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INTEGRATED DIGITAL UWB-RADAR

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1. 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 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.

This article will be concentrated on UWB radar rather than on problems of impedance spectroscopy even if both measurement tasks are closely related. The UWB radar measures the time variant impulse response function (IRF) of the scenario of investigation which is determined by the arrangement and movement of scatterer and the illumination of the scene by the antennas. The classical methods of determining a systems impulse response function is based on pulse excitation methods as the function's name implies. Other methods deal with the determination of the frequency response function (FRF) – a Fourier transformed version of the IRF - over a very large bandwidth. These methods typically use sine-waves which are stepped or swept through the band of interest or a Multi-Sine concept is applied. However all these methods are burdened by critical disadvantages concerning their technical parameters or/and the price (or/and size) of the equipment. In the opinion of the authors, this is the main explanation for the gap between

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the current status in the practical use of UWB sensor technique in industry and its actual physical potential.

The article will present a UWB-measurement concept which will reduce the above mentioned gap. The principle was tested for the first time in an anti-personnel landmine (APL) detection radar. The radar profits from the use of UWB pseudo random binary sequences which may be generated up to a very large bandwidth and which have high energy even at low signal levels. This favours the noise suppression as well as the system integration on low cost RF-circuits using SiGe-technology. Furthermore, nearly the complete system conception is based on digital circuit schemes promoting high flexibility and cost efficient implementation for applications. In what follows, the basic principle will be introduced and the main design rules will be described. In contrast to simple pulse systems, the presented principle is robust against drift appearances as sine-wave systems. Thus, the reduction of systematic errors by calibration procedures is an interesting feature to improve the performance. Two basic methods will be mentioned. Some results from the first practical implementation will close the article.

2. THE BASIC PRINCIPLE

The key for an efficient system configuration as well as its technical implementation is the appropriate choice of the test signal. Referring to UWB measurements, the theory of linear 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.

Avoid high signal peaks within the analogue part of the electronics: Low level signals do not burden the electronics extremely which provides for stable and fast (high bandwidth) operation. Furthermore, low level signals may be constructed in integrated RF-circuits promoting a further improvement in bandwidth. The signal must however also contain as much energy as possible in order to provide good suppression of perturbations. Thus, the instant signal power must be distributed equally over time i.e. the crest factor should be close to one. However, the reaction of a device under test on such a signal does not correspond to the IRF. Therefore, the reaction of the device must be compressed in the receiver in order to provide the IRF.

Use under-sampling: 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.

Use digital impulse compression: Impulse compression may be performed by matched filtering or correlation. Most principles use analogue techniques which is the usual method to improve the signal to noise ratio. This method is however only rarely applied in UWB systems since it is very difficult to build analogue 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 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 underlying less critical technical constraints as for analogue system solutions.

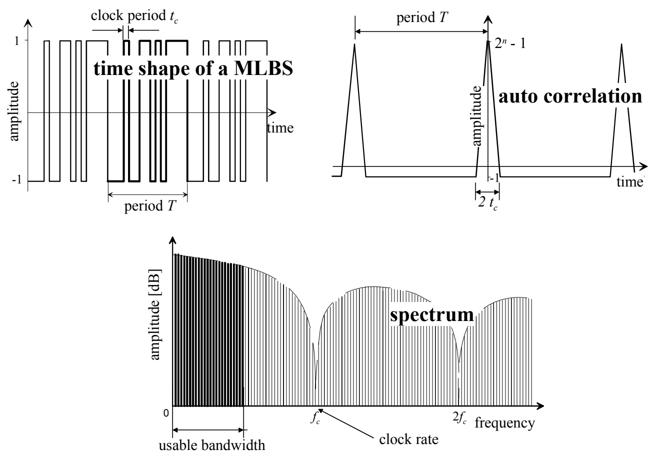

Figure 1: Time shape, auto correlation function and spectrum of a MLBS. For a better graphical representation the shown spectrum applies to a MLBS of higher order than used in the upper diagrams.

A good choice for a test signal under these circumstances is the Maximum Length Binary Sequence (MLBS or M-sequence). It is a periodic wideband signal having a short ACF which is free of side lobes (in the ideal case). Figure 1, shows an example of its idealised time shape, spectrum and ACF. The MLBS stimulus may be generated easily by a digital shift register with appropriate feedback loops even at clock rates in the GHz domain. It is a periodic signal permitting under-sampling and synchronous averaging for noise suppression. Interestingly, 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. This sampling principle exploits the fact that the number of chips in a MLBS is always one less than a power of two. Due to the Nyquist sampling theorem the usable bandwidth is limited to half the clock rate (see figure 1) with this kind of sampling. However the 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. Finally, there are simple and fast algorithms to compress the MLBS response of a device under test (DUT) in order to receive its IRF.

