RANDOM WAVEFORM DESIGN

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ON THE DESIGN OF WAVEFORMS FOR NOISE-MIMO RADAR

G. GALATI AND G. PAVAN

Multiple-Input-Multiple-Output (MIMO) radar is an emerging technology that has significant potential for advancing the state-of-the-art of modern radar systems. Unlike standard phased-array radar, a MIMO Radar system can transmit, via its antennas, multiple signals that may be correlated or uncorrelated with each other. The orthogonal property is required for the transmitted signals to better separate them in reception. Although orthogonality may be imposed in the time, in frequency or in signals coding domain, to avoid changes in the radar cross-section of the target and undesirable Doppler effects, the waveforms have to be transmitted simultaneously and at the same carrier frequency. As a consequence, the orthogonality in the signals domain is the best choice and to successfully utilize such systems signal design plays a critical role. Good candidates as orthogonal signals for MIMO radar are the Phase Noise signals. In this paper, after an introduction to MIMO radar systems, we present the main characteristics of these signals through a statistical characterization, including an analysis of the autocorrelation, cross-correlation and spectral properties. Finally two novel methods to generate phase Noise signals will be proposed, i.e. a recursive method and non-recursive (closed form) one. Preliminary results will be presented.

Keywords: MIMO, Orthogonal Waveforms, Phase Noise.

1. INTRODUCTION

Recently a new field of radar research called Multiple Input Multiple Output (MIMO) radar has been developed [1], which can be thought as a generalization of the *multistatic radar* concept.

This kind of radar, as its name indicates, can be broadly defined as a radar system deploying multiple antennas to simultaneously transmit arbitrary waveforms and utilizing multiple antennas to receive the relevant echo signals.

The key ideas of MIMO radar concept has been picked up from communications, where the MIMO technique is used to increase data throughput and link range and to overcome the fading effects without neither additional bandwidth nor more transmission power. Conversely, a spatial diversity gain benefit is obtained in MIMO communications, often achieved by transmitting the same signal through different subchannels and combining the information at the receiver. Diversity gain is used against channel fading enhancing the link reliability of the system. Radar systems also suffer from fading (more precisely, fluctuation of the radar cross section) when there are complex and extended targets as it is the case very often.

It has long been understood that common radar targets are complex bodies, and large scintillations in the amount of energy back-scattered by a complex target can occur with very small changes (e.g. fractions of one degree) in the illuminating direction. If the antennas of MIMO radar are widely separated such that different antennas observe different aspects of the target, the target returns result from independent illuminations and can be combined together leading to a spatial diversity gain.

Diversity gain is only one of two key gains that MIMO communications can provide. The other gain is called spatial multiplexing, which expresses the ability to use the transmit and receive antennas to set up a multidimensional space for signaling. Then it is possible to form uncoupled, parallel channels that enable the rate of communication to grow in direct proportion to the number of such channels.

Similarly, in MIMO radar, a multidimensional signal space is created when returns from the multiple scatterers of a target combine to generate a rich backscatter. With proper design, transmit-receive paths can be separated and exploited for improving radar performance.

The transmit and receive antennas in a MIMO radar may be in the form of an array (see Fig. 1) and the transmit and receive arrays can be co-located (coherent MIMO) or widely separated (statistical MIMO).

Although MIMO radar system resembles phased-array radar system, there is a fundamental difference between these two approaches. In fact, unlike a standard phased-array, which transmits scaled, time-delayed version of a single waveform, MIMO radar systems transmits multiple signals and this waveform diversity enables superior capability and performance compared with standard phased-array radars.

In much of the current literature it is assumed that the waveforms coming from each transmit antenna are orthogonal. Although this is not a strict requirement for MIMO radar, orthogonality can facilitate the process of separation of the simultaneously received signals, avoiding the burden of further processing.

Orthogonality may be imposed in the time domain, in frequency domain or in the signals space. Time division or frequency division multiplexing are simple approaches but they both can suffer from potential performance degradation because the loss of coherence of the target response [15]. As a matter of fact the scattering response of the target or of the background (clutter) is commonly time-varying or frequency selective, limiting the ability to coherently combine the information from the antenna elements. As a consequence obtaining the orthogonality in the signals domain is the best choice.

The paper is organized as follows. Chapter 2 describes the MIMO radar system, underlining its significant characteristics, classifications and the main differences with respect to the classical phased array radar. In addiction, a general model for the signals transmitted by the antenna array elements is proposed.

