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Digital Ultra-Wideband-Sensor Electronics integrated in SiGe-Technology

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Ultra-Wideband (UWB) sensors are of great interest for a vast number of different applications. Due to their bandwidth they provide a great deal of information about the objects under test compared to classical narrow band sensors. This improves the reliability of the measurement results and opens up new fields of application for electrical sensors. New UWB-electronics are introduced which provide the opportunity for large volume application of such sensors. Its bandwidth ranges from nearly zero to 5 GHz. The RF interface comprises of integrated digital circuits in SiGe technology. A Maximum Length Binary Sequence (MLBS) is used as transmit signal. At the receiver, fast track-and-hold and interleaved sampling is applied for data recording. The received signal is processed by a DSP. Excellent timing stability, low power consumption and miniature size support synchronous multi-channel operation which is a prerequisite to build sensor arrays.

INTRODUCTION

Ultra wideband (UWB) radar and impedance spectroscopy are of great interest for a vast number of applications such as surface penetrating radar, surveillance and emergency radar, medical instrumentation, non-destructive testing in civil engineering and the food industry, industrial sensors and microwave imaging and many others. The fractional bandwidth of the sounding waves for such types of applications should be as close as possible to the maximum value of 200 % resulting in a high spatial resolution and good penetration in materials. An UWB radar is able to detect hidden objects and a high bandwidth not only results in good spatial resolution but also in improved capabilities for object recognition. Concerning the impedance spectroscopy, a large bandwidth covers different relaxation phenomenon of matter so that more information is available for the substance characterisation.

An UWB-sensor is built from three main parts. These are the applicators, the measurement electronics and the processing software. The applicators are specific electrode configurations adapted to the actual measurement scenario. They transform between the sounding waves and the measurable electrical signals. Typical applicators are impulse radiating antennas, open coax-lines, wave guides filled with material under test etc. The task of the measurement electronics is to provide the stimulation signals and to gather the reaction of the scenario under test (SUT) referring to that stimulation. The processing software is finally responsible to derive the wanted information about the SUT from the measured data. Usually, this is connected to an inverse problem. The confidence in its solution is dependent on the quantity of information which is captured from the SUT. This underlines the importance of UWB-measurements because they provide as much information as possible.

Generally, it can be supposed that the SUT behaves linear with respect to its stimulus and its state does not change during the measurement time. From this view point of system theory, the measurement undertaken by an UWB sensor results in a frequency response function (FRF) or an impulse response function (IRF) describing physically the reflection or transmission behaviour of the sensor (applicator) arrangement respectively its impedance or similar (see Figure 1).

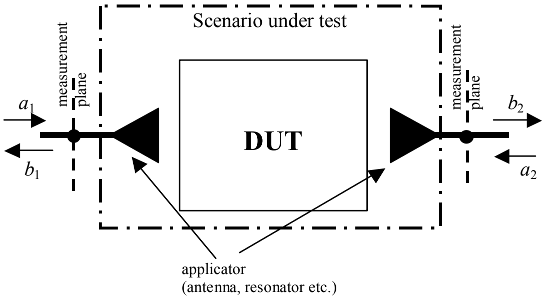

***Figure 1**: Generalised sensor arrangement. Note that the actual captured data refer to the measurement plane. This means they include the behaviour of the device under tests as well as the behaviour of the applicators.*

Finally, it is to note that the IRF and the FRF form the data base from which the wanted information about the SUT will be derived. This procedure strongly depends upon the actual kind of application and will not be considered here in detail. There are several methods known of how to measure the FRF or IRF over a wide spectral band, these are particularly impulse or sine wave techniques. Their usability for high volume sensor products is however very limited.

The type of applicator and the choice of an appropriate data processing strongly depends upon the actual application. It will not be possible for example to use exactly the same applicator configurations and processing software for mine detection by a high resolution radar and for impedance spectroscopy of the skin for cancer detection or for liquid sensing in a brewery. This must however not be the case for the electronics providing they are scalable to different basic requirements such as bandwidth, stability, measurement speed, costs etc. and the applicators are equipped with standardises RF-ports. Scalable, universal UWB-electronics and a toolkit of corresponding components is a prerequisite for large scale applications of UWB-sensors.

