A RECONFIGURABLE SDR RECEIVER FOR MULTI-MODE GNSS APPLICATIONS

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ABSTRACT
This paper describes a novel reconfigurable Software Defined Radio (SDR) receiver for the next-generation GPS and Galileo Global Navigation Satellite Systems (GNSS). The proposed receiver utilizes variable-rate BandPass Sampling (BPS) to downconvert the signal of interest to an Intermediate Frequency (IF) thereby simplifying the Analog Front-End (AFE) as well as rendering the simplified AFE mostly free of the perennial RF-impairments. A Continuous-Time (CT) quadrature bandpass sigma-delta (ΣΔ) modulator is used for data conversion which circumvents the need for an anti–alias filter and relaxes the RF band select filter specifications by attenuating adjacent band interference through its loop filter. System level simulations demonstrate that the BPS receiver with a CT ΣΔ–ADC is a feasible option that leads to reduced power consumption and high levels of integration making single chip radio receivers realistically viable.

1. INTRODUCTION
Third-generation (3G) mobile systems have introduced new standards and services, and it is anticipated that 4G will focus on integrating existing wireless standards on a single universal radio platform. With the development of the European navigation system Galileo and the modernization plan for GPS, integration of wireless communication systems with GNSS will become a practical reality thereby providing an important platform for the provision of a number of Location Based Services (LBS) as well as navigation information. It is evident that a flexible radio platform is needed that can be configured to tune into the bands of interest complying with the modulation standard and adapt to a dynamic communication environment to achieve acceptable performance.

The Software Radio (SR), envisioned in [1], is a wishful solution to this demand in which all RF and baseband processing is done in the digital domain. However, this puts extreme requirements on the Analog-to-Digital Converter (ADC) in terms of sampling rate, dynamic range, and power dissipation such that a feasibly conceivable solution will not be available in the near future. An alternative solution is to utilize BPS to relax the requirements of the ADC since the sampling rate reduces to twice that of the information bandwidth. A practical version of SR is the SDR in which the RF signal is sampled after a suitable band selection filter and a Low–Noise Amplifier (LNA) [2]. In order to support the idea of SDR, the ADC should be placed as near to the antenna as possible thereby minimizing the number of analog components required. An SDR platform utilizing a BPS ADC with variable sampling rate seems to be a viable option today to achieve wideband reception. The goal of this paper is to derive the requirements of the BPS receiver front-end by providing a signal chain noise analysis for the next-generation GNSS applications.

The paper is organized as follows. Section 2 introduces BPS for the open service GNSS signals. Section 3 analyzes the design considerations for the receiver and Section 4 presents the experimental results. Finally the concluding remarks are given in Section 5.

2. BANDPASS SAMPLING FOR GNSS SIGNALS
Open service radionavigation signals will be in multiple bands within the next decade, as depicted in Figure 1, offering improved position accuracy and robustness against multipath and interference effects. Therefore, the next generation radionavigation terminal has to be a flexible wideband receiver able to tune into different bands as and when desired and to dynamically sense and adapt to the environmental conditions (i.e. the presence of interferers). However, realization of such a receiver is not feasible today mainly due to the requirements on the AFE and the ADC. The current ADC technology is not mature enough to provide direct RF sampling at Nyquist. On the other hand, utilizing LowPass Sampling (LPS) at IF to process multi-frequency signals is still a challenging task due to the need for off-chip analog filters and high-frequency local oscillators which makes it difficult to integrate the receiver monolithically.

