-air.

INTRODUCTION

Hardware realization of a self-configuring radio comprises several stages, including the implementation of an RF front end, estimation of the modulation parameters of the signal, and modulation classification. Parameter estimation for uncharacterized or unknown modulated data streams is a key step in many spectrum management functions like signal intelligence, channel allocation, and interoperability activities. A particular example is a self-configuring radio that can sense spectral activity, blindly analyze participants, formulate a model for a foreign transmitter, and create a corresponding receiver on the spot. While extensive literature is devoted to the classification of modulation based on likelihood and feature methods [1], scarce published work exists on how to estimate the symbol rate, carrier frequency, and spectral shaping roll-off factor once a carrier has been detected, under the hypothesis that it is a single carrier linearly modulated signal.

The estimates of the carrier frequency f_c, symbol rate f_s, roll-off factor α, and carrier plus noise to noise ratio *(C+N)/N* can accurately describe the front end of a radio receiver suited for the signal of interest [2]. Analysis in the frequency domain seems the most appropriate method to obtain the aforementioned parameters; however, it is known that the Discrete Fourier Transform (DFT) is not a consistent spectral estimate [3]. In this paper, the general concept of Rapid Radio is presented, along with a description of a parameter estimation technique. The simulation results are compared with transmission tests carried out using FPGA configurable computing resources to implement the transmitter and receiver using a 70 MHz intermediate frequency, as well as a 2.05 GHz RF front end for transmission over the air.

PREVIOUS WORK

The algorithm proposed by Koh [4] focuses on the estimation of the symbol rate of an unknown linearly modulated signal, using the squared envelope of the passband signal rather than the baseband PAM signal, and performing the Fourier transform to detect spectral lines appearing at DC and at the symbol rate.

In Xu [5], the method proposed to estimate the roll-off factor and the symbol rate of linearly modulated signals is preprocessing the power spectrum by subtracting the noise floor and performing an IFFT. The roll-off factor is estimated as the ratio between the second and the maximum peak. A coarse estimate of the symbol rate is obtained searching for the first minimum of the IFFT magnitude, and a fine estimate is obtained by least squares fitting in the time domain. The estimation of the

Carrier Frequency Error and the SNR is not considered, although SNR estimates can be obtained from the power spectrum histograms proposed at the preprocessing stage.

Hamkins and Simon [6] present a complete system for an autonomous radio, able to detect the front-end parameters of the signal, classify the modulation type, and reconfigure itself for demodulation. They propose an iterative message-passing architecture to calculate fine estimates. This work, however, assumes that the parameters belong to a finite set of values allowed in NASA deep space applications.

THE RAPID RADIO FRAMEWORK

In many events, the quick building of a radio receiver is required to extract information from a signal with unknown modulation parameters. Traditional approaches to accomplish this task require signal analysis, design of a radio receiver, and construction of a prototype. A fairly efficient design can be obtained under this paradigm. However, if instant results are required, it is possible to trade efficiency, measured by area and power consumption, by speed of implementation. The Rapid Radio framework [7] conducts a methodology of design by analysis: the signal is analyzed in a process where several tools are made available to the human. Using a graphical interface, the partial results are presented to the user, who must approve the result of intermediate steps. The approved result of an intermediate outcome is written into an XML description of a radio receiver. A full hardware-independent Radio Description File is then assembled as a result of the analysis process, and can be used to quickly implement a radio receiver in a prototyping platform, such as FPGAs.

SIGNAL IDENTIFICATION AND ISOLATION

A user who is monitoring the spectrum can find the signals of interest, tag the desired signals, and record the initial and final frequencies for each. The process is illustrated in Figure 1. This preliminary step allows the calculation of a coarse downconversion stage, which is used to obtain a quasi-baseband signal, as illustrated in Figure 2. The complex quasi-baseband signal can be downsampled to maximize its resolution in the frequency domain and to minimize the number of samples per observation window.

