

Integrated UWB Radar Circuit for Base Band Applications from DC to 5 GHz

J. Sachs, M. Roßberg, P. Rauschenbach; TU Ilmenau
P. Peyerl, J. Friedrich; MEODAT GmbH Ilmenau
e-mail: sac@e-technik.tu-ilmenau.de

Motivation

Ultra wideband (UWB) short range radar and UWB measurements 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, industrial sensors, microwave imaging etc. In order to achieve high spatial resolution and maximum information, the spectrum of the sounding waves for such kinds of application should cover a very large bandwidth somewhere in the technically available frequency band. In air launching radar the stimulation band is generally located around a certain carrier frequency. It is, however, also often required that the waves must penetrate into solid or liquid bodies which restricts the upper frequency value. This demands carrier free waves having a relative bandwidth as large as possible. Correspondingly, the method is also often called base band radar. Since in the present case the technical challenge is seen mainly in mastering a large relative bandwidth rather than in creation of a large absolute bandwidth at high carrier frequencies, the following article will concentrate without restriction on base band type of a new UWB measurement principle in general. This is for the reason that when a base band head is available, it is possible to move the stimulation band by a carrier wave to an appropriate frequency plane.

One reason why UWB technology has not been applied industrially on a large scale up to now is that no adequate measurement principles have been available. The traditionally used impulse and stepped or swept sinewave methods work either too slow and too unstable or they are too expensive or inflexible. UWB technology has as such been limited to particular applications. To overcome these drawbacks a flexible principle is needed which may be adapted by simple measures to a vast number of different applications. The ultimate goal of our development is as such to create a kind of toolbox including hard- and software components which permit the configuration of systems with a high flexibility in technical (bandwidth, dynamic, measurement rate, EMI-resistance, number of channels etc.) and economical (the costs for example) parameters.

In what follows, the new UWB principle will be shortly introduced. For a more detailed description the reader is referenced to [1] - [4]. Then some results will be shown referring to an experimental model based on a SiGe-chip set manufactured at IHP Frankfurt/Oder.

Basic Principle

The basic requirement on flexibility can only be fulfilled if hardware is largely exchanged for software in the radar head. In an analogy to the catchword "software radio", such systems may be called "software radar". The hardware, especially the RF-hardware, should be reduced to a few elementary circuits. Furthermore, high dynamic signals should only appear in the digital domain and should be kept away from the analogue parts to reduce limiting effects and over-

loading of analogue circuits. Moreover, a low level modulation of analogue circuits increases their usable bandwidth and reduces power consumption.

The key to an UWB software radar is the choice of an appropriate stimulation signal because the whole device structure depends upon it. The most important demands directly influenced by the stimulus are summarised by the following points:

- Spread spectrum signal
- Low crest factor signal
- Simple generation in the RF- and microwave domain
- Periodic signal to perform under-sampling and synchronous averaging
- Simple synchronisation of sampling clock
- Fast and simple signal processing algorithms to perform impulse compression or corresponding operations
- High immunity against EMI

Pseudo random codes especially the **Maximum Length Binary Sequence** meet all these requirements.