The article will refer to a new UWB-conception tested for the first time in a mine detection radar array. Its applications is however as already suggested not restricted to such an applications The principle profits from the use of UWB pseudo random binary sequences resulting in a simple and flexible system structure with scalable performance values. Its flexibility and scalability is based on the consequent use of digital basic circuits also within the RF-part, and the high degree of software. The possible scope of its technical implementation – and by that the system costs - covers opportunities to use PCB- and Hybrid-technique for low and medium quantities as well as a complete integration for high volume quantities.

In what follows, the basic principle will be shortly introduced, the key figures will be given and some results with a laboratory sample are shown.

THE PHILOSOPHHY

The key for an efficient system configuration and its technical implementation is the appropriate choice of test signal. Referring to UWB measurements, the theory of linear and time invariant systems does not suppose a specific shape of stimulus rather it only requires that the signal has a very short auto correlation function (ACF) i.e. a large bandwidth. On this basis, one is not exclusively tied to pulses or sine-waves which increases the degree of freedom in the system conception. Now, aspects of the technical implementation may be moved into the foreground of interest. Those cover mainly three points:

Avoid high signal peaks: The signal to noise ratio of the measurement data strongly depends upon the quantity of energy which is available to stimulate the scenario under test and which can be gathered by the receiver. However, the maximum signal level of the RF-electronics is quite limited. Thus, the instant signal power must be distributed equally over time in order to maximise the total energy. The crest factor CF is a quality parameter which characterises the homogeneity of energy distribution. It is given by the ratio between maximum and effective values of the signal:

$$CF = \frac{\hat{X}}{X_{eff}} = \sqrt{\frac{P_{peak}}{P_{av}}} \qquad (1)$$

It is seen from (1) that a signal with a crest factor close to one provides more energy even at a relative low signal level than a high crest factor signal such as an impulse. Low level signals do not burden the electronics extremely which provides for stable and fast (high bandwidth) operation. Furthermore, they may be handled in integrated RF-circuits promoting a further improvement in bandwidth.

Do not measure faster than necessary: By under-sampling, the speed of data gathering can be reduced to a value that is determined by the time variance of the test scenario (for example the maximum speed of a target) rather than the bandwidth of the sounding waves. This greatly simplifies the electronics and reduces costs. Under-sampling may be applied in connection with any periodic stimulus signals. However, attention must be paid to provide an efficient and stable control of the sampling instant.

Perform digital signal processing: Digital signal processing provides a high degree of freedom and flexibility. Only the software fixes whether the sensor is dealing with time domain data (in case of a radar) or if it is dealing with frequency domain data (in case of an impedance spectrometer). Depending upon the required measurement result - IRF or FRF - two different processing strategies have to apply. This does not however have influence on the hardware structure of the measurement head.

The IRF-measurement for radar applications for example requires an impulse compression which may be performed by matched filtering or correlation. Both can in principle be undertaken by analogue or digital means. In the case of UWB applications, analogue compression methods are usually not applied since it is very difficult to build such systems with a large fractional bandwidth and a high compression factor. A high compression factor requires the storing of a huge amount of energy distributed over a wide spectrum within the impulse compression systems. Herewith, analogue solutions quickly meet with their technical limits. Moreover, the

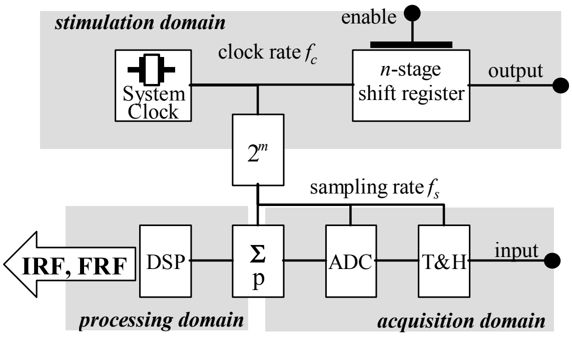

Figure 2: *Basic block schema of the UWB front-end*

output signal of an analogue impulsion compression is again a high crest factor signal which should be avoided in analogue circuits (see above). The digital impulse compression is not sensible to these effects. The "energy storage" is a question of memory size and a high crest factor signal supposes an appropriate word length respectively number format. Both points, may be respected by a corresponding design. They are subjected less critical technical constraints than for analogue system solutions.

Replacing the impulse compression by a FFT algorithm provides results in the frequency domain i.e. the measurement of the FRF and the use of parameter estimation techniques even supply the results in a compressed form of model parameters.