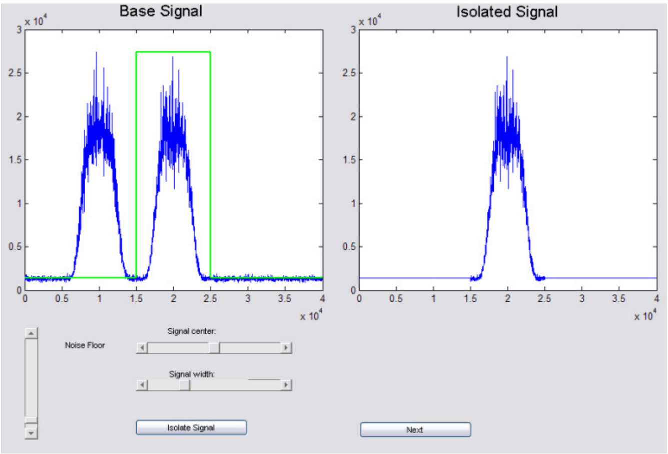


Figure 1. Tagging of carriers in spectrum.

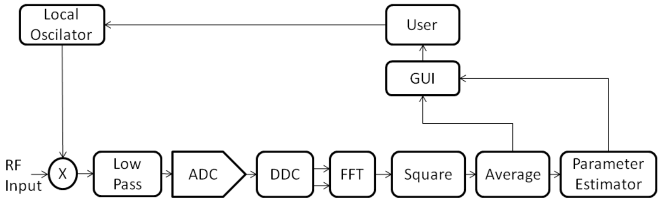


Figure 2. Setup for signal analysis.

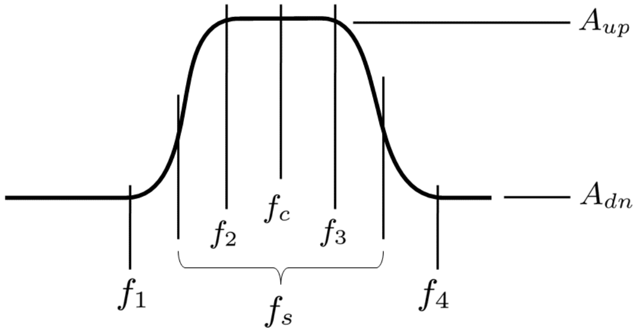
The downconverted and downsampled complex signal is used to estimate the power spectral density (PSD) in the frequency domain using the Discrete Fourier Transform (DFT) over an observation window of N consecutive samples; however, the PSD estimate obtained using the direct DFT suffers from several problems. First, it has a high variance and is non-consistent, which means that the estimate variance is not reduced even if the value of N is increased [8]. While there are several solutions to this problem, they fall into two categories: spectral smoothing and spectral averaging. Spectral smoothing demands fewer computations, but distorts the spectrum shape. Spectral averaging, also termed the Bartlett method, requires the calculation of an average over L PSD estimates, implying a higher computational load, and producing the benefit of not distorting the spectral shape. Therefore, it is the proposed method to obtain a spectral representation $P(k)$ of the signal PSD at the frequency sample k. The second problem of the DFT as a spectral estimator is the dependency of the estimate variance with the amplitude strength.

SPECTRAL SHAPE DESCRIPTION

The method proposed to estimate the RF front end parameters of the signal is based in fitting the ideal spectral shape of the carrier signal under the hypothesis of a modulation type. This concept is evaluated assuming a linear digital modulation with root-raised cosine pulse shaping defined by the analytical function $y(f(k),\beta)$, which is evaluated at the frequencies $f(k)$ and is defined by the set of parameters $\beta = \{f_c, f_s, \alpha, A_{up}, A_{dn}\}$, as follows:

$$y(f(k),\beta) = \begin{cases} A_{dn} & f(k) \in [0, f_1) \\ A_{dn} + \mu\left[1 - \cos\left(\dfrac{f(k) - f_1}{\alpha f_s}\right)\right] & f(k) \in [f_1, f_2) \\ A_{up} & f(k) \in [f_2, f_3) \\ A_{dn} + \mu\left[1 + \cos\left(\dfrac{f(k) - f_3}{\alpha f_s}\right)\right] & f(k) \in [f_3, f_4) \\ A_{dn} & f(k) \in [f_4, \infty) \end{cases} \tag{1}$$

where $\mu = (A_{up} - A_{dn})/2$. The set of parameters of the PSD of the root-raised cosine spectrum is presented in Figure 3.



$$f_1 = f_c - \tfrac{1}{2}(1+\alpha)f_s \qquad f_2 = f_c - \tfrac{1}{2}(1-\alpha)f_s$$
$$f_3 = f_c + \tfrac{1}{2}(1-\alpha)f_s \qquad f_4 = f_c + \tfrac{1}{2}(1+\alpha)f_s$$

Figure 3. The analytical spectral shape.

