NEW DIGITAL SPECTROMETERS FOR GROUND BASED DECAMETER RADIO ASTRONOMY

P. Kleewein*, C. Rosolen*, and A. Lecacheux*

Abstract

We present a new generation of spectrum analyzers currently being developed at Meudon Observatory (France) in collaboration with the Space Research Institute in Graz (Austria). These devices will perform spectral analysis digitally in real time. The new type of instrument emerged from the need for spectral analysis with at once high dynamic range, high resolution in time and frequency and high sensitivity. A special parallel multiprocessor architecture using the latest available digital signal processors has been concieved and is currently being tested. The final version of the device yields a dynamic range greater than 70 dB and 1024 frequency channels over a bandwidth of 12.5 MHz. Further, the possibility of polarisation measurements will be provided. We discuss design and architecture of the device and present results obtained in laboratory and on telescope with a prototype. Future perspectives will briefly be addressed.

1 Introduction

The instrument under development is intended to be used primarily for studying polarised decameter emissions from Jupiter and the Sun in the 10 to 40 MHz frequency range. The key characteristics of the device have been defined as follows:

- Wide band spectral analysis over an instantaneous bandwidth of greater than 10 MHz.
- Capability of very high time resolution (one millisecond approximately).
- Great number of frequency channels (about 1000).
- Very high sensitivity in order to detect weaker emissions.

Groupe de Recherche Instrumentale en Radioastronomie, CNRS URA 1757, Laboratoire ARPEGES, Observatoire de Meudon, 5 place Jules Janssen, 92195 Meudon Cedex, FRANCE

Additionally, the device was designed to allow for polarisation measurements.

Man–made interferences usually pose serious problems in spectral analysis of decametric emissions; these interference signals can be several orders of magnitude greater than the signal actually to be observed. The spectrum analyzer's linearity is a major issue since any nonlinearity affects the weaker components by means of intermodulation. The new device should be as rugged as possible regarding interference in order to be useable in a "hostile" environment.

Devices currently in use for decametric observations are mainly:

- Swept Frequency Analyzers (SFA): These offer a very high dynamic range but their sensitivity and time resolution is low.
- Acousto-Optic Spectrometers (AOS): Their sensitivity is excellent but they lack of dynamic range.
- Filterbanks: They combine medium dynamic range with good sensitivity but their physical dimensions may be prohibitive if the number of frequency channels gets large.

The design approach for the analyzer we will describe in the following sections was to sample the preprocessed and downconverted RF-signal using an analog-to-digital converter (ADC) and to perform spectral analysis entirely in the digital domain.

2 System Architecture

Figure 1 shows an overview of the device. The system can be divided into four main sections; the first one (on top) contains the analog preprocessor performing amplification, filtering, and up/down conversion of the RF signal. An AD-converter module (below) samples and quantizes the preprocessed signal, while several digital processing modules, each containing a cluster of digital signal processors (DSP), perform spectral analysis in real time. Finally, a workstation (bottom) serves as host.

The analog preprocessor comprises three sections: radio frequency (RF)-, intermediate frequency (IF)-, and baseband section. After upconverting the RF-signal to the IF (centered at 70 MHz) in the first mixer stage it is passed through a bandpass-filter of 10 MHz bandwidth. The second mixer downconverts the signal to baseband, that is, the 10 MHz IF-passband is transferred to the frequency range between zero and the Nyquist frequency (half the sampling frequency) of the ADC. The 10 MHz band to be processed can be placed arbitrarily between RF-frequencies from 5 to 50 MHz by means of the first mixer stage which contains a tunable local oscillator (LO 1). This LO is driven from a direct digital synthesizer. The ADC has a resolution of 12 bits and runs at a sampling frequency of 25 MHz.