Wideband reception in the receiver may be achieved by employing BPS which is a subsampling technique by intentional aliasing of the signal band of to realize
frequency downconversion. One of the challenges in BPS is determination of the sampling rate.

\[ \frac{2f_u}{m} \leq f_s \leq \frac{2f_l}{m-1} \]

where \( f_u \) is the upper and \( f_l \) is the lower frequency of the bandpass signal, \( B \) is the signal bandwidth, \( f_s \) is the uniform BPS rate and \( m \) is the subsampling factor given by:

\[ 1 \leq m \leq \frac{f_u}{B} \]

When a fixed sampling rate is to be chosen for multi-band signals, the spectral replicas of the signals at different bands must be placed in a particular order in the frequency spectrum to minimize aliasing as before. The Ladder diagram concept reported in [4] has been used for the open service E1–L1, L2C, E5a–L5, and E5b GNSS bands and is illustrated in Figure 2. As can be observed from the figure, the minimum feasible sampling frequency is approximately 165 MHz since going below that will result in two or more bands overlapping. Moreover, if it is desired to incorporate commercial service radionavigation bands, the sampling rate might be significantly increased up to 331 MHz [5]. Compared to LPS, high sampling rates will result in significant overhead that leads to large power dissipation in the ADC as well as the DSP. One solution that can avoid high sampling rates is to subsample and therefore downconvert one navigation band at a time and adjust the sampling frequency with respect to the desired signal and environment conditions. A slight drawback of this approach is the need to accurately generate more than one clock frequency which may be favorable as opposed to working at much higher data rates.

3. DESIGN CONSIDERATIONS

GNSS applications require very strict Noise Figures (NFs) since the received signal power at the antenna is well below the thermal noise floor and it will remain so until a digital correlator at the digital backend successfully tracks the signal. As an example, the signal power level of the Galileo E1 band (32.73 MHz), depicted in Figure 3, has a Signal-to-Noise Ratio (SNR) of \(-26 \, \text{dB}\) at the receiver antenna. This will improve when an appropriate PseudoRandom Noise (PRN) code is successfully correlated with the signal. The improvement is called the processing gain, \( G_p \), which is given by [6]:

\[ G_p = 10 \log \left( \frac{f_{PRN}}{f_d} \right) \]

where \( f_{PRN} \) is the code rate and \( f_d \) is the data rate.
Table I. Signal Levels for Open Service GNSS Bands

<table>
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<tr>
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<tbody>
<tr>
<td>L1/C/A</td>
<td>-127.66</td>
<td>1.023</td>
<td>43.1</td>
<td>-16.7</td>
<td>26.3</td>
<td>16.3</td>
</tr>
<tr>
<td>L2/C</td>
<td>-130.0</td>
<td>1.023</td>
<td>43.1</td>
<td>-19.1</td>
<td>24.0</td>
<td>14.0</td>
</tr>
<tr>
<td>I</td>
<td>-124.0</td>
<td>10.23</td>
<td>53.1</td>
<td>-23.1</td>
<td>30.0</td>
<td>20.0</td>
</tr>
<tr>
<td>Q</td>
<td>-125.69</td>
<td>10.23</td>
<td>53.1</td>
<td>-17.8</td>
<td>25.2</td>
<td>19.0</td>
</tr>
<tr>
<td>E1</td>
<td>-125.0</td>
<td>10.23</td>
<td>53.1</td>
<td>-24.1</td>
<td>29.0</td>
<td>20.0</td>
</tr>
<tr>
<td>E5a</td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
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<tr>
<td>E5b</td>
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</tbody>
</table>

The required NF of the AFE and the ADC can be written as:

\[
NF_{AFE+ADC} = P_{IN} + N_0 + G_p - SNR_{\text{min}}
\]  

(3.2)

where \(P_{IN}\) is the received signal power, \(N_0 = -174\text{dBm/Hz}\) is the thermal noise power density, and \(SNR_{\text{min}}\) is the minimum required SNR to be able to successfully track the navigation signal. Table I presents the signal levels of the open service GNSS bands (using 3.2) when the minimum required SNR at the output of a digital correlator is assumed to be 10 dB. As can be seen, the most stringent NF requirement on the AFE and the ADC is for the GPS L2C signal which is approximately 14dB.