Chapter 3 underlines the importance of the orthogonal waveform design and presents the Phase Noise signals as a good solution to the problem. This chapter contains the description of a new technique to generate phase noise signals. It is based on an iterative procedure that permits to obtain low Peak-Side-Lobe-Ratio (PSLR) or Peak-to-Average Ratio (PAR), limiting the spectrum in a desired band. Chapter 4 contains the conclusions.

2. INTRODUCTION TO MIMO RADAR

MIMO radar is capable of significantly improving target detection, parameter estimation, tracking and recognition performance, using multiple transmit and multiple receive antennas (see Fig. 1). These antennas may be closely spaced in the form of an array or may be widely spaced forming a "netted radar like" structure. In this paper we refer to the former case.

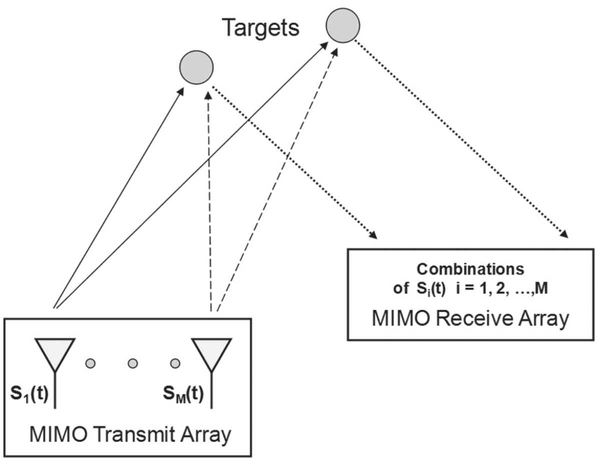

Fig. 1. Multiple transmit and multiple receive antennas for MIMO radar system

Every antenna element in a MIMO radar system (unlike standard phased-array radar which transmits delayed versions of a single waveform) can transmit different waveforms (waveform diversity). These may be orthogonal, mutually uncorrelated or linearly independent.

To benefit from this diversity, in the MIMO radar receiver there are as many matched filters as the number of transmitted signals. If the number of transmitting antenna elements is M and the number of receiving antenna elements is N, there are MN outputs of these matched filters totally (Fig. 2). MIMO radar processes these outputs jointly to decide whether a target is present or not.

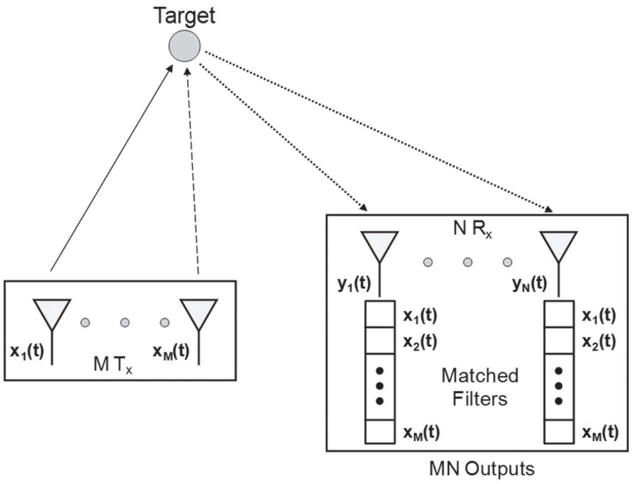

Fig. 2. Separation of different signals at the receiver: M matched filters to every transmitted signal in every receiver

As regards the kinds of MIMO radar systems, they can be classified into two categories according to their configurations and in particular to the distance between the antenna elements [1].

In the first category, referred to as *coherent* MIMO radar, the transmit and receive array elements are closely spaced so it is assumed that the target's scattering response is the same for each antenna pair, up to some small delay (the antennas are close enough such that all the elements view the same aspect of the target, or more precisely, are in the main lobe of the diffraction pattern of the target).

In the second category, referred to as *non-coherent* (or *statistical*) MIMO radar, the elements are broadly spaced, providing an independent scattering response for each antenna pair (the antennas are widely separated in order to capture the spatial diversity of the target's RCS).

2.1. Signal model

Consider a MIMO radar system that has a transmit and a receive array consisting of M and N elements respectively. Also, let denote the location parameter(s) of a generic target, for example, its azimuth angle and its range.

