Digital Stepping Correlator in a Wideband Radio Channel Measurement System

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ABSTRACT

This paper explains the realization and the performance of a digital stepping correlator in a wideband radio channel measurement system. Radio channel sounding is based on the estimation of the radio channel by use of maximum length sequence (m-sequence) as an effective sounding signal. Signal is detected using the stepping correlator receiver. The received signal is not converted to the baseband but the sampling is done direct from the IF-frequency. Digital implementation gives also a possibility to use a windowing property when those parts of the impulse responses that do not contain signal power are abandoned during measurements. The windowing enables high radio channel sampling rates.

INTRODUCTION

Design of the new radio communication concepts and devices needs accurate information of the real propagation channels for the good optimization of the systems. That is why the radio channel has been studied all over the world for years.

The most frequently used technique in the correlator receiver measurement systems is based on the analog sliding correlator technique. The Telecommunication Laboratory has two kinds of correlator receivers implemented to the wideband radio channel measurement system; the analog sliding correlator and the digital stepping correlator. The latter is the case in this paper. The sliding correlator based measurement system has been presented to the scientific forum twice before [1],[2].

The use of digital correlator gives a possibility to make measurements not only to the whole impulse response of the channel but also for those parts of the impulse responses where significant signal power is assumed to locate. This is because of the certain spreading code phase within the whole code cycle corresponds to the specified excess delay in proportion to the shortest propagation path. This process of cutting the impulse response is called *windowing*. When that windowing property is used the channel sampling rate is greater compared to the sampling rate of the sliding correlator. Analog sliding correlator techniques does not support the windowing.

THEORETICAL LIMITS

The dynamic range of the impulse response has been specified to be 25 dB [1]. That is the ratio between the maximum true correlation peak and the noise floor.

The effective sampling rate R_c of the estimated impulse response of the channel for the analog sliding correlator and the digital stepping correlator are

$$R_c = \frac{1}{kLT_c} = \frac{1}{L_w LT_c}\ , \tag{1}$$

correspondingly, where k is defined to be the time scaling factor of the sliding correlator, L is the length of the m-sequence in chips, T_c is the chip duration and L_w is the window size in chips. When the windowing property is not used $L_w=L$ ($L_w \le L$). All parameters in equations 1,3 and 4 can be calculated for the analog sliding system replacing L_w with k.

The maximum theoretical excess delay of the multipath component that can be detected without aliasing in the delay domain can be defined using equation

$$\tau_{\max} = LT_c = L_w T_c\ , \tag{2}$$

for analog and digital correlator, respectively.

The delay resolution, in theory, is one chip for both kind of correlators. Shifting the reference code by one chip in the correlation procedure gives, however, the delay resolution about two chips in digital system.

Using these parameters mentioned above the maximum unaliased Doppler shift f_d can be determined using equation

$$f_d \le \frac{1}{2LL_w T_c}\ . \tag{3}$$

Based on the Eq. 3 the digital stepping correlator receiver gives a relative small range to the Doppler shift if the whole impulse responses are calculated due to long measurement time of one impulse response. That limits the use of the receiver to the cases where the radio channel is expected to be relatively constant. If $L<k$ the Doppler range is larger in digital than analog implementation. The windowing property, however, gives more range at the Doppler domain to operate due to shorter measurement time of one impulse response as can be seen from Eq. 3.

The Doppler resolution Δf_d is given by equation

$$\Delta f_d = \frac{1}{N_{FFT} L L_w T_c} \ , \tag{4}$$

where N_{FFT} is the length of the FFT in post processing. Maximum value of N_{FFT} is ruled by the stationary of the channel. If N_{FFT} is chosen to be too large the Doppler resolution is more accurate than the fade rate of the channel which yields to Doppler shifts that do not represent the channel.

In theory, dynamic range of the impulse response reaches infinity if the side lobes of the autocorrelation function are removed. Theoretical dynamic range is limited in the real implementation by the AD-converter that has a finite 8 bits word length. That limits the dynamic range to 48 dB.

MEASUREMENT SYSTEM

At the transmitter the carrier frequency is modulated with the pseudo random noise sequence, namely m-sequence, and emitted by a power amplifier.

The maximum length spreading code is chosen because of its autocorrelation properties. Normalized autocorrelation function yields 1 when the transmitted code and the reference code are in the same code phase. In every other code phase the correlator outputs an absolute value that is inversely proportional to the code length *L*.

The length of the pseudo random noise (PRN) can be changed at the range of 63 to 65535 (2^6-1 ... 2^{16}-1) by the operator [1].