3.1. Analog-to-Digital Converter

Although the ADC comes after the AFE in the signal chain, it needs to be analyzed first in order to derive the requirements of the AFE. Traditionally \(\Sigma\Delta\) modulation techniques have been applied to narrowband signals. Switched–Capacitor (SC) \(\Sigma\Delta\) modulators reported in the literature so far have bandwidths not much greater than 10MHz [7], [8], with the exception of [9] having a bandwidth of 40MHz. Since the output of each SC integrator must settle between two clock pulses and the maximum clock frequency is limited by the required precision, the task of deriving wideband SC \(\Sigma\Delta\) modulator is difficult [10]. On the other hand, recent published CT \(\Sigma\Delta\) modulators operate up to a few GHz [11], [12]. CT \(\Sigma\Delta\)–ADCs having programmable bandwidth and dynamic range have also been reported in the literature [13], [14]. Apart from the wideband E5 signal, current IC technologies are able to facilitate bandpass \(\Sigma\Delta\)–ADCs for the open service GNSS bands. The composite E5 band has 2 subbands, E5a and E5b each with a bandwidth of 20.46MHz which are convenient to process individually.

The prominent benefit of utilizing a CT \(\Sigma\Delta\) modulator, shown in Figure 4, is that the anti–alias filter requirement may be circumvented since it provides inherent anti–alias filtering on the input signal path [15], [16]. This feature can also be exploited to relax the \(Q\)–factor of the RF filter by letting the CT \(\Sigma\Delta\) modulator filter out some of the adjacent band interferers.

The output of the modulator in Figure 4 can be written as:

\[
Y(z) = \frac{H(z)X(z) + 1}{1 + H(z)G(z)} E(z)
\]  

(3.3)

where \(H(z)\) is the CT filter transfer function and \(G(z)\) is transfer function of the Digital–to–Analog Converter (DAC). Equation 3.3 can be written in the \(z\)–domain as:

\[
Y(z) = \frac{H(z)}{1 + K(z)} X(z) + \frac{1}{1 + K(z)} E(z)
\]  

(3.4)

In (3.4) \(K(z)\) is the \(z\)–domain loop transfer function which is defined as:

\[
K(z) = Z \left[ L[1[H(z)G(s)]_s] \right]_{z=1}
\]  

(3.5)
where $T_s$ is the sampling time and $n = 0, 1, 2, \ldots$ is the sampling index. Using (3.4) and (3.5), the Signal Transfer Function (STF) can be defined as:

$$STF(z) = \frac{H(z)}{1 + K(z)}$$  \hspace{1cm} (3.6)

The transfer function for the CT filter can be express as:

$$H(s) = \frac{As}{s^2 + \omega_c s + \omega_c^2}$$  \hspace{1cm} (3.7)

where $A$ is the gain, $\omega_c$ is the angular carrier frequency, and $Q$ is the filter quality factor. The transfer function of the DAC for ease of analysis will be approximated as a zero–order hold:

$$G(s) = \frac{1 - e^{-sp}}{s}$$  \hspace{1cm} (3.8)

where $p$ is the opening aperture. For a Return–to–Zero (RZ) DAC $p = T/2$, and for a Non–Return–to–Zero (NRZ) DAC $p = T$, where $T$ is the sampling period. Substituting 3.7 and 3.8 into 3.6 gives the STF in the $z$–domain [15]:

$$STF(z) = H(z) \frac{1 + e^{-\frac{\omega_c}{Q} z^{-2}}}{1 + ge^{-\frac{0.5}{Q} z^{-2}} + e^{-\frac{\omega_c}{Q} + ge^{-\frac{0.5}{Q}} z^{-2}}}$$  \hspace{1cm} (3.9)

To see the effect of the $Q$–factor on the STF of the E1–L1 band, the magnitude spectrum is plotted in Figure 5 for different values of $Q$.