The basic structure of the MLBS front-end which is aimed to measure the IRF of a DUT can be derived in a straight forward way as mentioned above, figure 2 represents the result. A RF single-tone clock generator pushes in parallel the shift register which generates the stimulus signal and the binary clock divider, which controls the data gathering. This guarantees a very stable synchronism between signal generation and signal capturing. 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. Finally, the impulse compression is performed either by a matched (FIR) filter or by the Fast Hadamard Transform (FHT). The FHT algorithm is organised in a very close manner as the FFT algorithm but it only requires additions thus it can 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 the FHT respectively the FIR-filter.

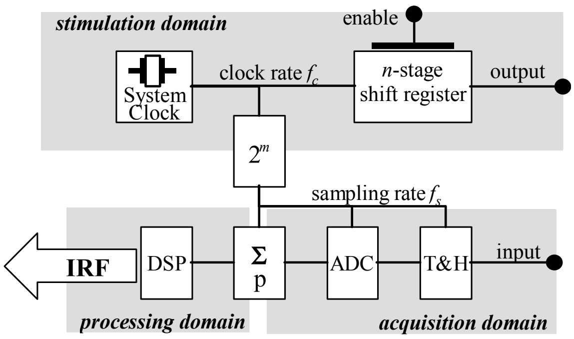

Figure 2: Basic block schema of the UWB front-end

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. However, usually only one signal generator should be active at a certain time. This is controlled by enabling/disabling the individual shift registers[1]. There is no need for RF-switches.

3. THE KEY FIGURES

Most of the technical parameters of the MLBS front-end are determined respectively influenced by easily controllable values. This opens up the opportunity to adapt the system performance by simple means to the requirements of different applications. The concerning quantities are the number n of shift register stages, the number m of binary divider stages, the number of synchronous averages p and the RF-clock rate f_c. In what follows, a short overview about the most important technical parameters and their relations to the design parameters shall be given.

Bandwidth, range resolution: The usable bandwidth B and by that the range resolution δ_r in case of radar applications are fixed by the Nyquist sampling theorem.

Supposing an equivalent sampling rate of f_c as in our case they result in (c – propagation speed of the waves):

$$B = \frac{1}{2} f_c \qquad\qquad \delta_r = \frac{c}{f_c}$$

It should be noted that for most applications which require a certain digital post-processing one has to see that the spectral components of the received signal are really limited to the range between DC and $f_c/2$. If the DUT by itself does not limit the upper frequency parts, an anti-aliasing filter has to force it.

Time window, ambiguity range: The length of the IRF T i.e. their number of points depends upon the period length of the MLBS and this is given by the shift register length. Since the measurement results are periodic in T, the radar ambiguity range R corresponds to that period.

$$T = \frac{2^n - 1}{f_c} \qquad\qquad R \approx \frac{2^{n-1}\, c}{f_c}$$

Measurement time/rate, maximum target velocity: The measurement time T_{obs} is the time over which the object is “observed” to get one complete IRF. It should be noted that the measurement principle supposes time invariance of the DUT during the data gathering . From this it follows:

$$T_{obs} \approx \frac{p\, 2^{n+m}}{f_c} \qquad\qquad 2\, T_{obs}\, v_{\max} \le \delta_r$$

If there is no lag in data processing or data transport, the measurement rate r_m is the inverse of T_{obs}.

Signal to Noise Ratio (SNR): The SNR-value is given by the quotient between the maximum possible value of the IRF referred to the RMS noise level. Supposing the dynamic of the analogue input part (amplifier, T&H, ADC) is expressed by ENOB (effective number of bits), the SNR-value L_n [dB] can be approximated by:

$$L_n\left[dB\right] \approx 6\, ENOB + 3\, n + 10 \log_{10} p$$

Peak to side lobe ratio: The ideal MLBS has an infinity peak to side lobe ratio but by the required band limitation because of the limited equivalent sampling rate side lobes appear in the system IRF. Then, the actual size of the side lobes mainly depends upon the anti-aliasing filter which the system designer can largely influence. By increasing the equivalent sampling rate, the need of a band limitation will disappear. Thus the resulting side lobes will increasingly vanish.

Peak to spurious ratio: The peak to spurious ratio expresses the influence of non-linear effects onto the IRF. In the case of the MLBS front-end such effects may be observed mainly in the shift register which generates the stimulus signal. Thus, the actual created test signal differs from the supposed one. This is expressed in small peaks which are distributed in a defined manner over the basis of the system IRF[2]. By an appropriate signal correction these kind of perturbations may be mostly removed (see below).

Jitter, Drift: Most of impulse systems applying under-sampling are sensible to drift and jitter. This is not the case here. The complete control of the measurement process is done by steep signal flanks which suppress the effect of additive noise and temperature dependent offset voltages on the trigger events. Thus, the system is as stable as a network analyser. An influence of the shift register, binary divider and the T&H onto the jitter behaviour could not be observed i.e. the measurements indicated that their effect must be less than 140 fs of RMS jitter. The time axis representation of the measurements is absolutely linear because of the digitally controlled sampling. The only source of jitter and drift which could be observed is provoked by the single tone RF-source but a variety of commercial solutions are available for this. In the end, the application dictates the requirements on the RF clock generator and finally its price. The higher the ambiguity range is the lower the phase noise of the RF source must be. Its medium term frequency drift limits the maximum averaging time and thus the maximum dynamic range and its long term drift reduces the efficiency to remove systematic errors by calibration.