THE MLBS PRINCIPLE

Referring to the above mentioned points different solutions are theoretically possible. But a good choice for a test signal under these circumstances is the Maximum Length Binary Sequence (MLBS or M-sequence) (see Alrutz (2)) which mainly profits from following aspects:

The M-sequence may be generated easily by a digital shift register with appropriate feedback loops even at clock rates in the GHz domain. A shift register of length n provides a sequence of 2^n-1 chips. Due to its periodicity the M-sequence permits under-sampling and synchronous averaging. The MLBS has a sinc^2-spectrum whose zeros are identical with the clock rate f_c of the shift register. It is a wideband signal having a short ACF which is free of side lobes.

There is a very simple and stable way to control the sampling instants for the data gathering: If one is satisfied with an equivalent sampling rate of f_c (i.e. one sample per chip), a simple binary divider is able to provide a stable sampling clock. If m is the number of divider stages, 2^m cycles of the M-sequence are needed to gather a complete data set, since there is always a lack of one between the period of the M-sequence and the divider output. The actual data gathering (sampling) rate is given by $f_s = f_c/2^m$. Due to the Nyquist sampling theorem the usable bandwidth is limited from zero to half the clock rate with this kind of sampling. But this lower part of the spectrum contains the majority (more than 70 %) of signal energy. Thus, an increased equivalent sampling rate will only scarcely improve the usable frequency band because the immunity against noise will rapidly decrease by going beyond the $f_c/2$ value.

Synchronous averaging improve the noise performance by a factor of $\sqrt{p}$. Herein p is the number of averages. Averaging uses the lack between the (mostly low) data rate required by a slowly varying SUT and the high speed capability of modern digital electronics.

These three points determine the basic structure of the measurement head which is shown in figure 2. A RF-single-tone clock generator pushes in parallel the shift register and the binary clock divider. The first one generates the stimulus signal and the second one controls the data recording. This structure guarantees a very stable synchronism between signal generation and signal capturing. This stability is to be seen in a twofold sense. First, the steep flanks of all control signals suppress jitter and drift tendencies. Thus, the high frequency parts of the measurement signal will not be smoothed out by averaging. Second, the sampling time instants are absolutely linear distributed over time. Nonlinear time axis distortions as in ramp controlled sequential sampling oscilloscopes do not appear. An important fact for a sophisticated data processing.

The input signals are sampled by a T&H or S&H circuit before they are transformed into the digital domain. The signals are then synchronously averaged to increase the dynamic range and to reduce the data rate. Averaging is often useful because the sampled data are usually very noisy because of their large bandwidth.

Finally, the gathered data must be processed in an appropriate way. In the simplest case the FFT-algorithm is used in order to provide the FRF of the sensor respectively the FHT (fast Hadamard transform) provides its IRF. The FHT algorithm is organised in a very close manner as the FFT algorithm but it only requires additions thus it may be implemented in a very fast way. It should also be noted that no high crest factor signals are loading the analogue circuits. High signal peaks only appear after an impulse compression in the digital domain.

As identified in figure 2, the measurement head may be divided into three parts of different speed respectively rate levels. The clock rate in the stimulation domain is determined by the physics of wave propagation. Usually it must be beyond several GHz. The rate in the processing domain covers the range of one IRF/FRF per hour up to thousand of functions per second depending on the application. The rate in the acquisition domain must be in between. The higher it is the better the measurement dynamic but also the higher the system costs.

The key figures of the basic system may be summarised by:

Bandwidth $$B = f_c / 2 \quad (2)$$

Time window length: $$T = \left(2^n - 1\right) / f_c \quad (3)$$

Number of spectral lines: $$N \approx 2^{n-1} \quad (4)$$

Measurement time: $$T_{obs} \approx p\, 2^{n+m} / f_c \quad (5)$$

Dynamic range (IRF):

$$L_{n1}[dB] \approx 6\, ENOB + 3n + 10 \log_{10} p \quad (6)$$

Dynamic range for an individual spectral line:

$$L_{n2}[dB] \approx 6\, ENOB + 10 \log_{10} p \quad (7)$$

Herein the dynamic range of the whole receiver track is summarised by its effective number of bits (ENOB). The jitter performance is fixed by the phase noise of the RF-master clock. Measurements have shown that the other circuit parts do not contribute to the jitter. This means that the self jitter of the critical RF-components as shift register, binary divider and T&H were less than

140 fs. The requirements on the stability parameters of the RF-generator is finally dictated by the application. Large delay times within the SUT suppose a low phase noise and an excessive averaging is limited by its medium term drift.

