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REVIEW ARTICLE

NOISE ALIASING TECHNIQUE FOR SOFTWARE DEFINED RADIO: A REVIEW

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Abstract— In communication the signals area unit processed by the sampling devices while not loss of data. As associate in nursing interface between radio front-ends and digital signal process blocks. Digital radio communications completed by sampling devices. In the idea of software defined radio, radio systems area unit that mixes analog, digital and software technology. One goal of software defined radio is to place the analog to digital convertor as nearest as double to the antenna. Band pass sampling allows one to own an interface between the radio frequency or the upper intermediate frequency signal and also the analog to digital convertor, and it might be an answer to software defined radio. Three types of sources perform degradation present in harmful signal spectral overlapping, noise aliasing, rate systems and sampling temporal order noise. In this Research Paper, Optimized Construction BandPass Sampling (OCBPS) is completely studied with specialise in the noise aliasing drawback in software defined Radio.

Keywords: - AA-Anti-Aliasing; A/D- Analog-to-Digital; BER-Bit Error Rate; BK-Basis-Kernel; CDMA-Code Division Multiple Access; CF-Continuous-Frequency

I. INTRODUCTION

According to the study, there is so far no design for fully integrated multi standard subsampling receivers because of the well-known noise aliasing problem in the subsampling system , although some single chip RF subsampling receivers have been designed for GSM , Bluetooth and 802.11b . BandPass charge sampling with intrinsic FIR moving average operation and a 1^{st} order IIR filter were used to treat the noise aliasing problem. Direct RF sampling by quadrature BPS in voltage mode was used without any specific treatment to noise aliasing. A more general solution compared to the work in using quadrature BandPass charge sampling with composite FIR and IIR filtering was proposed. Both simulation results and circuit implementations on an IF signal have shown that this solution is promising to suppress noise aliasing in subsampling or band pass sampling receivers. However, it is known that other processing blocks in radio receiver front-ends operate in voltage mode. Before using charge sampling, an analog voltage signal needs to be first converted to a current signal by a Trans-conductance cell, but it is not necessary for using bandpass voltage sampling and the corresponding front-end receiver architecture is simpler. In this synopsis, bandpass voltage sampling is mainly discussed and also compared with bandpass charge sampling. Three sources of performance degradation in bandpass sampling systems, harmful signal spectral overlapping, noise aliasing and sampling timing jitter, are comprehensively studied. With respect to noise aliasing problem, the theory of generalized BPS in voltage mode is proposed, including the examples of Generalized Quadrature BandPass Sampling (GQBPS) and Generalized Uniform BandPass Sampling (GUBPS). GQBPS and GUBPS perform also FIR filtering that can use either real or complex coefficients besides sampling. The input signal of GQBPS and GUBPS is an RF or a higher IF bandpass signal.

1.1 Conventional Receiver Architectures

Superheterodyne Receivers

The conventional radio receiver architecture, superheterodyne, has existed for almost one century and was proposed by Edwin H. Armstrong in the 1910s. In the literature there is usually no distinction between heterodyne and superheterodyne architectures. To "heterodyne" means to mix two frequencies and produce a beat frequency defined by either the difference or the sum of the two. To "superheterodyne" is only to produce a beat frequency defined by the difference of the two frequencies.

Two stages of down-conversion (dual-IF, IF stands for intermediate frequency) based on the theme of superheterodyne is mostly used in today's RF receivers, see Figure 1.1. This receiver translates the signal to a low frequency band by two stages of down-conversion mixing. If the second IF of a dual-IF receiver is equal to zero, the second down-conversion normally separates the signal to *I* (in-phase) and *Q* (quadrature) components for Single-SideBand (SSB) communication systems or frequency-/phase-modulated signals, and the corresponding demodulation and detection are performed at baseband, see Figure 1.2. This down-conversion is realized by quadrature mixers, which have a 90± phase shift between two Local Oscillators (LOs) signals. Any offset from the nominal 90± phase shift and amplitude mismatches between *I* and *Q* components will raise the Bit Error Rate (BER). If the second IF is not equal to zero, the receiver becomes a digital-IF receiver. The IF bandpass signal is processed by an A/D converter, and the *I/Q* mismatches can be avoided by signal processing in the digital domain.