Under the assumption that the transmitted signals are narrowband and that the propagation is non-dispersive, the received signal can be written as [2]:

$$y(t) = diag(b(\theta)) \cdot A \cdot diag(a(\theta)) \cdot x(t-\tau) + w(t), \quad (1)$$

where the vectors $x(t)$ and $y(t)$ represent the transmitted and received signals:

$$x(t) = \left[x_1(t)x_2(t)\dots x_M(t)\right]^T \quad (2)$$

$$y(t) = \left[y_1(t)y_2(t)\dots y_N(t)\right]^T \quad (3)$$

and $w(t)$ denotes the interference-plus-noise term; A is a $N \times M$ matrix whose entries correspond to the bistatic RCS between each pair of transmitter and receiver; $a(\theta)$ and $b(\theta)$, which are some functions of the target location θ, are the $M \times 1$ transmit steering vector and the $N \times 1$ receive steering vector, respectively:

$$a(\theta) = \left[e^{-j\psi_1} e^{-j\psi_2} .... e^{-j\psi_M} \right]^T , \tag{4}$$

$$b(\theta) = \left[e^{-j\varphi_1} e^{-j\varphi_2} .... e^{-j\varphi_N} \right]^T \tag{5}$$

where:

$$\tau = \tau_{t1}(\theta) - \tau_{r1}(\theta) , \tag{6}$$

$$\psi_m = 2\pi f_0 (\tau_{tm}(\theta) - \tau_{t1}(\theta)) , \tag{7}$$

$$\psi_n = 2\pi f_0 (\tau_{rn}(\theta) - \tau_{r1}(\theta)) , \tag{8}$$

with τ_{tm} time delay between the target and the m^{th} transmit antenna and τ_{rn} represents the time delay between the target and the n^{th} receive antenna.

2.2. Coherent MIMO Radar

Coherent MIMO radar resembles the phased array radar but every antenna element sends different waveforms (waveform diversity) and this diversity enables superior capabilities as compared to standard phased-array radar. For coherent MIMO radar the benefits are [3], [5]:

(*a*) *Higher resolution*: the performance of MIMO radar systems can be characterized by a virtual array constructed by the bi-dimensional convolution of the real transmit and receive antenna locations, assumed to belong to a common plane. This virtual array can be much larger than each constituting a real array. The aperture extension results in narrower beams and therefore in a higher angular resolution and a better detection performance. Moreover, some of the virtual sensor locations are identical, which can be interpreted as spatial tapering, and results in lower side lobes.

(*b*) *Extension of spatial coverage*: in conventional radar systems several directional beams are usually transmitted in order to scan a given region of interest and the time on target is equal to the total interval assigned for covering the region of interest divided by the required number of beams [4], while MIMO radar transmits orthogonal signals, with virtually omnidirectional beams and hence with an extended spatial coverage; therefore, the time on target for each beam increases, and may be set equal to the interval which is assigned to scan the whole area.

(*c*) *Transmit beam-pattern synthesis*: through the choice of a signal cross-correlation matrix, it is possible to create spatial beam-patterns ranging from the high directionality of phased-array systems to the omni-directionality of MIMO systems with orthogonal signals. In detail, by properly designing the cross-correlation matrix of the transmitted signals $R = a^H(\theta)\langle x(t)x^H(t)\rangle a(\theta)$, where $\langle\cdot\rangle$ denotes time average, it is possible to maximize the total spatial power at a number of given target locations, or more generally, to match a desired transmit beam-pattern and minimize the cross-correlation between the transmitted signals at a number of given target locations [6]. Two specific problems are addressed. On one hand there is the optimization problem of finding the matrix R which makes the transmit beampattern close to a desired beampattern. This is approached using convex optimization techniques. On the other hand there is the not easy problem of designing multiple constant-modulus waveforms with a given cross-correlation R.

(*d*) *Direct application of adaptive techniques for parameter estimation*: because of the different phase shifts associated with the different propagations path from the transmitting antenna to the targets, these independent waveforms are linearly combined at the target locations with different phase factors. As a result, the signals reflected form different targets are linearly independent of each other. Therefore the direct application of adaptive techniques becomes possible without the need for secondary range bins or even for range compression [3]. Example of adaptive array algorithms applied to MIMO radar are Capon and APES (Amplitude and Phase Estimation). The paper [7] discusses these adaptive radar algorithms.