Considering figure 2 it does not provide any difficulties to expand the principle to a multi-channel measurement system as required in UWB arrays. As a matter of principle any number of measurement channels may be synchronously driven by the binary divider. The same is for the shift registers so that several stimulation channels are available. There is no need for RF-switches.

For more details on the function principle and system modifications see Sachs et al (2) and (3).

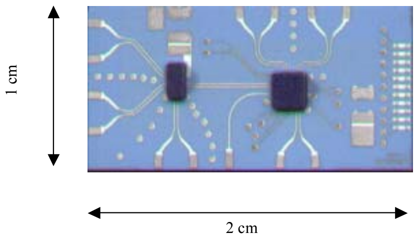

***Figure 3**: RF-part showing two SiGe chips mounted onto a 4 layer LTCC.*

EXPERIMENTAL RESULTS

A first integrated version of the UWB front-end was built for a GPR array with 6 transmitter and 6 receiver antennas for anti-personnel landmine detection purposes (EU-Project: ESPRIT 29902 DEMINE). The shift register, the binary divider and the T&H-circuit were integrated in 0.8 µm SiGe-HBT-technology (see Roßberg et al (4)) . The complete analogue part of the device was designed by differential circuits. The wiring of the RF components was performed by four layer low temperature co-fired ceramics (LTCC). Figure 3 shows such a ceramic board with the RF-electronics.

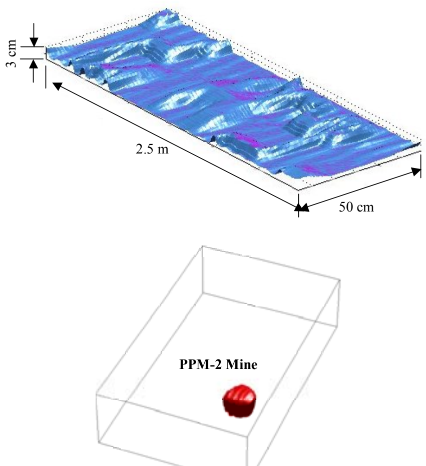

***Figure 4**: Surface profile of a demining lane (above) and iso-surface of the wave energy scattered by a mine (with courtesy of Vrije Universiteit Brussel).*

A Video-ADC captures the sampled data which were either stored directly on a PCI-board or they passed through a fast averaging circuit (FPGA) and the impulse compression (FHT-algorithm running on a DSP) to the main processor. The shift register length was $n = 9$ resulting in a 511 points IRF. The RF-clock rate was $f_c = 9$ GHz i.e. the useful bandwidth was approximately 4.5 GHz. The maximum measurement rate was approximately 30 000 IRF/s if only data storage and off-line processing was applied.

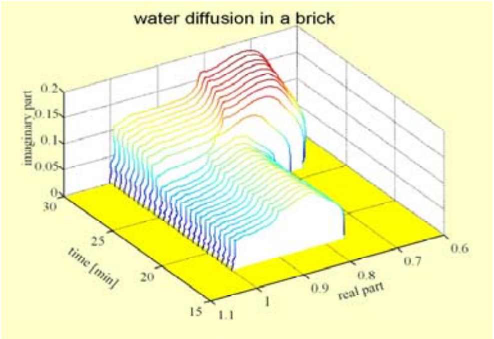

***Figure 5**: Complex reflection coefficient depending upon the time exposed to water.*

The figures 4 and 5 indicate two arbitrary measurement examples. The first one refers to a radar application for mine detection. It is showing the surface profile of the deminig lane and a buried object. Figure 5 demonstrates an example of impedance spectroscopy. It shows the complex reflection coefficient of an open coaxial line in contact with a brick which was exposed to water diffusion (see also Sachs et al (5)).

CONCLUSIONS

The functioning principles of an UWB sensor electronic were explained and some experimental results were shown. The system concept follows from a consequent exploitation of the advantages of M-sequences. It results in flexible design and stable operation. It forms the basis for a largely integrated system. Thus, the prerequisite is met for large scale applications of UWB-sensors for radar purposes as well as for impedance spectroscopy.

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