The digital sections of the device are built around a VME-bus. This industry-standard bus system is used for data transfers between the host and the DSP-processors and also provides the power supply for the ADC-board. However, data from the ADCs is transferred

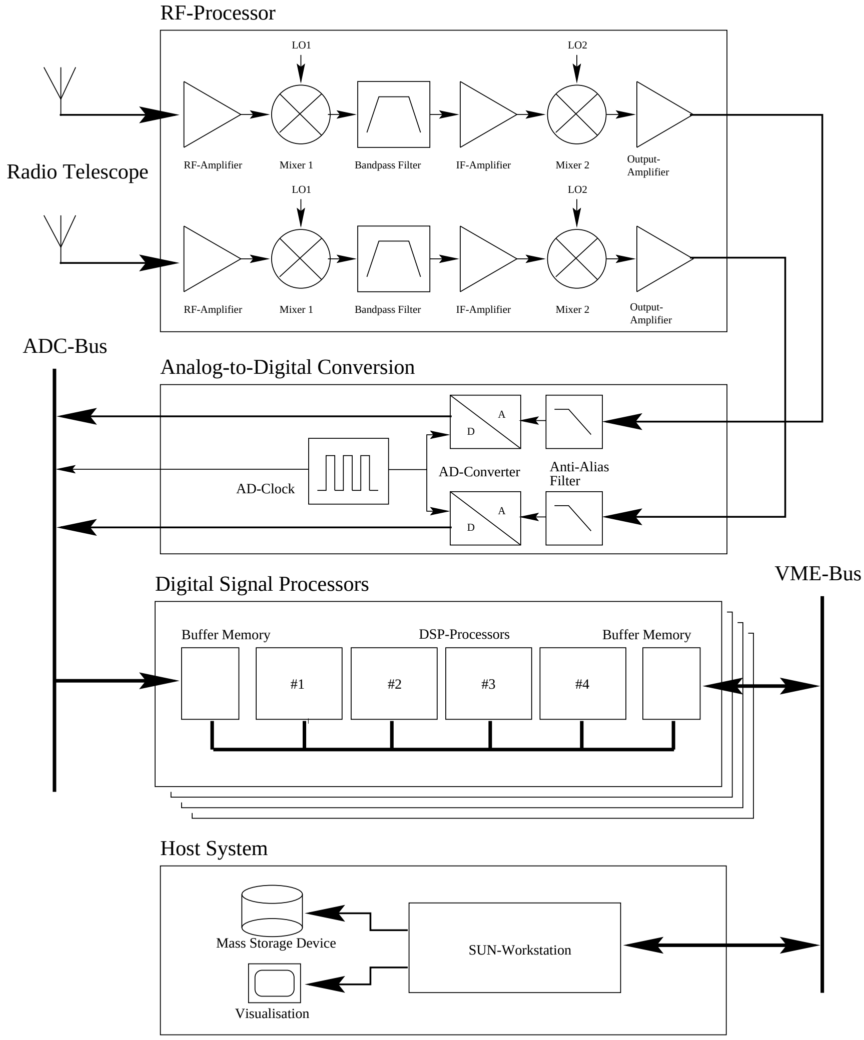

Figure 1: System Overview

via a special bus (ADC-Bus in the figure) to the DSP-cards in order to avoid saturation on the VME. As shown in the figure, the device is equipped with two ADCs as well as a two-channel analog preprocessor. Furthermore, data from both ADCs is accessible on each DSP-unit. This feature allows the computation of correlation between signals acquired from two different antennas and has been provided for polarisation measurements.

3 Power Spectrum Estimation

A method for estimating spectral power using the fast Fourier transform (FFT) has been presented by P. D. Welch [1967]. This method is computationally efficient and requires a relatively small amount of memory to be available on the hardware platform. It involves sectioning the sample record, computing the modified periodogram on each section, and averaging these periodograms. The so-called modified periodogram is the squared magnitude of the Fourier transformed computed from the windowed data section. The process can be outlined as follows:

- Windowing of data vectors:
$$w(t) \cdot x_k(t),$$
where $x_k(t)$ denotes the data vector in the time domain, and $w(t)$ the weighting function.