The attenuation of the anti–alias filter at an image frequency can be approximated as [16]:

$$\text{Attenuation} = \frac{1}{\sqrt{(2Q/m)^2 k^2 + 1}}$$  \hspace{1cm} (3.10)

where $k$ is the index for the image. The graph in Figure 6 illustrates the relation between $Q/m$ ratio and image attenuation. Notice that 20dB attenuation of the first image requires a $Q$ factor that is 5 times the subsampling ratio, $m$.

The resulting SNR of the ADC is the key design parameter in the BPS receivers since the effects of non–idealities will be amplified due to subsampling. The dominant noise sources in an ADC that utilizes BPS are quantization noise, wideband thermal noise, and the S/H clock jitter noise. Assuming that the quantization noise is distributed over the Nyquist band, the quantization noise can be defined as [17]:

$$P_q = \frac{\Delta^2}{12} = \frac{V_{pp}^2}{6N^2}$$  \hspace{1cm} (3.11)

where $\Delta$ is the quantization step, $V_{pp}$ is the peak–to–peak voltage swing of the ADC, and $N$ is the number of quantization levels. The S/H circuit can be modeled as a Switched–Capacitor (SC) in which the input voltage charges the capacitor, $C_m$ through the switch. At the end of the sampling time, the switch turns off and the capacitor holds the input voltage as well as the noise introduced by the on–resistance, $R_{on}$ of the switch. The noise introduced by the sampling switch can be estimated by assuming that the total noise power $kT/C_m$ is sampled by the subsampling rate given as [18]:

$$P_n = \frac{kT}{\bar{C}_m} \times \frac{2B}{F_s}$$  \hspace{1cm} (3.12)
where \( k \) is the Boltzmann constant and \( T \) is the absolute temperature.

The aperture jitter on the S/H clock, defined as the uncertainty in time between the effective sampling intervals, will result in phase noise. The jitter noise power for a given sinusoidal input signal, \( A \sin(2\pi f_c t) \), can be written as:

\[
P_j = P_s \left( \frac{m2\pi f_s \sigma_j}{\pi} \right)^2
\]

(3.13)

where \( P_s \) is the sampled signal power and \( \sigma_j \) is the variance of the jitter error. Notice that the SNR degradation on thermal noise is proportional to the subsampling ratio while the SNR degradation on sampling jitter is proportional to the subsampling ratio squared since \( \sigma_j^2 \ll 1/f_c^2 \).

In CT \( \Sigma \Delta \)–ADCs since both the Sample–and–Hold (S/H) and the quantizer circuits are inside the feedback loop, noise generated by their non-ideal operation can be shaped out. Figure 7 illustrates the output spectrum of the forth order CT and SC \( \Sigma \Delta \) modulators when a sinusoidal input signal (used for convenience to evaluate the SNR) is applied.

3.2. AFE
A block diagram of the proposed GNSS receiver is presented in Figure 8. To the author’s best knowledge, the AFE of the receiver is the simplest architecture among the GNSS receiver front-ends. It consists of adjustable LNA and limiting amplifier (LA) to be able to achieve the required NF and a tunable RF bandpass filter for blocking out–of–band interferers. Realization of the fully integrated tunable RF filters has been reported recently in [19], however, achieving high \( Q \) on a mixed signal integrated solution is still a challenging task. For this reason, we propose a relaxed \( Q \)-factor topology in which adjacent out–of–band interference is attenuated by the CT \( \Sigma \Delta \)–ADC. The RF filter with a bandwidth of 50 MHz will have a \( Q \) factor of approximately 30 for E1–L1 band. This eliminates the need for a SAW filter and makes the RF filter viable with the current IC technology. The simplified AFE also minimizes the signal distortion caused by the RF impairments [20], [21] since quadrature mixer and parallel I/Q paths are eliminated.

