Generation of Pseudo-Random Sequences for Noise Radar Applications

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Abstract — **Noise Radar Technology (NRT) is nowadays a promising tool in radar systems. It is based on the transmission of waveforms composed of many noisy samples, which behave as LPI (Low Probability of Intercept) and antispoofing signals. Each noisy sequence is theoretically uncorrelated with the others. In the paper we propose a scheme to generate a "tailored" pseudo-random sequences (limited in amplitude). It will be followed by an analysis of the main performances in terms of the Peak Side Lobe Ratio (PSLR) of the autocorrelation function, cross-correlation analysis to evaluate the orthogonality, bandwidth and energy efficiency.**

Keywords – Noise radar, low PSLR, orthogonal waveforms.

I. INTRODUCTION

For the first time in 1959 the concept of noisy signal was introduced to implement a system able to measure the distance using a noise modulated signal [1]. During the '60s - '80s the research became more intense [2], [3], [4]. In the '90s new methods for efficient generation of chaotic signals were developed in the millimeter band [5]. In the '2000s several research groups have developed new applications for noise radar and made significant contributions towards detection, surveillance, tracking, and imaging of targets [6]. More elements about features, problems and potential of noise radar systems can be found in [7]. The progress of technology now offers the possibility to produce and process noisy waveforms. In fact, high-speed and high dynamic range Digital-to-Analog Converters (DACs) and high-speed large-scale Field Programmable Gate Arrays (FPGAs) facilitate generating high-performance precision digital waveforms. Moreover FPGAs and ADCs allow directly sampling of fairly wide bandwidth signals, and modern high speed processors allow more and more sophisticated filtering and detection algorithms to be employed.

The main advantage of noise radars waveforms is to achieve high resolution in both range and Doppler, which can be independently controlled by varying the bandwidth B and the integration time T respectively (the compression ratio BT is related to the number of independent samples to be process).

In general, one of the most important problem of the compression technique is the masking effect due to the presence of high sidelobes at the output of the correlation filter, i.e. weak echoes can be masked by the strong ones. In pulse and CW radars this problem is solved exploiting time and frequency separation respectively. However in CW noise radars the noisy signal occupies the entire bandwidth at each instant of time, thus making impossible frequency separation and reducing radar sensitivity and detection range. Methods to solve masking effect can be conceptually divided in: (a) methods requiring extra signal processing at the receiver (for example the iterative algorithm known as CLEAN [8]) and (b) methods based on the waveform design with desired auto and cross correlation characteristics at the transmitter [9]. The main concept of (b) is filtering the noisy signal in the frequency domain with a window whose inverse Fourier Transform provides a low sidelobes level (low Peak Side Lobe Ratio). In the following a specific frequency window is used for the purpose to reducing the masking effect in transmitting section; it will be seen that a coherent azimuthal average of the outputs of the compression filter is necessary in order to achieve this goal. The waveforms design is one of the main focus of the research in modern radar systems, as MIMO (Multiple Input Multiple Output) radar [10] and Multifunction Radar [11], [12]. The MIMO and Multifunction radar applications require (in addition to the above mentioned low Peak Side Lobe Ratio) also good orthogonality properties (while maintaining the spectral occupancy into a limited bandwidth) and a low degradation in the main lobe (low Signal to Noise Ratio loss). Orthogonality may be imposed in time domain, in frequency domain or in signal space. Time division or frequency division multiplexing are simple approaches but they can suffer from potential performance degradation because the loss of coherence of the target response [15]. As a consequence, obtaining the orthogonality in signal space domain is the best choice, therefore this one will be considered in this paper. To obtain a low Signal to Noise Ratio loss an amplitude limiter (Zero-Memory-Non-Linearity transformation) will be used to maximize the transmitted power.

Fig. 1 presents a general architecture for coherent Noise Radar. The Digital Waveform Generator generates (I,Q) components of the transmitted noise signal which is stored in a memory and then modulated, amplified and transmitted. In the receiving section a coherent demodulator provides the base-band signal (I',Q'); the output of the Digital Correlator is the output of the filter matched to the generated signal (I,Q).