- Computation of FFT on each vector:
$$W(f) * X_k(f) = FFT\{w(t) \cdot x_k(t)\},$$
where $X_k(f)$ denotes the Fourier transformed of $x_k(t)$, $W(f)$ the Fourier transformed of $w(t)$, and the star ($*$) the convolution operator.

- Computation of periodogram $P_k(f)$:
$$P_k(f) = [W(f) * X_k(f)] \cdot [W(f) * X_k(f)]^*,$$
where ($[]^*$) denotes the complex conjugate of ($[]$).

- Average over K individual periodograms:
$$P(f) = \langle P_k(f) \rangle.$$

For polarisation measurements, the cross spectrum can be calculated from the Fourier transformed of the two associated antenna signals:

$$P_{cross}(f) = \langle [W(f) * X_{1k}(f)] \cdot [W(f) * X_{2k}(f)]^* \rangle.$$

From this complex spectrum and the (real) power spectra of the two antenna signals, the Stokes parameters can be calculated.

Assuming $x_k(t)$ to be a sample from a stationary random process and $P(f)$ to be flat over the passband, the variance of the estimate $P(f)$ is inversely proportional to the number of individual periodograms K:

$$\frac{var\{P(f)\}}{var\{P_k(f)\}} = \frac{1}{K}.$$

Furthermore, it can be shown that the expected value of the periodogram is the true power spectrum convolved with the magnitude squared Fourier transform of the window [DeFatta et al., 1984]. The estimate is asymptotically unbiased except for a normalizing factor which can be calculated from the window function.

The number K cannot be arbitrarily increased since a certain time resolution T is required. Rather, $P(f)$ has to be computed using only the samples which arrive within the time interval T. On the other hand, if we want to obtain N frequency channels, we need $2N$ samples from which an individual periodogram $P_k(f)$ can be computed. Straightforwardly, the number of individual periodograms K can be calculated from:

$$K = \frac{f_s T}{2N},$$

where f_s is the sampling frequency of the AD-converter. In this case, the series of samples within the time interval T are sectioned into K *non-overlapping* data vectors of a length of $2N$ samples. There are two reasons for using *overlapped* data vectors to compute the final estimate $P(f)$. First, short events may be missed in a non-overlapped analysis due to the weighting function's fall-off near its boundaries. Second, a considerable loss of sensitivity with respect to a device with "ideal sensitivity" like an AOS occurs. To enhance performance, overlapped sections can be used, which in turn require additional computation power. A reasonable value for the number of overlapped periodograms K_{ov} can be obtained from:

$$K_{ov} = k \cdot \frac{f_s T}{2N},$$

where k is chosen between 2 and 4, yielding *fractional overlaps* between 50 and 75%. At 75% overlap, the variance of the estimate $P(f)$ is only slightly higher than the one at ≈100% for all usual weighting functions [Harris, 1978]: The loss in sensitivity is very well below one dB in this case, since consecutive data sections are already strongly correlated with each other. The value for k must be determined in a tradeoff-analysis considering available computation power and desired performance. Obviously, the shape of the weighting function is of importance.

4 Technical Considerations

Although published in 1967, Welch's method is still uncommon in the field of radio astronomy. This may be due to problems concerning analog-to-digital conversion on the one hand and computation power requirements on the other. According to theory, a continuous signal of a bandwidth B must be sampled at least at a frequency

$$f_s \geq 2B$$

to be represented unambiguously in the discrete-time domain, while the signal-to-noise ratio (SNR) of an AD-converter can be estimated [Oppenheim et al., 1975] from its number of bits b :

$$SNR[dB] \approx 6 \cdot b.$$

Thus, high dynamic range and wide bandwidth require a fast ADC with high resolution; for example, a 12-bit-ADC running at 20 MHz theoretically yields a bandwidth of 10 MHz and a signal-to-noise ratio of 72 dB. Such converters became available only a short time ago.