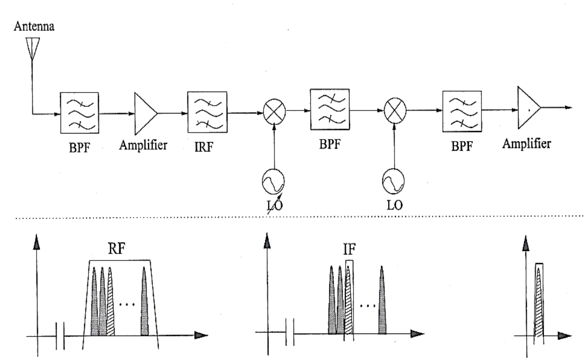

Figure 1.1: Conventional dual-IF superheterodyne receiver architecture

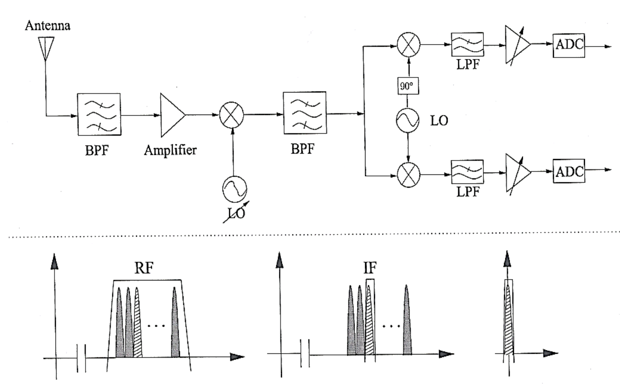

Figure 1.2: Superheterodyne receiver architecture with the second IF being equal to zero.

The choice of IFs influences the trade-off between the image rejection (or sensitivity) and channel-selection (or selectivity). If the IF is high, the image band appears far away from the information band such that the image can be easily suppressed by an Image Reject Filter (IRF). However, the channel selection filter will require a high *Q*-factor to select a narrow channel at a high IF. On the contrary, if the IF is low, the design of the channel selection filter becomes easier but the image band is so close to the information band that it becomes difficult to achieve proper image suppression by a BandPass Filter (BPF). More than one stage of down-conversion makes the trade-off easily achievable. In a dual-IF superheterodyne receiver, the first IF is selected high enough to efficiently suppress the image, and the second IF is selected low enough to relax the requirement on the channel selection filter. The selectivity and sensitivity of the superheterodyne makes it a dominant candidate in RF receiver architectures. Unfortunately, the high *Q*-factors of the discrete-components in the superheterodyne receiver make it difficult to fully integrate the complete front-end on a single chip.

Homodyne Receivers

In a homodyne receiver, no IF stage exists between RF and baseband. The input of the A/D converter is located at baseband, see Figure 1.3. The channel selection filter is just a LowPass Filter (LPF) prior to the A/D converter. The homodyne receiver has two advantages compared to the superheterodyne receiver. First, the architecture is simpler. Second, the image problem can be avoided due to zero IF such that no IRF is needed.