The paper is organized as follows. Chapter II describes the generation method based on a filtering of a complex white Gaussian process followed by a Zero-Memory-Non-Linearity transformation to implement the hard or soft amplitude

limitation. In Chapter III the criteria for setting the threshold of the soft limiter will be analyzed with respect to loss and orthogonality. Chapter IV contains some considerations regarding the PSLR. Chapter V reports conclusions and future perspectives.

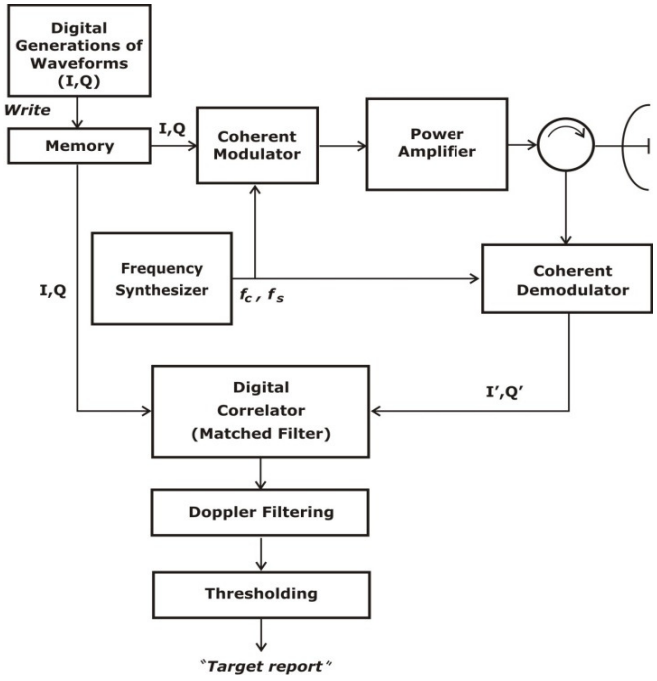

Figure 1. General architecture of a solid-state, coherent, Noise Radar.

II. WAVEFORMS DESIGN

The proposed scheme to generate pairs of pseudorandom signals is shown in Fig. 2. Starting from a pair of *white* Gaussian noise $x_1(n)$ and $x_2(n)$, the low-pass filter $H(f)$ has the goal of limiting the signal bandwidth and defining the autocorrelation function properties. The frequency response of the filter is obtained by $H(f) = \sqrt{S(f)}$ where for $S(f)$ a Blackman-Nuttall spectrum, [16], has been considered, whose inverse Fourier Transform gives a function with very low sidelobes (order of $-100\ dB$). The samples of $S(f)$, i.e. the desidered spectrum for $y(t)$, are:

$$S(k) = a_0 - a_1 cos\left(\frac{2\pi}{N}k\right) + a_2 cos\left(\frac{4\pi}{N}k\right) - a_3 cos\left(\frac{6\pi}{N}k\right) \quad (1)$$

for $k = 0,1, \dots, N-1$, where $a_0, \dots, a_3$ are defined in [16].

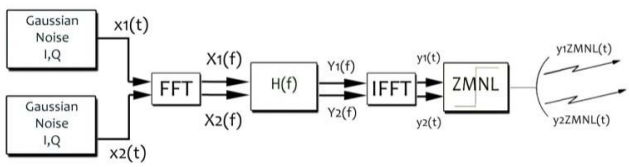

Figure 2. Block diagram to generate pairs of pseudorandom sequences.

The block diagram in Fig. 2, considering a single signal $y_{ZMNL}(t)$, represents the Digital Waveform Generator in Fig.1; a pair y_{1ZMNL} and y_{2ZMNL} is generated to calculate the cross-correlation (i.e. to check orthogonality).