The feedback loop of the shift register of the PRN generator can be chosen by the user as well. At the receiver, spurious correlation peaks have been found and the feedback of the generator has been chosen so that the distance between the true and spurious correlation peaks is maximized. The spurious peaks have been found from the analog correlator as well [2].

The system operates at the fixed 15 MHz chip rate [1]. The modulation scheme is BPSK-modulation, without any data modulation, with signal bandwidth B = 30 MHz. The carrier frequency can be chosen in range from 100 MHz to 1500 MHz. The block diagram of the transmitter is presented in figure 1.

The Global Positioning System (GPS) is used to locate the transmitter and receiver during the field measurements.

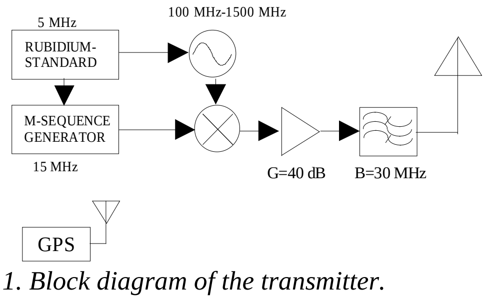

Figure 1. Block diagram of the transmitter.

In order to archive a phase coherence between the transmitter and the receiver common frequency standards have to be used in all ends. Signal sources are synchronized to the rubidium frequency standards that have been synchronized to each other before the measurement sessions.

At the receiver, the former analog sliding correlator has been replaced with the new, stepping correlator that has been implemented using digital technology. The received signal is sampled directly at the 195 MHz IF-frequency (f_{IF}) using quadrature sampling. RF to IF conversion is done at the separate down converter part and the f_{IF}-signal is the input signal to the digital correlator.

The block diagram of the receiver is presented in the figure 2. Sampling rate at the system is 60 MHz. At the front-end of the receiver is the digital step attenuator which is used in AGC-control. During the measurements the despread I- and Q-samples and log(I^2+Q^2) can be monitored by oscilloscope.

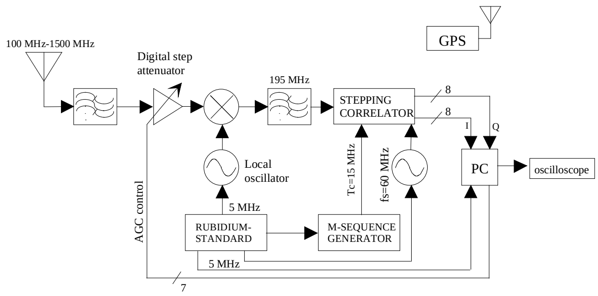

Figure 2. Block diagram of the correlator receiver.

The bandpass theorem for uniform sampling states that the acceptable sampling rate f_S must be [3]

$$\frac{2f_U}{n} \le f_S \le \frac{2f_L}{n-1} \tag{5}$$

where f_U is the highest and f_L is the lowest frequency component in bandpass signal and the integer n is [3]

$$1 \le n \le I_g\left[\frac{f_U}{B}\right]. \tag{6}$$

Operation $I_g[x]$ denotes the largest integer within x. In our case this parameter n has been chosen to be the largest possible value (n_{max}=7) so the sampling frequency (Eq. 5) can be as low as possible.

When the sampling rate has been chosen to be the lowest one the out-of-band signal components have to be rejected before the sampling very carefully. This is done using anti-alias filter before A/D-converter. Otherwise these unwanted signal components are aliased to the signal band and they are made the performance of the receiver worse. If the n has been chosen to be smaller than the maximum value the quard band between the signal bands can be found.

When quadrature sampling (a special case of 2^{nd} order sampling) is used, both the sampling rate and the IF-frequency have to be chosen very carefully to satisfy the requirements for obtaining I- and Q-values directly. Suitable sampling rates depend on the IF-frequency by equation [3]

$$f_S = \frac{4f_{IF}}{2n-1} = \frac{4\cdot 195\text{MHz}}{2\cdot 7 - 1} = 60\text{MHz} \ . \tag{7}$$

Due to the bandpass sampling (four samples per chip), the measurement signal spectrum will be shifted around the 15 MHz (Fig. 3).

If the parameters in Eq. 3 have been chosen carefully there will not be any aliasing caused by desired signal in the frequency domain, as shown in figure 3. Although desired signal do not cause aliasing, the input noise can and will always cause some aliasing if it is wider than desired signal [3]. This is an another reason to use anti-alias filter before A/D-converter.