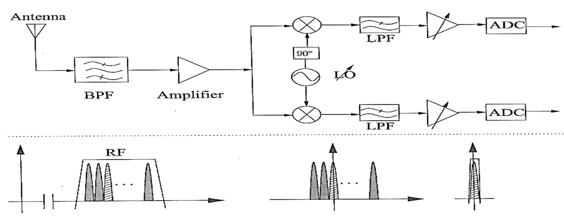

Figure 1.3: Homodyne receiver architecture

The homodyne receiver allows a higher level of integration than the superheterodyne receiver as the number of discrete components is reduced. However, this receiver inevitably suffers from the problems of LO leakage and DC-offset. The output of the LO may lead to the signal input port of the mixer or the Low Noise Amplifier (LNA) due to improper isolation. The leaked signal will be mixed with the output of the LO (i.e., the origin of the leaked signal) and produce a DC component at the output of the mixer. This is called self-mixing. LO leakage to the antenna may result in a time-varying DC offset due to self-mixing. The undesired DC component and DC offset will corrupt the information signal that is present in the baseband. *I/Q* mismatches are other associated problems due to the quadrature down-conversion in homodyne receivers. Because the down-converted signal is located at zero frequency, flicker noise or $1/f$ noise of devices will also corrupt the information signal.

IF Receivers

IF receivers combine the advantage of both superheterodyne and homodyne receivers. The received RF signal is down-converted to IF by an LO, where the IF could be either one or two times the information bandwidth (in low-IF receivers) or arbitrary (in wideband-IF receivers) depending on the system specifications in terms of sensitivity and selectivity.

Low-IF receivers

In the low-IF receiver, a low IF stage is present between RF and baseband, and the low IF signal is directly digitized in the A/D converter, see Figure 1.4. This architecture is also simple and promising for a higher level integration. As compared to homodyne receivers, low-IF receivers have no DC-offset problem since the signal after the first down-conversion is not around DC. The IF is very low (one or two times the information bandwidth), and it is hard to reject the image signal in the RF BPF. Both image and wanted signal will be sampled and quantized in the A/D converter. Further frequency down-conversion from the low IF to baseband is realized in the digital domain to avoid problems such as *I/Q* mismatches in the analog domain

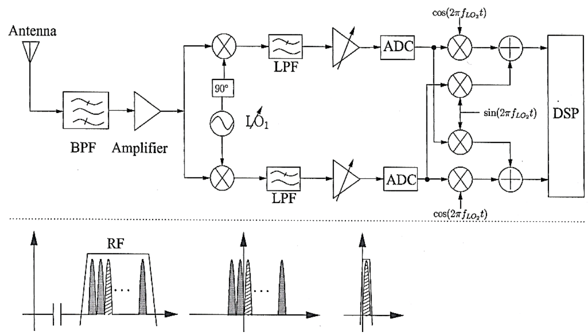

Figure 1.4: Low-IF receiver architecture

II. RECONSTRUCTION

Depending on the context, "reconstruction" has different definitions. Image reconstruction is defined in imaging technology wherein data is gathered through methods such as computerized tomography-scan (CT-scan) and magnetic resonance imaging (MRI), and then reconstructed into viewable images. In analog signal processing, reconstruction mostly means that a CT signal is obtained from the DT data by an interpolation filter or some other filtering processes. Although modern data processing always use a DT version of the original signal, obtained by a certain sampling pattern on a discrete set, reconstruction to a continuous version of sampled data is also needed for some specific applications. In Hi-Fi applications such as digital audio, to maintain high quality in the resulting reconstructed analog signal, a very high quality analog reconstruction filter (postfilter) is required. Reconstruction from one discrete set to another is also useful in the nonfractional sampling rate alternation in digital signal processing. Additionally, in radio receiver front-ends, if the output of the sampling process is not uniformly distributed, a reconstruction process is needed to reconstruct the nonuniform samples to uniform distributed samples prior to the quantizer. According to Shannon's sampling theorem , a band-limited signal can be exactly reconstructed from its samples by US. The perfect reconstruction formula derived by Whittaker for critical uniform sampling is given by

$$x(t) = \sum_{n=-\infty}^{\infty} x(nT_s)\text{sinc}[2B(t - nT_s)], \tag{2.1}$$

where $x(nTs)$ represents samples at the series of equidistant sample instants, $Ts = 1/(2B)$, and sinc(x) = sin(¼x)/(¼x). The reconstruction of the input signal is realized by a convolution summation of uniform distributed samples $x(nTs)$ with a sinc function equivalent to ideal low-pass filtering. In practice, then feeding it into an LPF or other RAs. The reconstruction discussed in the thesis is only realized by a certain RA without any enhancement from the zero-order holding. For NUS, even if there is a large number of samples, only few of them possess a uniform distribution property with respect to the average sampling rate. The expansion of $X(f)$ does not consist of periodic replicas of the fundamental spectrum. Consequently, the signal cannot be determined uniquely by the samples with only

a lowpass filter. Based on the Fourier series expansion, $X(f)$ can be generally expanded as