The number of the generated samples N (for each sequence) depends on the bilateral band B, on the pulse length T and on the sampling frequency F_s. After filtering, the sequences $y_1(n)$ and $y_2(n)$ have limited bandwidth and non-constant amplitude. The latter causes "*losses*" in the transmission power (meaning that the transmitter is not operating in saturation and the average power is lower than the peak power). To reduce the loss, Mean Envelope to Peak Power Ratio:

$$MEPPR = \frac{\frac{1}{T}\int_0^T |s(t)|^2 dt}{max\{|s(t)|^2\}} \quad (2)$$

has to be close to 1; this is obtained using a Zero-Memory-Non-Linearity (ZMNL) transformation. This amplitude limitation may be either *hard* or *soft*.

Given a complex signal $(I + jQ)$, the *hard* limitation (to the unit level) is obtained by $\left(I' = \frac{I}{\sqrt{I^2+Q^2}}, Q' = \frac{Q}{\sqrt{I^2+Q^2}}\right)$, keeping the phase of the (unit level) signal unchanged. For a *soft* limiter (see Fig. 3) the unit level is reached if the input amplitude exceeds the threshold $L = k\sigma$, where σ is the *mode* of the Rayleigh distribution; $k = 0$ corresponds to the hard limiter case, while for $k \gg 1$ most input amplitude samples remains in the linear zone. For example with $k = 4$, 99.97 % is not limited.

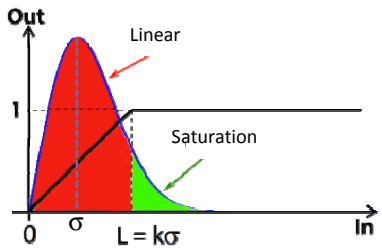

Figure 3. Soft limiter characteristic with threshold $L = k\sigma$.

III. CHOICE OF THE LIMITER THRESHOLD

The choice of the threshold (parameter k) is driven by the characteristic of the transmitter amplifier. For a given BT, variations of k produce limited variations on the PSLR, as shown in Fig. 4 for BT = 10000 considering two realizations for k = 0 and k = 5, and in Fig. 5 (continuous line) averaging 25 values of PSLR, for the compression ratio BT varying (BT = 5000, 10000, 30000, 50000, 100000). The difference between the two cases ($k = 0$ and $k = 5$) is less than 1 dB.

Considering the normalized cross-correlation function, Fig. 5 (dashed line) shows that also the mean cross-correlation peak is almost constant varying k with assigned BT; the difference is less than 1 dB. The minimum peak cross-correlation (for k = 0) results lower than 14 dB circa compared with the one of LFM *up* and *down* chirp, equal to $-10log_{10}(BT) - 3$, i.e. -43 dB for BT = 10000 [17].

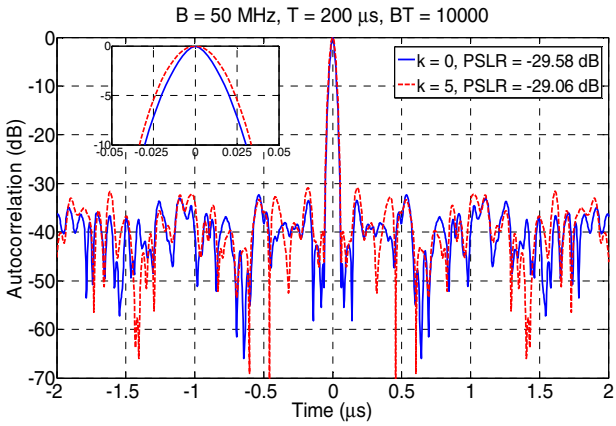

Figure 4. PSLR of a noise signal with hard limiter ($k = 0$) and witout amplitude limitation ($k = 5$).