An often used structure for digital downconversion to the baseband is shown in figure 4. After A/D-converter digital signal is split into two separate branches. The lower branch samples are Hilbert transformed by a digital Hilbert filter and upper branch samples are delayed to compensate the delay introduced by causal Hilbert filter in lower branch. Signals $a(nT_S)$ and $\hat{a}(nT_S)$ together form an analytic signal that contain only positive frequency components. Then the analytic signal is shifted to the baseband to form a complex signal $I(nT_s) + jQ(nT_s)$.

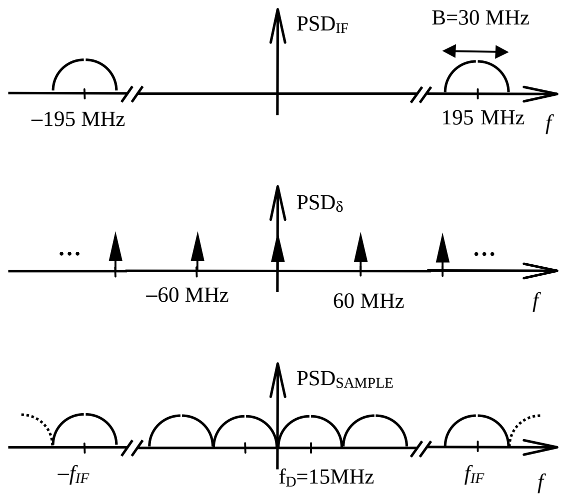

Figure 3. Principle figures of bandpass sampling in frequency domain (PSD = Power Spectral Density)

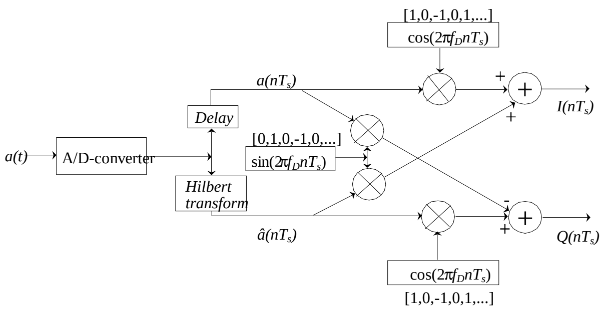

Figure 4. The idea of a digital downconversion to the baseband.

Because now $f_D = 1/(4T_S)$ the operands in sine and cosine generators are $2\pi f_D nT_S = n\pi/2$. This means that sine generator forms the sequence 0,1,0,-1, ... and cosine generator forms the sequence 1,0,-1,0, ... (Fig. 4). When quadrature sampling is used (as in this measurement system) the consecutive samples are taken so that a phase of the sampled signal has changed ±90° between the two samples. This means that Hilbert filter is now unnecessary to form an analytic signal due to quadrature sampling. The desired analytical signal can be formed from the actual sampled signal and the same signal delayed by one sample. This process approximates a receiver structure based on Hilbert transformation.

Further simplification for downconversion can be obtained from the fact that only every other sample is needed to be calculated because of zeroes from sine and cosine generator. The simplified and realised structure for digital downconverter used in this measurement system is shown in figure 5.

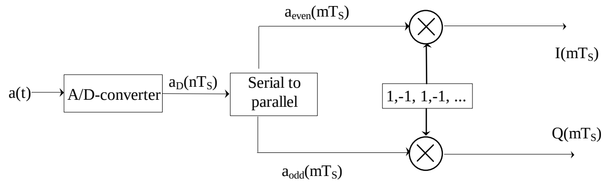

Figure 5. The structure of a practical digital downconverter.

After A/D-conversion (word length 8 bits), the data samples, $a_D(nT_S)$, are handled as consecutive I/Q-sample pairs. Serial-to-parallel conversion is applied to the digital samples, so that the first sample represents the in-phase sample $I(mT_s)$ and the other represents the quadrature sample $Q(mT_s)$. After serial-to-parallel conversion every other sample pair is multiplyed by -1 to complete the downconversion to the baseband.

The I/Q-interleaving procedure causes a slight error for each I/Q-pair because I- and Q-samples are taken at different time instances at the constructed system. Ideally both samples should be taken at the same time [3].

This error caused by unideal sampling is paid for a simpler receiver structure for quadrature sampling. The error depends inversely on the ratio of the chip frequency and the sampling frequency. Also, the pulse shape will affect for the level of error.