Observe that a spectral representation can be augmented to track more than one carrier, as well as detecting a signal by comparing the values of A_{up} and A_{dn} against predefined thresholds. Consider also that the signal plus noise to noise ratio $(C+N)/N$ can be estimated by A_{up}/A_{dn}; therefore, an estimate of the SNR can be obtained as:

$$SNR \cong \frac{A_{up} - A_{dn}}{A_{dn}} \tag{2}$$

We can observe that the root-raised cosine shape is non-linear in its parameters. Further, while the expected value of the PSD estimate is unbiased, the noise variance is unequal over the frequency range. These conditions can be numerically expressed as:

$$E\{P(k)\} = y(f(k), \beta^*) \tag{3}$$

and,

$$Var\{P(k)\} = \frac{1}{L}[y(f(k), \beta^*)]^2 \tag{4}$$

where β^* represents the true set of parameters. An analysis of the spectral shape properties for fading channels is performed in [9].

ALGORITHM FOR FEATURE EXTRACTION

A non-linear curve fitting algorithm can be used for estimation of the parameter set, using the values that minimize the mean squared error between the periodogram $P(k)$ and the signal description $y(f(k),\beta)$. This condition may be expressed as:

$$\hat{\beta} = \arg\min_{\beta}\left\{\|y(f(k),\beta) - P(k)\|_Q^2\right\} \tag{5}$$

where Q represents a diagonal positive definite weighting matrix required to compensate the unequal estimate variance. It is defined as:

$$[Q]_{ij} = \begin{cases} \dfrac{L}{[y(f(k),\beta)]^2} & i = j = k+1 \\ 0 & i \neq j \end{cases} \tag{6}$$

A non-linear optimization algorithm is required to find an estimate $\hat{\beta}$. Iterative methods over linear approximations of the model are commonly used [10]. Such approximations are based in second order models of the squared error surface, described by the Jacobian and the Hessian matrices. The Jacobian $J(\beta)$ is calculated as the derivative of the analytical function with respect to each of the parameters, and is therefore defined as:

$$[J]_{ij} = \frac{\partial y(f(i-1),\beta)}{\partial \beta_j} \qquad (7)$$

where β_j represents the jth entry of the set of parameters β. The Hessian matrix is approximated in terms of the Jacobian matrix as:

$$H(\beta) = J(\beta)^T J(\beta) \qquad (8)$$

where $\{\,^T\}$ represents the transpose operator.

The Levenberg-Marquardt algorithm was chosen for the iterations of the algorithm, given its computational simplicity. The nth iteration is expressed as:

$$s_n = -\left[J(\hat{\beta}_n)^T Q_n J(\hat{\beta}_n) + \lambda_n I\right]^{-1} J(\hat{\beta}_n)^T Q_n r_n \qquad (9)$$

$$\hat{\beta}_{n+1} = \hat{\beta}_n + s_n \qquad (10)$$

where $\hat{\beta}_n$ represents the updated estimate of the parameter set, s_n is the iteration update, Q_n is the weighting matrix based on the nth parameter estimate, I is the identity matrix, λ_n is an adaptable damping factor depending on the degree of trust of the current iteration, and r_n is the residual $r_n = y\big(f(k) - \hat{\beta}_n\big) - P(k)$. The iteration stops when $\|r_n\| < \varepsilon$, where ε is some small positive real number.