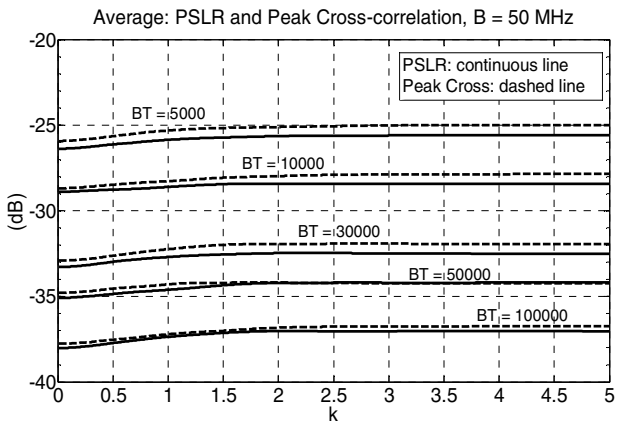

Figure 5. Mean PSLR (continuous line) and mean Peak cross-correlation (dashed line), averaging 25 PSLR values, versus k, varying BT.

The effect of the parameter k on the spectrum is significant. Fig. 6 highlights the difference between the spectrum of an amplitude limited sequence, hard limiter ($k = 0$) and the one of the same limited sequence, soft limiter ($k = 5$).

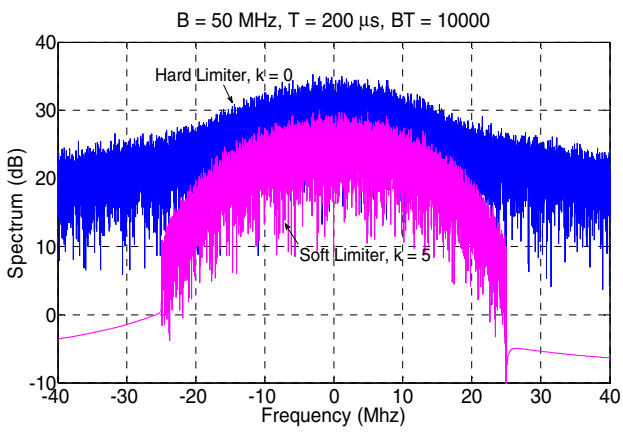

Figure 6. Comparison between the spectrum obtained with hard limiter ($k = 0$) and soft limiter ($k = 5$).

In general the limiter, being a ZMNL device, causes the increase of the spectrum tails outside the band. This effect is stronger when k is close to zero, while increasing k, the limiter becomes similar to a linear characteristic mitigating the rise of the tails [18].

In many applications the minimization of the power loss in transmission may be an important requirement. The power loss can be written as function of the threshold $L = k\sigma$:

$$Loss_{(dB)}(k\sigma) = 10 \cdot log_{10}\{MEPPR(k\sigma)\} \qquad (3)$$

Fig. 7 shows the power loss versus k. For $k = 0$ the amplifier works always in saturation (no loss). With a quasi linear characteristic, the loss is $10 \div 11$ dB.

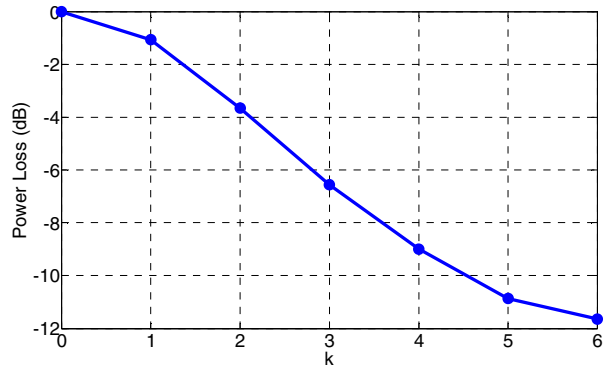

Figure 7. Loss of power versus the soft limiter parameter threshold (k).