The correlator unit (Fig. 6) is controlled by a DSP-processor (TMS320C30). The reference PRN generator (Stepping Code Generator) is inside the correlator unit and it is realised by a single IC chip (Stanford Telecom STEL-1032), having three internal and independent code generators (registers). The independency of the code registers makes it possible to set the reference code into an arbitrary initial phase. While one register is used to feed the reference code to the correlator, the two other registers can be set up to other code phases. The stepping code phase is generated by starting the next code phase one chip earlier than the current phase. So, the last chip of the current reference code is doubled and it will be become to the first chip of the next code phase. The consequence is that the reference code will be stepped compared to the transmitted code by one chip.

The correlation is calculated over the whole m-sequence period before the integrator is dumped. After L_w recalculations in I- and Q-correlators the estimate of one channel impulse response is completed.

The real and the imaginary parts of the channel's impulse response are in the outputs of the I-correlator and Q-correlator (Stanford Telecom STEL-2410). Total word length of the I/Q-sample pair is 23 bits.

The realized digital correlator supports two independent windows that can be set up by the operator. This windowing property is not possible when the sliding correlator is used.

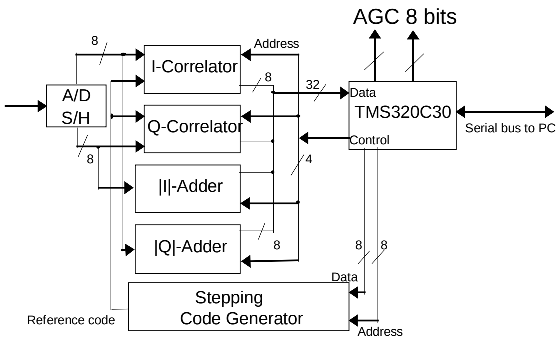

Figure 6. Simplified block diagram of the digital stepping correlator.

The reference signal level defines the rms-value that the A/D-input signal is tried to keep and can be defined by the operator. The complex signal power level is calculated using the sequences of the wideband I- and Q-samples. The final power estimate for the impulse response is determinated as the average value of the sample powers during the period LT_c. The difference signal between the estimate of the received signal power and the reference signal level before the A/D-converter is finally calculated and the automatic gain control (AGC) control is generated. The AGC period can be chosen by the operator and it has to be multiple of the impulse responses.

The absolute values of the I/Q-samples are summed in the same kind of correlator circuits that are used in calculations of the correlation (Fig. 6, $|I|$ -Adder and $|Q|$ -Adder). Those results are then averaged and the estimate of the wideband signal power is obtained. The estimated received power is then used in the AGC algorithm that controls the digital step attenuator (Fig. 2). Using the correlated I/Q-sample values the narrowband AGC-control can also be calculated and used to control the AGC attenuator value.

LABORATORY TESTS

The measurement system has been tested at the laboratory.

The tests were made using the coaxial cable as a single path channel between the transmitter and the receiver. The power amplifier is not used at the tests. The signal sources were synchronized to the same frequency standard that made the test system coherent.

The range of the input power is limited between -103 dBm ... -30 dBm. Within this range the power level at the input of the receiver does not introduce any distortion to the system, yet giving a 25 dB dynamic range for the impulse response. Because of the digital implementation and the different IF-parts of the receiver the range is better using digital correlator rather than analog correlator. Also the noise floor in the stepping correlator is 10 dB lower.

The correlator is dumped once during one code period. In the case of a single path channel the correlator output produces a single correlation peak repeated in time of LL_wT_c. The delay resolution of the system has been measured in laboratory using the radio channel simulator. The delay resolution was defined to be the smallest time difference between two equal power paths that can be separated. These tests showed that the delay resolution is about 125 ns (Fig. 7).

Without any time averaging in post-processing maximum noise components are about -30 dB below true correlation peaks when using chip length 511 and sufficient input power. This means that dynamic range of 25 dB is easily obtained. When using the same parameters and statistically averaging 256 consecutive impulse responses a noise floor was observed at level -40 dB.

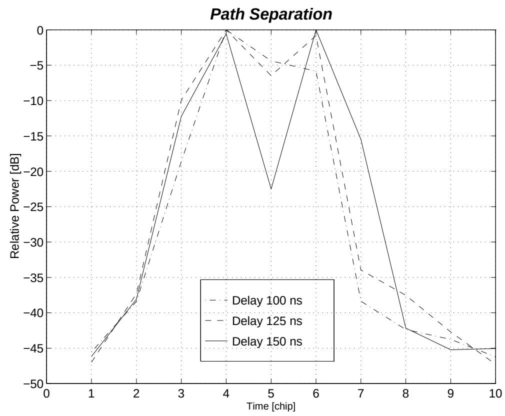

Figure 7. Definition of the delay resolution.