Numerous fast Fourier transform algorithms have been published since 1965 [Cooley et al., 1965]. In general, these require a number of operations (OPS, meaning multiplications and additions in this context) which can roughly be calculated

$$OPS_{FFT} = k_{FFT} \cdot N log_2 N,$$

while the direct computation of the discrete Fourier transform requires

$$OPS_{DFT} = 8N^2 - 2N$$

operations. N is the number of frequency bins and k_{FFT} usually lies between 4 and 5 (depending on the algorithm). Using the formula for K_{ov} from the previous chapter, assuming a sampling frequency of 25 MHz, a fractional overlap of 50%, and 1024 frequency channels, we obtain a computational load of more than one billion operations per second for a real-time spectral analysis; this number does not include any computation other than the FFT, such as windowing and averaging.

We investigated two kinds of processors which could be used for real-time spectral analysis: specialized FFT-processors on the one hand, and more "conventional" programmable DSP-processors on the other. We adopted a solution using only the latter type mainly because of its greater fexibility, although hardwired chips typically execute the FFT several times faster than programmable ones. Data compression, techniques for handling man-made interferences, automatic signal recognition or polarisation measurements can only be implemented on a freely programmable device. Another consideration concerned the limited dynamic range of many specialized processors.

The high amount of computation power implies the use of a multiprocessor architecture. Further, using the method of Welch, the process of spectral analysis can be easily decomposed into several equal tasks which can be executed in parallel. The architecture we finally designed was a *homogeneous* multiprocessing system with a *symmetrical* implementation of Welch's method. As processor we chose the ADSP-21060 ("SHARC") from Analog Devices, a DSP-processor with high performance in FFT-computations and well-suited for multiprocessing applications. About twenty of them, each providing a computation power of about 100 MFLOPS (million floating point operations per second), are needed for a machine with 12.5 MHz bandwidth, 1024 frequency channels, and good sensitivity (50% fractional overlap).

5 Results Obtained with a Prototype

The multiprocessor system described above is currently under development. Nevertheless, a prototype of the spectrum analyzer has already been tested both in laboratory and on antenna. It comprises a one-channel analog reception unit, the ADC-converter board, and the host system as described in section 2, but contains only a computation unit with a single DSP-processor. This machine does not provide good sensitivity since it only computes one periodogram per millisecond instead of the desired 25, thus having a low time coverage of the input signal, but other key characteristics such as dynamic range or frequency resolution are not affected by the lack of processing power.

Figure 2 shows the system's response to a monochromatic signal applied to its input. It can be seen that several spurs occur, which are at least about 75 dB weaker than the test signal. The test signal was just below the saturation of the AD-converter (0 dB in the figure). The response to broadband noise can be seen from Figure 3.

Response to broadband noise has also been measured; the power of the noise source has been varied over a range of 64 dB in steps of one dB. The power at the system's output has been traced against the injected noise power, which provides a measure of linearity regarding broadband sources. The response is linear over a range of about 50 dB.