$X(f) = -j2$¼$fts(n)$, (2.2)

where $ts(n)$ is the set of sampling instants either uniformly or nonuniformly distributed. Using inverse Fourier transform, we obtain the general reconstruction formula:

$$X(f) = \sum_{n=-\infty}^{\infty} c_n e^{-j2\pi f t_s(n)}, \tag{2.2}$$

$$\begin{aligned} x(t) &= \int_{-B}^{B} \left(\sum_{n=-\infty}^{\infty} c_n e^{-j2\pi f t_s(n)} \right) e^{j2\pi f t} df \\ &= 2B \sum_{n=-\infty}^{\infty} c_n \text{sinc}[2B(t - t_s(n))]. \end{aligned} \quad ,(2.3)$$

For US $ts(n) = nTs$, $cn = x(nTs)/2B$. However, for NUS, since $ts(n) = tn$ and cn 6= $x(tn)/2B$ except when $tn = nTs$, the reconstruction formula of eq. (2.21) cannot directly represent the original signal $x(t)$ unless cn is determined. RAs are expected to accurately predict the original signal $x(t)$ from the nonuniform samples $x(tn)$.

In biomedical image processing, CT-scan and MRI frequently use the NUS pattern in the frequency domain. Four sampling patterns are shown in Figure 2.1.

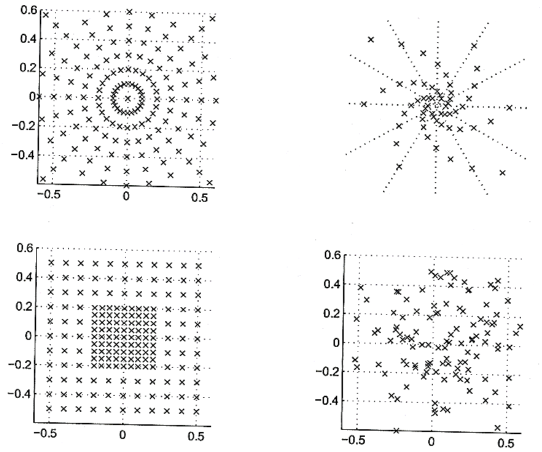

Figure 2.1: Sampling patterns of nonuniform sampling . (*Top-left*): Polar sampling grid; (*Top-right*): Spiral sampling grid; (*Bottom-left*): variable-density nonuniform sampling grid; (*Bottom-right*): general nonuniform sampling grid.

The sampled data of CT-scan and MRI are measured in the Fourier frequency domain. The RA is needed to derive the Cartesian US grid from the acquired data prior to the inverse Fourier transform operation. Inspired by the

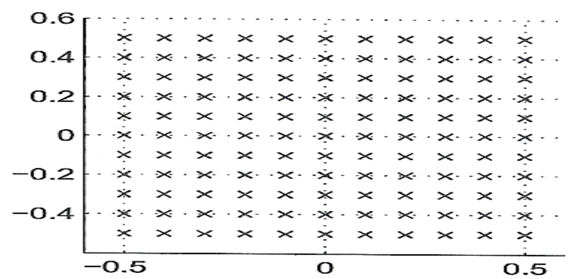

Figure 2.2: Cartesian uniform sampling grid.

applications in biomedical image processing, some RAs extensively used in image reconstructions are proposed for the applications of radio communications. However, the reconstruction process in radio communications is different from that in biomedical image processing. In radio communications, both sampling and reconstruction are in the time domain while they are in the frequency domain in image processing. Additionally, in radio communications, the RA can be used to reconstruct a set of unknown data at a regular time set from the NUS sequence. Then the reconstructed result can be directly fed into the following digital signal processing block (e.g., A/D converter). It is also possible to convert the samples by NUS to a CT signal when an analog signal is needed (e.g., in Hi-Fi) in the processing steam, which is different from the reconstruction in image processing.