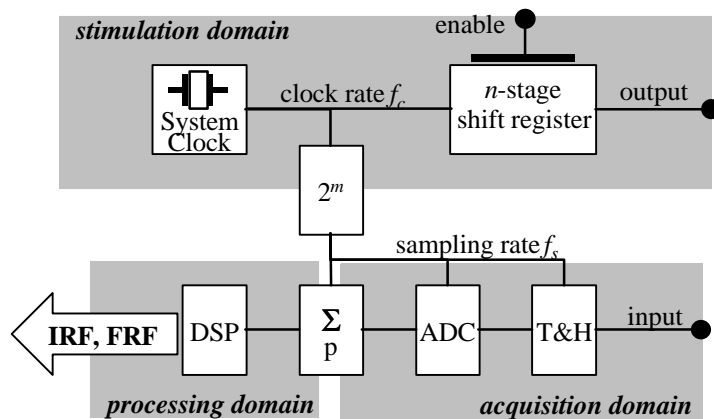


Figure 1 Basic block diagram of the UWB radar head

Figure 1 summarises the basic components of the base band variant of the MLBS radar head. Controlled by a single tone clock, a digital shift register generates the MLBS signal and a binary divider (2^m) provides the sampling clock. The measurement data are captured by a T&H-circuit, transformed into the digital domain (ADC), synchronously averaged (Σ) and finally processed (DSP) in an appropriate manner. Usually the DSP calculates the impulse response function (IRF) for

example, the frequency response function (FRF) or others corresponding to the required result of the measurement. The IRF is determined by correlating the captured signal with the MLBS by the fast Hadamard-algorithm. The calculation of the FRF uses the well known Fast-Fourier-Transformation.

Except for applications above 500 MHz bandwidth all components of the radar head are standard circuits which are commonly applied. Higher bandwidth systems, however, require customer integrated circuits (shift register, binary divider, T&H) as corresponding components are not yet commercially available. As shown in Figure 1, the radar head is divided into three parts of different processing speed. Their rates can be chosen largely independently one from each other since the three domains are only loosely coupled by the binary divider and the averager. Dividing factor and number of averages may be selected arbitrarily. As such, a large scope exists to optimise the system performance with regard to technical parameters and costs.

Fixed by the operation mode of the sampling clock control, the equivalent sampling frequency is given by f_c . Thus, the preceding principle may be used for short range radar applications and any wideband measurement respectively of linear systems in the frequency range from dc

sation line is required and that no RF switches are needed. A carrier frequency system (Figure 3) requires each two receiving channels in order to provide an I/Q-demodulation. The usable frequency band covers $f_{RF} - f_c/2 \dots f_{RF} + f_c/2$, where f_{RF} is the carrier frequency.

Experimental Results

Three systems of different circuit technology having an usable bandwidth up to 60 MHz, 300 MHz and 5 GHz were designed and tested. The digital part (12-bit/40-MHz-Video-ADC, averager, DSP) is identical for all systems.

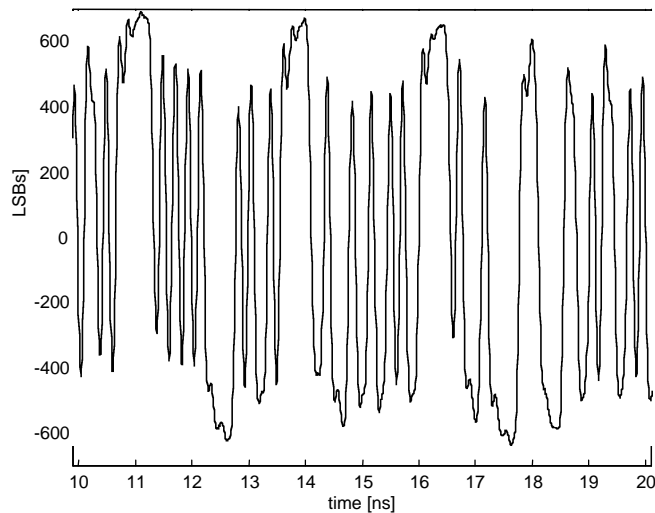


Figure 4 Section of the MLBS test signal.

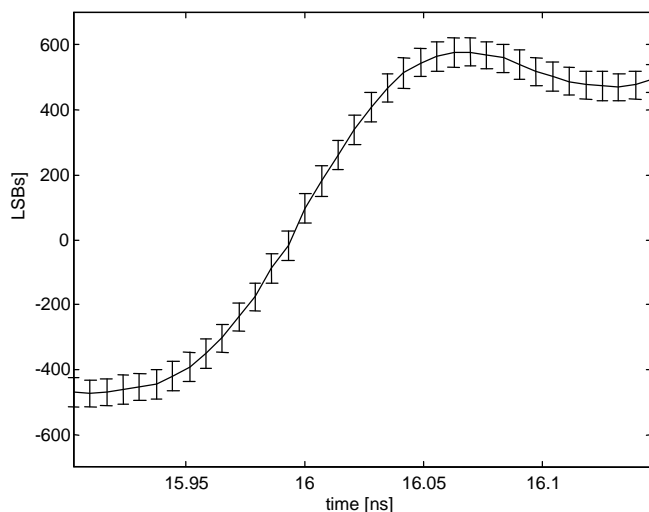


Figure 5 Any signal slope showing the standard deviation of all measurement points.