At the tests the minimum code length that can be used is found to be 511. The signal processor (TMS320C30) is not capable to handle samples that come faster if the code length is shorter. The idea of stepping correlator is not the limiting thing but the hardware used is.

During the laboratory tests a few spurious correlation peaks have been found. These peaks locate at definite delays, that depend on the length of m-sequence. When the statistical behavior of the spurious peaks is known their influence can be removed during the analysis of the measured data.

Windowing property is also tested at the laboratory and the window size of 15 chips has used which gives the measurement time $T_i=1/R_c$ = 0.51 ms and $f_{d.max}$= 980 Hz. The problem that arises when too small window is used is that because of the distance change between the transmitter and receiver, the correlation peaks roam and they might go outside the window. The output of the correlator is then zero and the information of the channel is lost.

At a present system the window is settled manually. One possibility to solve out the problem of roaming of correlator peaks is an automatic window place setting. It might be based, for example, on the power level crossing measurements of the correlator peaks.

Table 1 shows the comparison of the theoretical and measured performances between the analog sliding correlator and the digital stepping correlator in our measurement system when whole impulse responses are calculated. Values for the sliding correlator are taken from [4].

Windowing property improves the Doppler range at the digital implementation radically.

The benefit of the digital correlator is that the self-noise of the analog correlator is vanished. Also, if the spreading code length is risen the noise floor is dropped unlike in the use of sliding correlator [2].

Table 1. Comparison between different type of correlator receivers.

	Anal.	Digit.
Transmission power:	4 W (36 dBm)	same
Frequency range:	100-1500 MHz	same
Dynamic range of IR:	25 dB	25 dB
Received power range:	-83...-30 dBm	-85...-30 dBm
Chip rate:	15 MHz	same
Chip duration:	67 ns	same
Modulation:	BPSK	same
Length of m-sequence:	63-65535	511-65535 *
IF-frequency:	160 MHz	195 MHz
Time scaling factor k:	1000-20000 †	-
Delay resolution:	83 ns	125 ns
Maximum theor. delay:	4,2 µs...68.2 µs#	34,2µs...4,4 ms
Doppler resolution:	depends on the FFT	
Maximum doppler L=63:	119 Hz@k=1000	1,89 kHz **
I/Qsample rate:	60 kHz	60 MHz
Quantization:	16 bits	8 bits
Storing capasity:	1 GByte	same

† Five fixed values: 1000,2000,5000,10000 and 20000
# Self-noise limits the maximum value
* Hardware limits the code length to 511.
** If L=511 and windowing is not used $f_{d,max}$=29 Hz. Using windowing, $f_{d,max}$ increase significantly

CONCLUSION

The digital stepping correlator based on the direct IF sampling has been tested in a part of a wideband radio channel measurement system.

The comparison between the sliding and the stepping correlator shows that the stepping correlator gives worse measurement ranges for certain parameters based on the long calculation time of one whole impulse response. However, if the spreading code length is shorter than the time scaling factor of the sliding correlator, the digital correlator gives larger measurement ranges. New signal processors can give higher calculation rates which make it possible to use shorter codes than 511 which then gives shorter measurement times for the impulse response. That improves the performance of the digital correlator.

If the windowing property of the stepping correlator is used the digital correlator gives much higher channel sampling rates than the sliding correlator because those parts of the impulse responses that do not include any signal power can be abandoned. The improvement of the channel sampling rate is significant and very large Doppler spreads can be measured. The windowing is a property of the digital stepping correlator which cannot be used in a sliding correlator.

ACKNOWLEDGMENTS

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REFERENCES

[1] J.Talvitie et.al. (1992) "A Wideband Channel Measurement System for Aircraft Air-to-Ground Links", *Proceedings of the IEEE Second International Symposium on Spread Spectrum Techniques and Applications,* Yokohama, Japan, Nov. 29-Dec. 2, 1992, p.187-190.

[2] J.Talvitie et.al. (1994) "Self-noise as a Factor Limiting The Dynamic Range in Impulse Response Measurements Using Sliding Correlation", *Proceedings of the IEEE Third International Symposium on Spread Spectrum Techniques and Applications*, Oulu, Finland, July 4-6,1994, p.619-623.

[3] R.G. Vaughan et.al (1991) "The Theory of Bandpass Sampling", *IEEE Transactions on Signal Processing*, vol. 39, no. 9, September 1991, p.1973-1984.

[4] J.Talvitie (1993) "Wideband Channel Measurement for VHF and UHF Mobile Communication", *Licentiate Thesis*, University of Oulu, Oulu, Finland, 94 p.