Summing up, MIMO radar systems could have better (i) resolution, (ii) parameter estimation accuracy and (iii) interference rejection capability.

3. ORTHOGONAL WAVEFORM DESIGN

The waveform design and optimization is one of the main focuses of the research in multistatic and multifunction radar [8], [16]. In MIMO radar applications typically M codes are required in the set, where M is the number of transmit elements. The main requirements of a pair of signals with complex envelope s_i and s_j with $i, j = 1,...,M$, pulsewidth T and same power, are defined by:

- Peak Side Lobe Ratio $(< 30dB)$

$$PSLR = \frac{\max_i(s_i)}{\max_k(m_k)} , \tag{9}$$

where s_i = *sidelobe samples*, m_k = *mainlobe samples*.

– Crest Factor (C) or Peak-to-Average Ratio (PAR), i.e. the peak amplitude of the waveform divided by the *rms* value of the waveform $s(t)$:

$$C = \frac{\max(|s(t)|)}{\sqrt{\frac{1}{T}\int_0^T |s(t)|^2 dt}} . \tag{10}$$

– Mean Envelope-to-Peak Power Ratio:

$$MEPPR = \frac{\frac{1}{T}\int_0^T |s(t)|^2 dt}{\max(|s(t)|^2)} . \tag{11}$$

It results: $MEPPR = \frac{1}{C^2}$.

– Normalized cross-correlation:

$$r_{ij}(t) = \frac{|R_{ij}(t)|}{|R_{i(j)}(0)|} , \tag{12}$$

where $R_{ij}(t) = \int s_i^*(\theta) s_j(t+\theta) d\theta$, $i \neq j$ measures the orthogonality, the desired value is $r_{ij}(t) < -30dB$.

– Spectral band occupancy; sometimes this item is overlooked, especially when noise-like waveforms are concerned, but it is of paramount importance in most real-world radars.

As explained before, the MIMO radar waveforms orthogonality in the signal space is preferred. Thus orthogonal waveform design plays a critical role in determining the feasibility of MIMO radar.

A good candidates to design deterministic signals that satisfy the orthogonal requirements are the well-known *"up"* and *"down"* chirp (Linear-FM and Non-LFM) [9], but in this case only one pair of signals can be defined. To obtain *M* pairs of signals the Costas codes represents a possible solution [10]. In addition Alltop sequences can be considered [11].

More recent research on orthogonal signals proposed the use of normal or interleaved OFDM techniques [12]. The main limitation of the OFDM approach is due to the non-constant envelope of the signals, i.e. $MEPPR < 1$, the transmitter does not work at its maximum power.

Another class of waveforms, i.e. the Phase Noise signals [13], has two main advantages as compared to the signals introduced before. The former is the possibility to generate a large number of orthogonal signals, which is of great importance in MIMO radar systems. The latter is about the detectability; in fact they are random signals so they place limitation on the detection, the identification and the eventual spoofing of the signal, an element of great importance in many military applications which require low detectability of the active system. Finally the *MEPPR* can reach the unity.

For a phase noise signal the complex envelope can be written as:

$$s(t) = A \cdot \exp\{j\varphi(t)\} \cdot rect_T(t)\,, \qquad (13)$$

where *A* is the constant amplitude, $rect_T$ is 0 outside the interval $[-T/2, +T/2]$ and 1 inside it (with *T* being the pulse length) and $\varphi(t)$ is the phase process modulating the noise signal $s(t)$.

In the following we present three methods to generate the phase noise signals highlighting their strengths and weaknesses.

3.1. Phase Noise Signals

In [13] Axelsson supposed for $\varphi(t)$ a zero-mean Gaussian process with root mean square (rms) σ and a given power with density spectrum within the band b. He showed that the normalized autocorrelation function of the signal $s(t)$ can be written in a closed-form expression as:

$$R(\tau) = \exp\{-\sigma^2[1-\rho(\tau)]\}\,, \qquad (14)$$

where $\rho(\tau)$ is the correlation coefficient of $\varphi(\tau)$. For example $\rho(\tau) = \frac{\sin(\pi b\tau)}{(\pi b\tau)}$ for a constant spectrum within the bandwidth *b* and zero outside.

Of course, $R(\tau)$ depends on the bandwidth b, on the pulse length T and on the *rms* phase fluctuation σ.