In case of the GHz-system, the RF-part (shift register, binary divider, T&H, clock recovery) was build from customer integrated chips manufactured in a 0,8 μm SiGe HBT technology [4]. The chips are mounted by wire-bonding or flip chip technique on a 4 layer LTCC. For measurement purposes, they were fitted in a RF test case provided with SMA connectors. All RF-ports are designed for differential and single-ended operation (100 Ω differential, 50 Ω single-ended). The output ports provide a power of about 3 dBm (single ended, 50 Ω load) / 6 dBm (differential, 100 Ω load). Digital control signals are fed by optocouplers. The shift register has 9 stages generating a MLBS with 511 bins. Applied in a radar system, this would provide 511 resolution cells. Its maximum toggle rate is about 10 GHz partly up to 12 GHz. The same is for the programmable binary divider and the pulse recovery stage. The rise and fall times of all these components are about 40 ps. The clock recovery stage is used in multi-channel applications for cascading the RF- and sampling clock lines. The bandwidth of the T&H stage is about 6 GHz. The maximum sampling rate of the T&H is in the range of several GHz and

its RMS noise voltage is about 3 ... 4 mV. The RF master clock of the whole system is provided either by a RF-synthesiser or a stable microwave DRO.

Due to the high measurement rate, the first step of the digital signal processing is done on hardware level by a FPGA. It comprises a fast cyclic averaging in order to reduce data rate and to increase dynamic [3]. The following procedures run on the DSP. Usually these are further averaging, calculation of the IRF, the FRF or others. Finally the data are transferred to a PC. There, a GUI supports the user with the data handling and the control of the measurement system. Complex data analyses are done off-line under MATLAB™.

Starting with figure 4, some current measurement results are summarised. Figure 4 represents a 10 ns portion of the period of the MLBS test signal stimulated by a 9 GHz clock. A detailed section of that signal is zoomed out in Figure 5 indicating also the standard deviation of the captured points. The measurements were done by connecting the input with the output port via an attenuator and a variable delay line. By that it was possible to capture the data with an equivalent sampling interval of 7 ps providing 16 point per bin instead of one point per bin in the actual operation mode.

As identifiable from Figure 5, the RMS-value of random noise is constant over the whole signal. An increase at signal slopes may not be observed which indicates an excellent jitter behaviour. The total noise n_{tot} at a signal flank is approximately given by:

$$n_{tot}^2 = n_a^2 + (c_v t_j)^2 \quad (1)$$

Herein n_a is the amplitude noise, c_v is the max. voltage slope steepness and t_j is the RMS-value of the jitter. Referring to Figure 5 and (1) and supposing that the difference between n_{tot} and n_a is at least smaller than 20 %, then the RMS-jitter of the whole system must be lower

than 2 ps. Note, that this jitter value is determined by the square sum of the jitter contribution of the shift register, the binary divider and the T&H-circuit.