IV. PSLR AS FUNCTION OF THE COMPRESSION RATIO

For random sequences filtered with a Blackman-Nuttall frequency window, the $PSLR$ depends on the number of independent ($N = BT$) samples of the signal with an average value of $-10 \cdot log_{10}(N)$. The signal has $N = BT$ (i.e. the compression ratio) samples only if it is sampled at sampling frequency $F_s = B$. For example, if this is the case, a $BT = 10000$ theoretically produces $-40\,dB$ sidelobes; however being the generated sequences (in according to the scheme of Fig. 2) realizations of a noisy random process, the $PSLR$ of each autocorrelation is really 13 dB worse (i.e. greater) in comparison with the theoretical case.

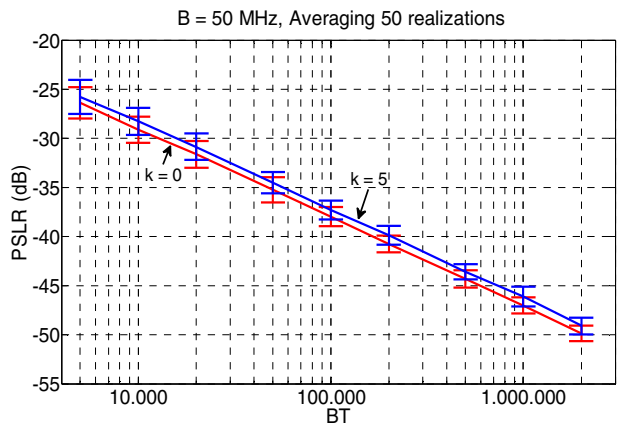

Figure 8. Mean PSLR (with confidence interval of twice the standard deviation) versus BT (each value is obtained averaging 50 PSLR values).

Fig. 8 shows the mean $PSLR$ (with confidence level of twice the standard deviation) versus BT, for $B = 50\ MHz$ and a threshold of $L = 0$ and $L = 5\sigma$. The difference between the two cases is very small ($\sim 1\ dB$).

In order to reduce the PSLR, the outputs of the correlation receiver can be averaged coherently (if we consider each realizations referred to a single angular section, the average can be called "azimuthal") obtaining an integration gain related with the number of averaged realizations, i.e. 12 dB averaging 16 realizations. Fig. 9 shows the PSLR improvement with respect to the averaged realizations.

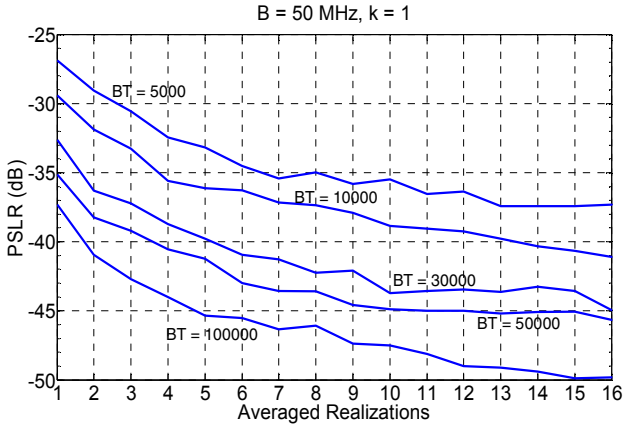

Figure 9. PSLR versus the number of averaged outputs.

The improvemeny in PSLR level is significant, but the azimuthal average leads to a poor angular resolution. To avoid this issue you can still consider to use more than one realization for each angular direction with the use of a Pulsed (increasing the scanning time) or a Continuous Wave-MIMO radar (with orthogonal waveforms).

V. Conclusions and Future Perspectives

A procedure for the generation of "tailored" pseudorandom signals has been proposed based on the use of a Low Pass Filter following by a soft limiter, whose threshold L has been set equal to $k\sigma$ (σ is the mode of the Rayleigh amplitude distribution). A soft limiter with $L = \sigma$ showed a low power loss of only 1 dB, while varying k the variations of the PSLR and the cross correlation are very limited (1 dB). Moreover, to improve the $PSLR$, it is appropriated to average coherently in azimuth the outputs (8 or 16 realizations) of the correlation receiver.

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