The bandwidth b is related to the width of the main peak and therefore, it determines the range resolution. An increase of *T*, and consequently of the compression ratio (the time-bandwidth product of the generated signal), causes a reduction of the range sidelobe level, whereas the mainlobe width remains fixed being independent of *T*. Finally the *rms* σ has two different effects. The former is on the sidelobe level: an increase of σ causes a decrease of the sidelobe level and an improvement of the PSLR.

The latter concerns the resolution. The *rms* value in fact establishes a connection between the bandwidth of the modulated signal and the bandwidth of the modulating signal. In detail, when σ increases the final bandwidth increases too. As a consequence a high *rms* value of σ gives an improved resolution (Fig. 3).

In [13] a simple relation between the *rms* bandwidth of the phase modulated signal (B_{rms}) and the *rms* bandwidth of the phase modulating noise (b_{rms}) has been found:

$$B_{rms} = \sigma \cdot b_{rms}\,. \qquad (15)$$

On the other hand, as regard the sidelobe suppression, the expression of the autocorrelation function introduced in [13] would show a progressive improvement of the sidelobe suppression as σ increases. However the periodic nature of $\varphi(t)$ with a folding in the $[-\pi, +\pi]$ interval has been neglected in [13], and in reality, the model can be used only for values of σ significantly smaller than π.

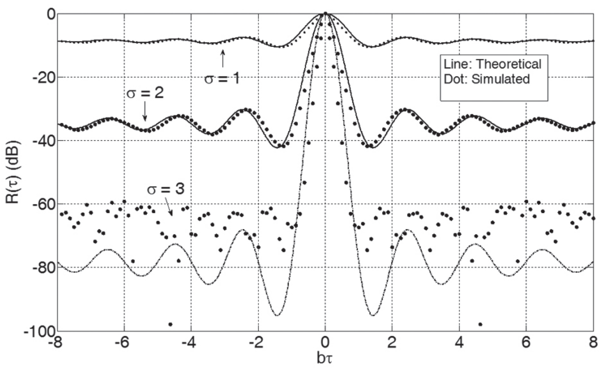

Fig. 3. Normalized autocorrelation for Phase Noise (compression ratio = 1000)

The Gaussian noise, used to modulate the signal phase, is to be compared with as a uniform distribution in the range $[-\pi, +\pi]$ with a standard deviation of $\pi/\sqrt{3} \cong 1.8\,rad$. Therefore, if σ is too large $(\sigma > \pi/\sqrt{3})$, the resultant phase does not have a Gaussian distribution and the mathematical formulation introduced in [13] does not apply. This is shown, inter alia, in Fig. 3 where the difference between Axelsson's theory and experiments (by simulation) is clear for $\sigma = 3$.

On the other hand, considering simulations and the relation with a potential real application, it would be better to generate the signal through a white Gaussian process with its *in phase* and *in quadrature* components (I, Q) that are band-limited as desired. This is described in the ensuing section.

3.2. An iterative algorithm to generate Phase Noise Signals

To control the spectral width and to reduce the PSLR of the generated phase noise signals, we propose an iterative algorithm as shown in Fig. 4. It is based on alternative projections in frequency and in time domain.

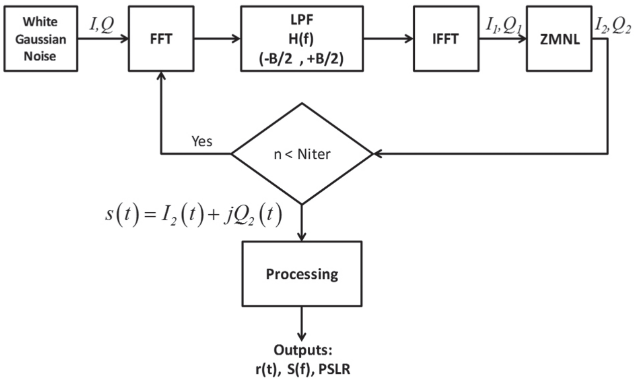

Fig. 4. Block diagram of the iterative algorithm to generate phase noise signals. *Legenda*: LPF = Low Pass Filter, ZMNL = Zero-Memory-Non-Linearity, N_{iter} = number of iterations

The filtering is implemented in frequency domain while the amplitude limitation (ZMNL = Zero-Memory-Non-Linearity) in time domain. The input to the algorithm is a *zero-mean white complex Gaussian process* $(I + jQ)$ with power $2\sigma^2$.