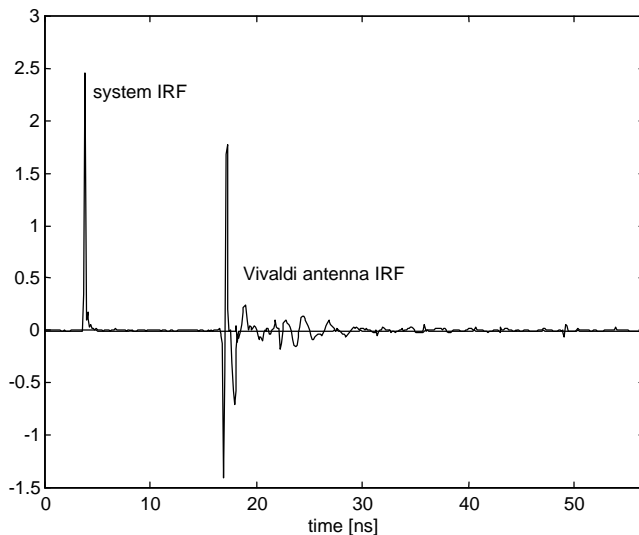


Figure 6 Measured impulse response functions (IRF)

Figure 6 shows the IRF^2 of the measurement system and the transmission between two differential fed Vivaldi-antennas. In order to give an impression of the measurement speed (wideband Doppler capabilities), a neon tube was placed in the radiation field of the Vivaldi antennas. The plasma within the tube acts as time variant scatterer which "pumps" with a 100 Hz rate. This effect can clearly

² Exactly spoken the shown IRFs are represented by the cross correlation function between measurement signal and ideal MLBS.

be seen in the different representations of the radargram (Figure 7). In the preceding case the measurement was about 34 000 IRFs per second and the data processing was done off-line.

Summary and Applications

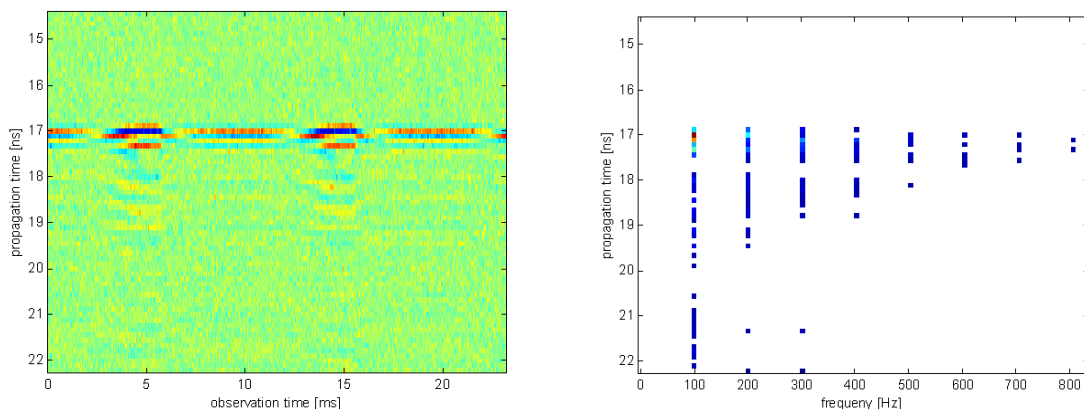


Figure 7 Left: "radargram" of a neon tube
Right: the observation domain is transformed into the frequency domain

A new ultra wideband measurement principle was introduced. It is based on the fusion of (mainly digital) hard- and software components which provide high flexibility for applications. It may be used in single and multi channel arrangements as base band or in carrier frequency systems. The measurement head may be divided into three parts working at different rates that can be selected relatively independently from each other. This ensures great flexibility within technical parameters and costs as indicated by the following three opportunities. With a high performance DSP and SiGe-front-ends, a measurement rate of 20 000 IRFs per second (with 511 points) will be possible opening wideband doppler performance and resolutions in the cm range. In the simplest case the whole "measurement head" is a PC-program controlling a digital output port as well as an ADC-board and performing the signal processing. For large scale applications or in cases requiring small sized sensors, a fully integrated version may be preferred.

In general, applications of the introduced principle are seen in those fields where the IRF or the FRF of static or moved objects is required to be measured. As well as for radar and microwave imaging, this could also be for moisture measurement, level indicators or other industrial sensors ,e.g., automotive sensors, medical sensors, wide band RLC-bridges etc.

References

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