4. CORRECTION OF SYSTEMETIC ERRORS

The ability to design a stable system opens up interesting opportunities to correct systematic errors by software. Since decades, this is for example a usual method to improve the system performance of network analysers. In our case, more or less extensive methods can be applied which again are dependent on the requirements posed by a specific application. By concentrating on the systematic errors of the MLBS front-end and omitting the influence of other components (antennas, cables, …), the difference between the real form of the MLBS and the ideal one cause the most serious deviations. However they are of deterministic nature so they may be eliminated to certain extent. Two different methods of error correction will be shortly mentioned here which consider the same problem from different sides. A third one refers to Volterra Filtering which will be omitted here (see paper [3] for more information).

Source Calibration: The measurement method as represented in figure 2 is based on an impulse compression referring to an ideal reference signal rather than the actual stimulus. Thus it is not surprising that spurious signals appear after the compression but they may be removed if the correct stimulus signal is used for impulse compression. For that purpose calibration measurements must be undertaken by using DUTs with known behaviour. This method is well known from the network analyser technique and must not be explained in detail here. Figure 3 demonstrates the results of a simple response and isolation calibration.

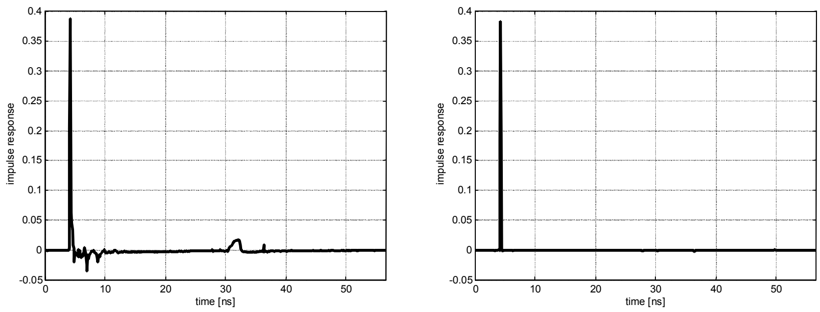

Figure 3: The system IRF before and after calibration. The FWHM value of the main peak is 170 ps.

Reference Channel: As with a network analyser, the actual stimulus signal will be captured by a second measurement channel. This signal will act as reference for the impulse compression. Such an arrangement provides the best results because it must no longer (as in the case above) be supposed that MLBS generator and sampling unit are absolutely stable over time. It is rather enough to require that only the difference between both measurement channels must not vary. This is technically quite less critical than to fix their absolute parameters particularly if one considers that both measurement channels are constructed in the same way. Nevertheless, a usual (network analyser) calibration technique will further improve the results.

5. EXPERIMENTAL RESULTS

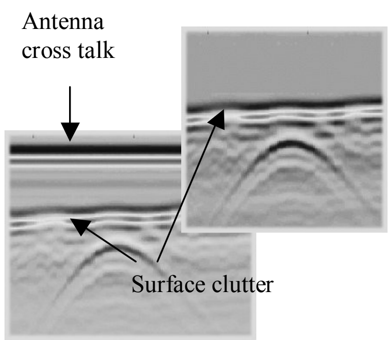

Figure 4: B-scan of a Tx-Rx-antenna pair indicating a buried object close to the surface. The left image represents the original data. On the right, the antenna cross-talk is removed. The perfect removal of antenna cross talk underlines the stability of the electronics.

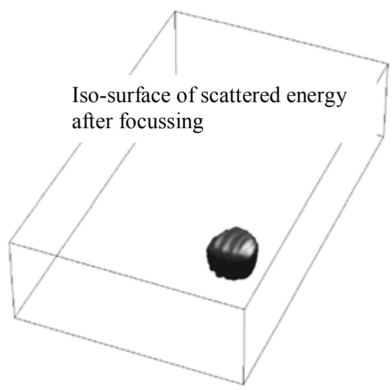

Figure 5: Results of data processing showing a small buried object (PPM2-mine). The data processing was performed by Vrije Universiteit Brussel.

A first integrated version of the UWB front-end was built for a GPR array with 6 transmitter and 6 receiver antennas for mine detection purposes (EU-Project: ESPRIT 29902 DEMINE[4]). The shift register, the binary divider and the T&H-circuit were integrated in 0.8 µm SiGe-HBT-technology. 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). 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.

The figures 4 and 5 summarise some measurement examples coming from a measurement campaign at a mine test field in Angola. Figure 4 represents a 2.5 m long B-scan measured along a demining lane. The GPR array gathered 21 of such B-scans at the same time each measured between different array elements. From this data the scattering energy of buried objects were calculated (see figure 5).

6. CONCLUSIONS

The function principle of a UWB front-end were explained and some experimental results were shown. The method works with low voltage MLBS signals and it is widely based on digital circuit conceptions and software routines. This promotes flexible system design, stable operation and 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.

7. REFERENCES

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