First we consider for ZMNL a *hard limiter*, i.e.:

$$I_2 = \frac{I_1}{\sqrt{I_1^2 + Q_1^2}}, \qquad Q_2 = \frac{Q_1}{\sqrt{I_1^2 + Q_1^2}}. \qquad (16)$$

Fig. 5 shows the obtained PSLR versus the number of iterations considering three different random sequences for the white Gaussian noise. The PSLR converges after some tens of iterations to –31 dB in the best case and it varies from –24 *dB* to –31 *dB*.

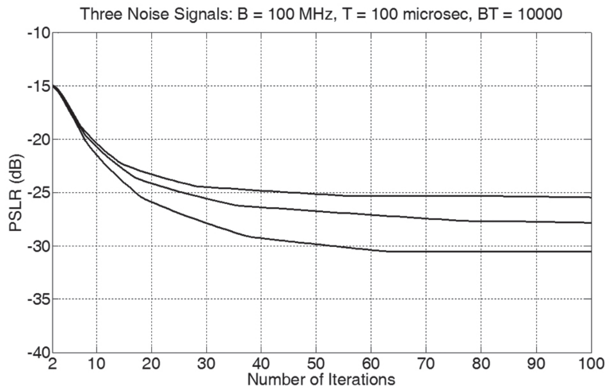

Fig. 5. PSLR versus the number of iterations. Three examples of convergence

Fig. 6 shows the normalized autocorrelation (around the main lobe) for two randomly generated phase noise signals.

Fig. 7 reports an example of density spectrum in comparison with them of Linear and Non-Linear *up* and *down* chirp. Due to the Low Pass Filter (LPF) of Fig. 4 the spectrum remains strictly band-limited as desired.

With respect to the orthogonality property, in Fig. 8 the cross-correlation is shown for a pair of generated phase noise signals. In comparison with the up and down chirp (LFM and NLFM), a degradation of 8-10 *dB* results.

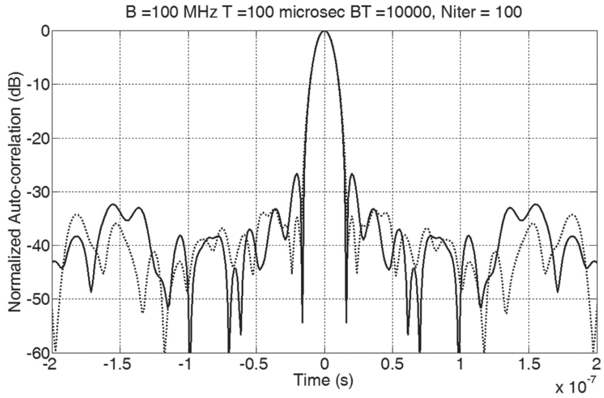

Fig. 6. PSRL near the mainlobe for two generated phase noise signals with a band of 100 MHz and a compression ratio of 10000

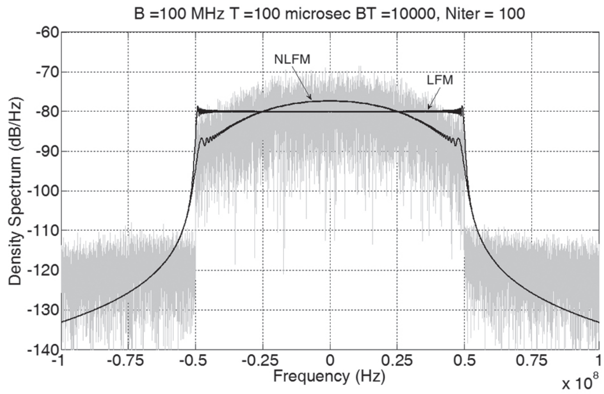

Fig. 7. Density spectrum of a phase noise signal with an allocated band of 100 MHz