OVER-THE-AIR EXPERIMENTS

Experiments of the method were conducted over-the-air using an FPGA system for signal generation and acquisition, and an RF front end at 2.05 GHz. A schematic of the receiver is presented in Figure 4. Test signals were generated with a symbol rate of 250 kHz and a roll-off factor of 0.75. While digital upconversion is required to achieve the 70 MHz IF signal required by the RF front end, bandpass aliased sampling was implemented with a frequency of 8 MHz, mapping the 70 MHz IF to an aliased band of 2 MHz in the digital circuitry. Downsampling to 1 MHz was performed to enhance the spectral resolution of the PSD estimate. A variable attenuator at the transmitter side was used to change the signal level and hence the signal to noise ratio. Each experiment was repeated 40 times to obtain an estimate of the variance of the method. A summary of the values used for experiments is presented in Table 1.

Table 1. Parameters used for simulation and for over-the-air tests.

Parameter	Value
Sampling Frequency	1 MHz
Carrier Frequency Error	0
Symbol Rate	250 kHz
Roll-off factor	0.75
DFT Window Size	2^{10}
Number of Observation Windows	32
Number of repetitions	40

PERFORMANCE

The theoretical performance of the method can be approximated [10] by Equation (11) assuming a multivariate Gaussian distribution for the estimate vector $\hat{\beta}_n$ given by:

$$\hat{\beta}_n - \beta^{\star} \sim \mathcal{N}\left(0, \sigma^2\left[J^T(\hat{\beta}_n) Q_n J(\hat{\beta}_n)\right]^{-1}\right) \qquad (11)$$

where the value of the variance term is σ^2=1 due to the election of the inverse variance of the periodogram estimator as the diagonal of the weighting matrix Q_n. The diagonal of the covariance matrix gives the theoretical variance of the estimates. The performance of the method was measured in terms of the bias and variance of the estimates obtained from simulation and from over-the-air tests. The results for the bias are presented in figures 5 to 7. The results for the standard deviation of the method are presented in figures 8 to 10. A variance floor is observed as a consequence of the non-consistency of the FFT as a power spectral density estimator.

The estimates of the carrier frequency error bias and standard deviation presented in figures 5 and 8, respectively, are normalized to the symbol rate, as the requirements for the carrier recovery circuit are determined by the magnitude of the phase shift presented during the duration of one symbol.