5. EXPERIMENTAL RESULTS
In order to study, analyze and evaluate the performance of the receiver, a behavioral model of the RF transmitter and the BPS receiver has been developed in Simulink™ as illustrated in Figure 8. A GPS/Galileo signal generator block [22] generates the desired signal as well as MPSK interference. Sampling clock jitter and thermal noise associated with the S/H circuit is introduced in the CT \( \Sigma \Delta \) modulator.

Figure 7. Illustration of the noise shaping in a forth order (a) CT \( \Sigma \Delta \)–ADC (b) SC \( \Sigma \Delta \)–ADC for the E1–L1 band with a subsampling rate of 32.99 MHz.

Figure 8. Proposed Bandpass Sampling GNSS Receiver Architecture

Figure 9. Behavioral Model of the RF Transmitter and the BPS Receiver

Subsampling rates and the resulting SNR of the 4th order CT \( \Sigma \Delta \)-ADC is determined experimentally and given in Table II in which \( P_N \) denotes the noise power. The required gain and noise budget for the AFE is derived from the requirements of the \( \Sigma \Delta \) modulator. Since the noise floor in the \( \Sigma \Delta \) modulator is around -73dBm, the signal power should be raised above this level in order to get sufficient signal power at the output of the ADC. Table III presents the gain and noise budget for the L2C band which is the worst–case scenario.
Table II. Sampling Rates and Resulting SNR for the GNSS Bands

<table>
<thead>
<tr>
<th>Band</th>
<th>( f_c ) [MHz]</th>
<th>( B ) [MHz]</th>
<th>( F_s ) [MHz]</th>
<th>( P_v ) [dBm]</th>
<th>SNR_{CTE}[dB]</th>
</tr>
</thead>
<tbody>
<tr>
<td>E1-L1</td>
<td>1575.4</td>
<td>4.092</td>
<td>32.99</td>
<td>-74.07</td>
<td>71.06</td>
</tr>
<tr>
<td>L2C</td>
<td>1227.6</td>
<td>2.046</td>
<td>16.53</td>
<td>-73.25</td>
<td>70.24</td>
</tr>
<tr>
<td>E5a-L5</td>
<td>1176.45</td>
<td>20.46</td>
<td>162.41</td>
<td>-73.74</td>
<td>70.73</td>
</tr>
<tr>
<td>E5b</td>
<td>1191.79</td>
<td>20.46</td>
<td>164.53</td>
<td>-74.30</td>
<td>71.29</td>
</tr>
</tbody>
</table>

Table III. Gain and Noise Budget for the AFE (L2C Band)

<table>
<thead>
<tr>
<th>Input</th>
<th>LNA</th>
<th>BPF</th>
<th>LA</th>
</tr>
</thead>
<tbody>
<tr>
<td>Gain [dB]</td>
<td>18</td>
<td>-2</td>
<td>50</td>
</tr>
<tr>
<td>NF [dB]</td>
<td>2</td>
<td>2</td>
<td>10</td>
</tr>
<tr>
<td>Carrier Power [dBm]</td>
<td>-130</td>
<td>-112</td>
<td>-114</td>
</tr>
<tr>
<td>Noise Power [dBm]</td>
<td>-110</td>
<td>-90</td>
<td>-90</td>
</tr>
<tr>
<td>SNR [dB]</td>
<td>-19</td>
<td>-21</td>
<td>-21</td>
</tr>
<tr>
<td>Cascaded NF [dB]</td>
<td>2</td>
<td>2</td>
<td>2.2</td>
</tr>
</tbody>
</table>

6. CONCLUSIONS

The proposed receiver provides a low-complexity and flexible platform which can handle multi-band GPS and Galileo signal acquisition with minimal RF component thereby eliminating and circumventing the cost and impairments associated with them. Exploiting the characteristics of the CT \( \Sigma \Delta \) modulator ADC to perform BPS can simplify the AFE by shifting the IF processing into the digital domain. Reconfiguration in the digital domain can be done in order to achieve dynamic adaptation to the signal environment. System level simulations indicate that the proposed receiver is viable with acceptable performance.

REFERENCES