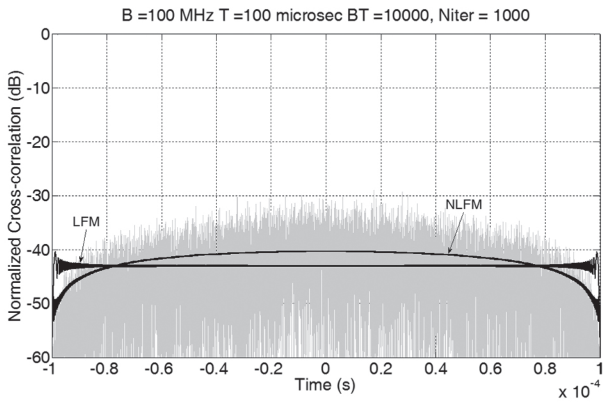

Fig. 8. Normalized cross-correlation of a pair of phase noise signal with a band of 100 MHz. Compression ratio of 10000. In black the cross-correlations of the pair up and down chirp (LFM and NLFM)

Considering now an *amplitude soft limiter* for the ZMNL (see Fig. 9 for the I/O characteristic):

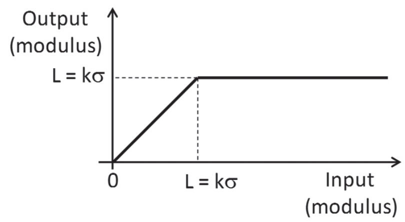

Fig. 9. Soft limiter I/O characteristic

and indicating with G_p the power gain of the Low Pass Filter (LPF) of Fig. 4, the Mean Envelope to Peak Power Ratio has been evaluated.

It depends on the number of iterations *m*, and on the ratio $k = \frac{L}{\sigma}$ being *L* the threshold of the soft limiter [17]. It results:

$$MEPPR = \frac{2(G_p)^{m-1}}{k^2}\left[1 - \exp\left(-\frac{k^2}{2(G_p)^{m-1}}\right)\right] \qquad (17)$$

with $m = 1,2,\ldots$.

Fig. 10 shows that only four iterations (in the worst case when $L = 5\sigma$ and the effect of the limiter is negligible) are needed to obtain a MEPPR between -1 *dB* and 0 *dB*.

By increasing the number of iterations, as shown in the case of hard limiter, the PSLR decreases up to −30 *dB circa* as shown in Fig. 11 for two different values of the threshold *L*.

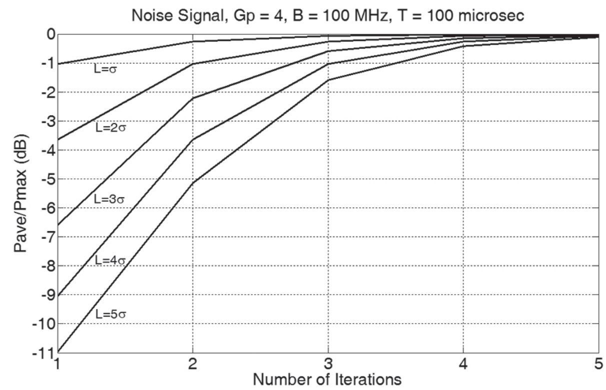

Fig. 10. Mean Envelope to Peak Power Ratio versus the number of iterations, varying the threshold $L = k\sigma$

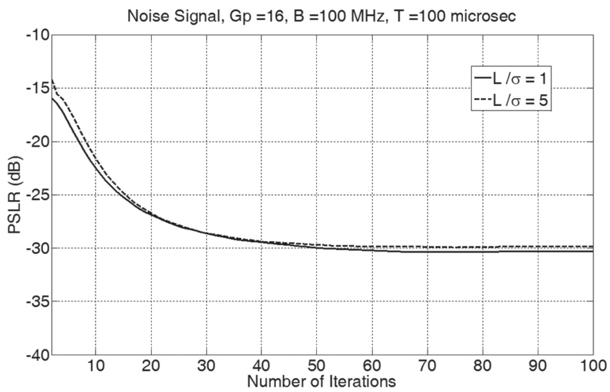

Fig. 11. PSLR versus the number of iterations considering a soft limiter. Two examples of convergence for $L = \sigma$ and $L = 5\sigma$