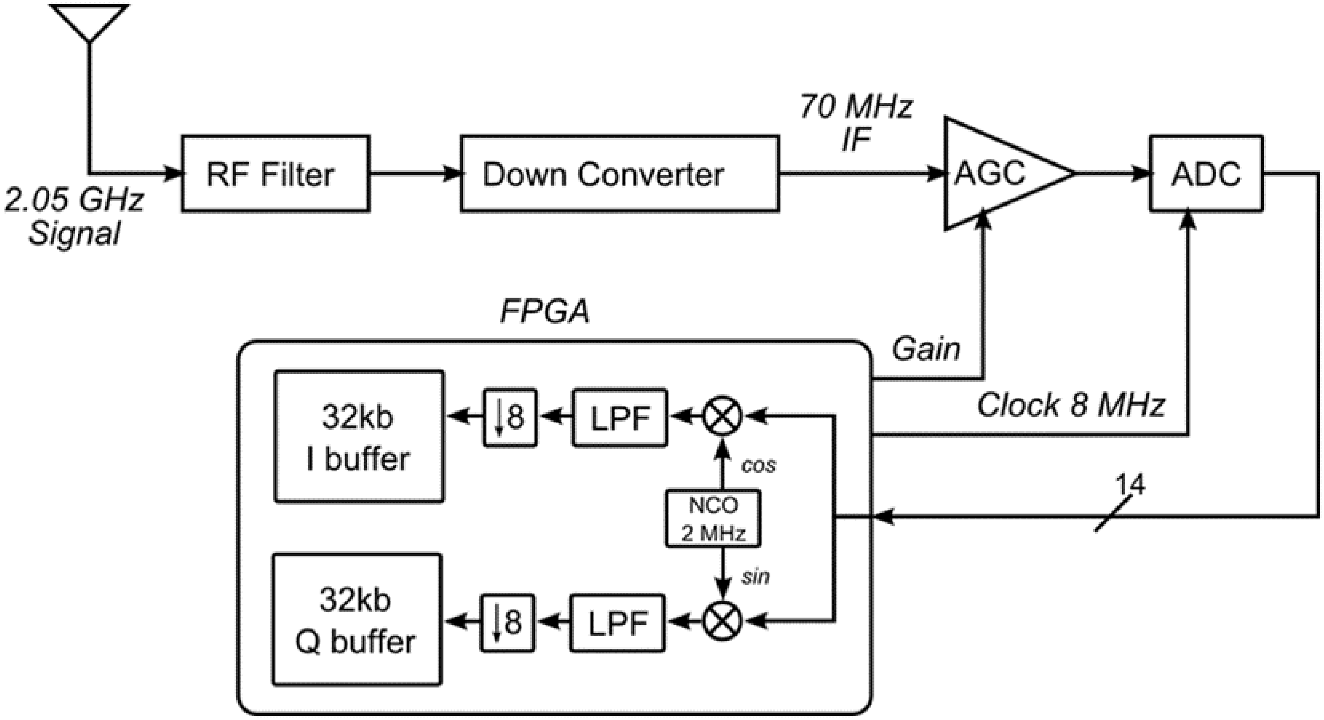


Figure 4. Schematic of the receiver.

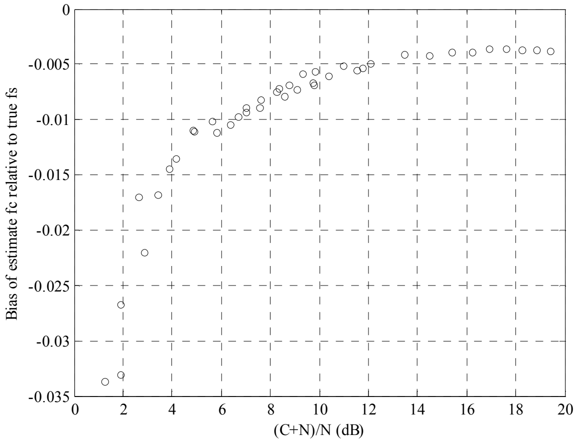


Figure 5. Bias of the frequency error estimate normalized to the symbol rate.

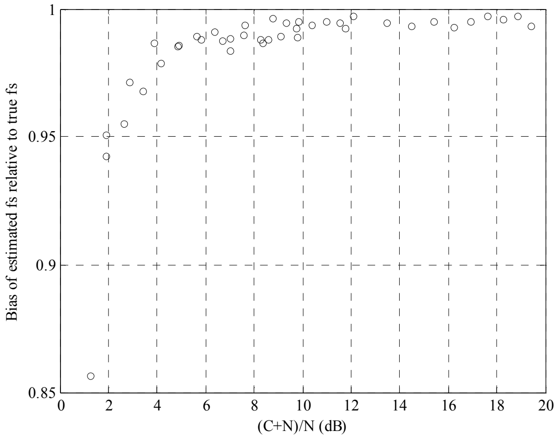


Figure 6. Normalized bias of the symbol rate estimate.

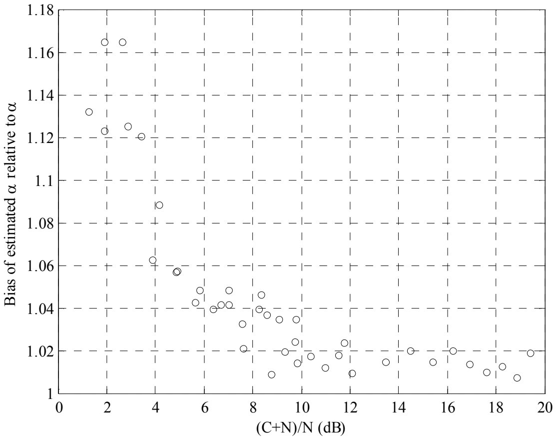


Figure 7. Normalized bias of the roll-off factor estimate.