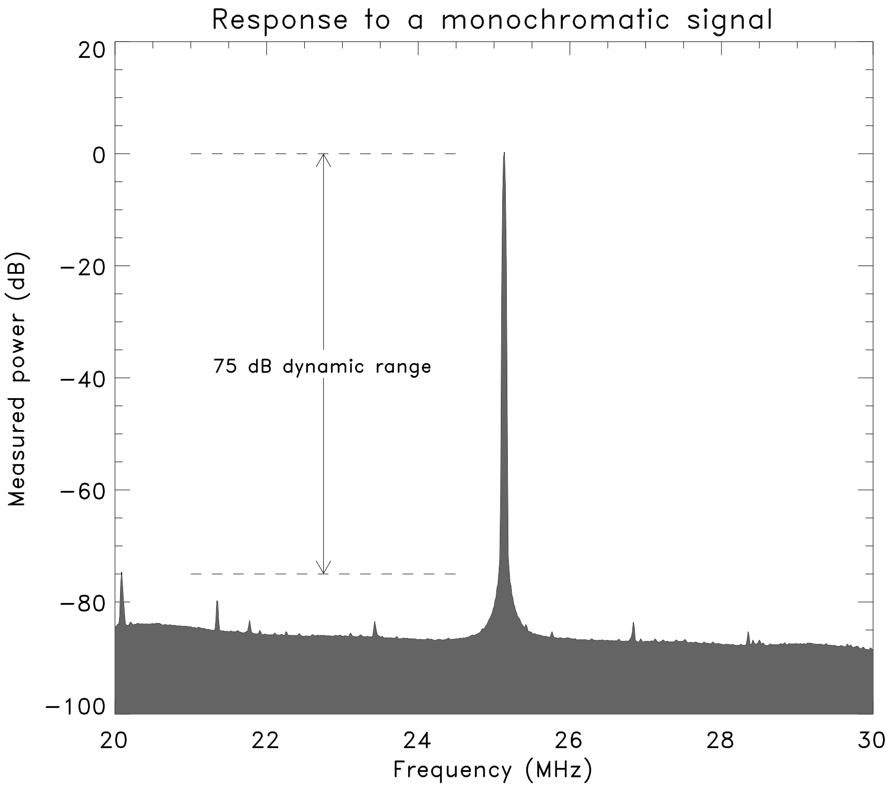

Figure 2: Monochromatic Signal

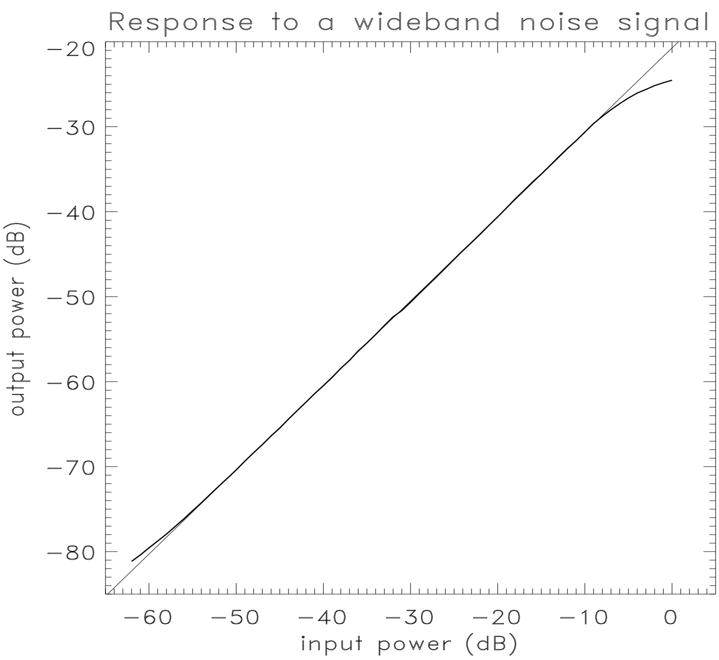

Figure 3: Broadband Noise

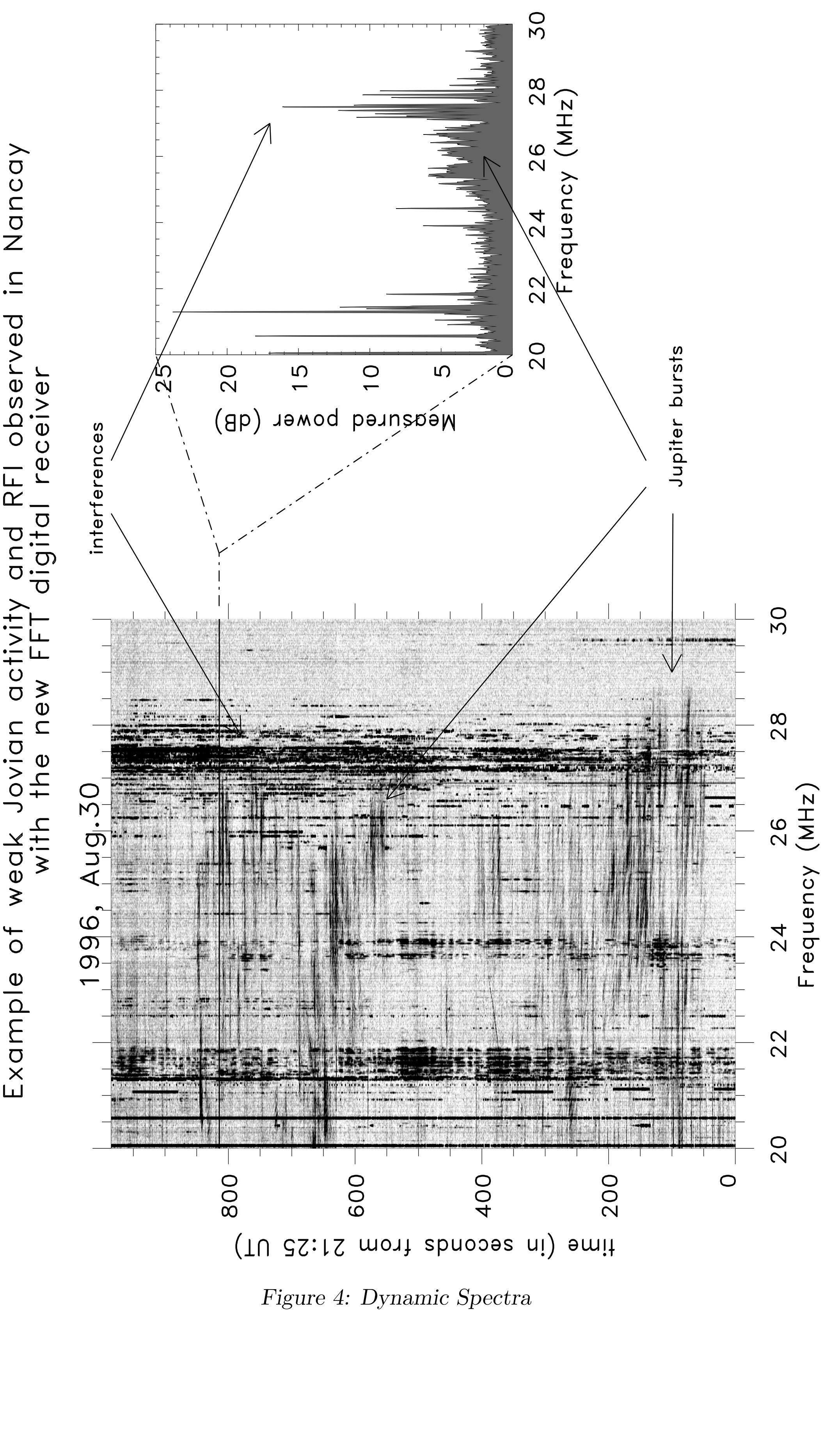

Figure 4: Dynamic Spectra

Finally, Figure 4 shows dynamic spectra acquired with the prototype at the decameter array in Nançay, France, at a time resolution of one second.

The laboratory measurements have been carried out without external filters at the device's input; only the internal filters (IF-filters, anti-alias filter at the ADC-input) have been used. A 4-term Blackman-Harris window [Harris, 1978] with highest sidelobes at -92 dB served as weighting function for all tests.

6 Conclusion

We summarize the results of our investigations and tests.

- The digital device is rugged towards interferences. Its response to broadband noise is linear over 50 dB, a spurious-free dynamic range of greater than 70 dB has been measured.
- Frequency resolution and channel separation is excellent due to filtering in the digital domain.
- A time resolution down to the theoretical limit of $\frac{1}{B_{channel}}$ could be obtained. This is not a matter of computation power but of data transfer capabilities and/or memory size.
- Polarisation measurements, data compression, automatic signal recognition and techniques for handling interferences could be implemented.

We will continue our work on this subject especially towards a system with both high sensitivity and high time resolution providing polarisation measurements.