The first part of this thesis reviews and compares different radio receiver architectures for conventional and subsampling receivers, single standard and multi standard receivers. The motivation of the thesis is to investigate novel receiver architecture for Software Defined Radio (SDR). Under the concept of SDR, subsampling receivers in BandPass Sampling (BPS) technique becomes more and more attractive. After that, sampling is discussed, including Uniform Sampling (US) and Non Uniform Sampling (NUS), voltage sampling and charge sampling, deterministic sampling and random sampling. A single ideal lowpass filter based on Shannon's sampling theorem is not good enough to reconstruct the signal from the samples by NUS. A general reconstruction formula in terms of a basis-kernel is proposed and nine reconstruction algorithms (RAs) starting from the formula are evaluated and compared especially for deterministic NUS in terms of reconstruction performance and computational complexity. It is investigated that most of these RAs are extensively used in off-line image processing, but algorithms based on *interpolation* are also possibly used in on-line radio communications. Reconstruction becomes hard for random sampling, but random sampling may be helpful for signal identification and eliminating the quantization distortions in the following A/D converters. Then the classic BPS theory is reviewed from the aspects of sampling rate selection, noise aliasing and jitter. The existing studies on BPS are presented and compared. It is noticed that noise aliasing plays an important role in the BPS applications. Starting from the Papoulis' optimised sampling theorem, optimised bandpass sampling including Optimised Quadrature BandPass Sampling (OQBPS) and Optimised Uniform BandPass Sampling (OUBPS) are invented especially for dealing with the performance degradation due to noise aliasing in BPS systems. It is observed that OQBPS and OUBPS perform intrinsic FIR filtering that uses either real or complex filter coefficients. The theoretical analysis and simulation results show that both OQBPS and OUBPS realize sampling, frequency down-conversion and noise aliasing suppression by well-designed intrinsic FIR filtering. Both noise and jitter performance will be theoretically increased by 3 dB when the length of FIR filtering is doubled. However, OQBPS has always limited noise and jitter performance improvement that is determined by the time resolution of sampling. Additionally, the samples by OQBPS are nonuniformly spaced for most cases. OUBPS has better noise and aliasing performance and is easier to be implemented as compared to OQBPS. In the final part, a optimised bandpass sampling receiver based on the concept of OUBPS is implemented at circuit level by Switched- Capacitor (SC) circuit technique. To obtain a better selectivity at the sampling output, an extra IIR filter combined with a decimation operation is introduced. Bandpass voltage sampling and bandpass charge sampling are two promising candidates for subsampling receivers. In this thesis, these two sampling methods are also analyzed and compared in theory. It is shown that optimised bandpass sampling in voltage-mode is more efficient to suppress noise aliasing than bandpass charge sampling with embedded filtering technique. This is determined by the different frequency responses of two sampling techniques. The same noise aliasing suppression may be achieved by lower order FIR filtering in optimised bandpass voltage sampling as compared to bandpass charge sampling. Regarding the problems in real SC circuit implementations, further studies are still needed for the proposed SC circuit architecture of optimised bandpass sampling receiver, e.g. cross-talk among multiple sampling branches, charge injection and clock feed through of switches, parasitic issues in physical level and the trade-off between other performances and power consumption, etc. A final silicon implementation of the proposed SC circuit architecture is expected. Both OQBPS and OUBPS were invented orienting to single-band RF/IF applications. More deep studies on multi-band RF bandpass sampling have come out, and it is of more interest to further investigate OQBPS and OUBPS for multi-band RF applications. It is promising to see multi standard subsampling receivers using the concept of generalize bandpass sampling in the near future.

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