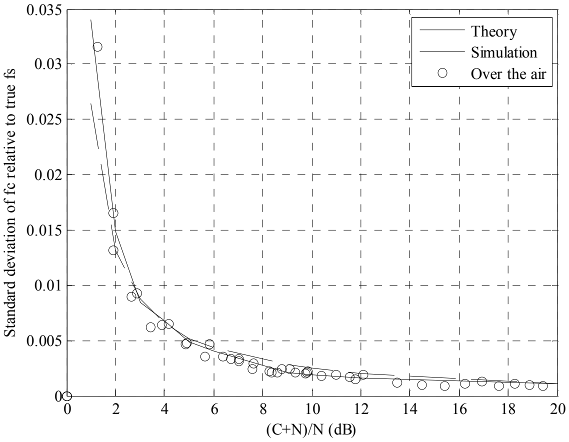


Figure 8. Standard deviation of the frequency error estimate normalized to the symbol rate.

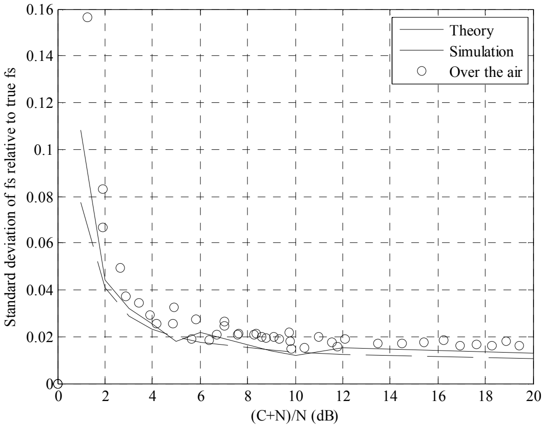


Figure 9. Normalized standard deviation of the symbol rate.

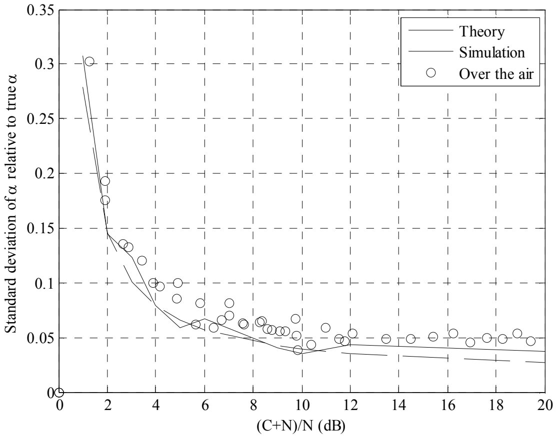


Figure 10. Normalized standard deviation of the roll-off factor estimate.

RAPID RADIO USE MODEL

Several applications can be derived from the general idea of fitting analytic functions in the frequency domain. Some of them are discussed in this section.

The method of parameter extraction from spectral fitting can be used for different spectral shapes; the resulting error metric between the PSD estimate and a set of candidate analytic functions can be used for hypothesis testing purposes when the modulation type is unknown.

The case of multiple carriers can be implemented assuming a shared noise floor. An advantage of this possible expansion is to allow partial spectral overlap between pairs of adjacent signals.

For low SNR scenarios, cyclostationarity is widely used. The application of spectral fitting to the cyclic spectrum is a clear expansion to allow the measurement of the symbol rate and roll-off factor.

As suggested, SNR estimation is also an application of the proposed method. As it involves the division of two random variables, the statistic behavior of the resultant random variable is a subject of further investigation.

The iterative nature of the algorithm is also an attractive feature for the tracking of time-varying signals. Fading profiles can be obtained using this method, as well as frequency changes caused by Doppler or by oscillator instabilities.

CONCLUSIONS

A method for extraction of signal parameters was devised as a first stage for the signal analysis algorithm of the Rapid Radio method. The method relies on the use of the unmodified periodogram to obtain PSD estimates, and on non-linear optimization algorithms that converge to generalized least squares estimates. Close agreement with the theoretical values of variance indicates that the method is usable under real conditions, and that can be expanded to cover scenarios with multiple carriers, heterogeneous modulation types and low signal to noise ratios.

ACKNOWLEDGMENTS

The authors would like to thank the Harris Corporation, Government Communications Division for supporting this research.

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