3.3. Closed-Form algorithm to generate Phase Noise Signals

A mathematical generation approach is based on the following considerations. For a real Gaussian process Van Vleck and Middleton [14] have shown that the autocorrelation coefficient (R_t with $t = t_2 - t_1$) of the output from a *hard limiter* is related with the input autocorrelation coefficient (here denoted r) by the well known *arcsine-law*:

$$R_t = \frac{2}{\pi}\arcsin(r)\,. \qquad (18)$$

Considering a complex Gaussian process, the correlation R_t after the *hard limiter* i.e. between: $\frac{z^*_1}{|z_1|} = \frac{x_1 - jy_1}{\sqrt{x_1^2 + y_1^2}}$ and $\frac{z_2}{|z_2|} = \frac{x_2 + jy_2}{\sqrt{x_2^2 + y_2^2}}$ is

$$R_t = E\left\{\frac{z_1^* z_2}{|z_1||z_2|}\right\} = \\ = E\left\{\frac{x_1x_2 + y_1y_2 + j(x_1y_2 + x_2y_1)}{\sqrt{x_1^2 + y_1^2}\sqrt{x_2^2 + y_2^2}}\right\} = u + jv \qquad (19)$$

where $E\{\cdot\}$ is the statistical *mean* operator. Supposing a symmetrical power density spectrum with respect to the origin, the correlation is real and $v = 0$. Equation (19) has been evaluated in [17] and it results:

$$R_t = b_0 r + \sum_{n=1}^{\infty} b_n \cdot r^{2n+1} \qquad (20)$$

being:

$$b_n = \frac{(2n-1)^2}{4n(n+1)} \cdot b_{n-1}, \quad b_0 = \frac{2}{\pi} \quad n = 1,2,3, \qquad (21)$$

Then R_t can be expressed as a sum of odd powers of r , where the coefficients b_n are very similar to those evaluated for the *arcsine-law*:

$$b_n = \frac{(2n-1)^2}{2n(2n+1)} \cdot b_{n-1}, \quad b_0 = \frac{2}{\pi} \quad n = 1,2,3, \qquad (22)$$

Fig.12 shows R_t versus the input correlation *r* for real and complex Gaussian process.

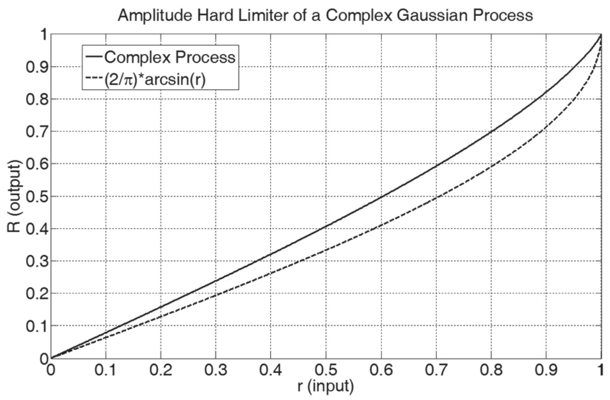

Fig. 12. Output autocorrelation (*R*) from a hard limiter versus the input autocorrelation (*r*)

Inverting eq. (20) it is possible to pre-distort the input autocorrelation to the hard limiter to obtain a desired R_t .

In such a way the requirements listed at the beginning of section 3 can be met with no need for iterations.

In fact, (*a*) the output autocorrelation is chosen in order to satisfy the PSLR requirement and the spectral band requirement, (*b*) the MEPPR requirement is satisfied by a suitable choice of the parameter *k* (ref. Fig. 9) of the limiter (the hard limiter being the situation $k \to 0$) and (*c*) the orthogonality is obtained by the randomness of the white Gaussian input sequence, and may be enhanced by proper choices of the generated output sequences.

3.4. Comparison of the algorithms

A quantitative comparison of the generation methodologies is ongoing; the qualitative comparison is shown in the following Table 1.

Table 1

Comparison of the algorithms (TBC = To Be Checked)

Quality	Algorithms to generate Phase Noise Signals		
	Axelsson [13]	Iterative	Closed Form
PSLR	good only for $\sigma < 2rad$	$< -30dB$	TBC
Orthogonality	TBC	$< -30dB$	TBC
Band occupancy	non controlled	controlled	controlled
MEPPR	1	~ 1	~1

4. CONCLUSIONS

Coherent MIMO radars call for the design of sets of orthogonal waveforms with (*i*) large enough Peak-to-Side-Lobe-Ratio of the autocorrelation function, (*ii*) fairly good mean power to peak power ratio and (*iii*) an assigned spectral occupancy.

Having shown that there are conceptual drawbacks in the Axelsson's method [13], we have started investigating two novel methods, i.e. a recursive and non-recursive (closed form) one. Preliminary results have been presented, but they are not sufficient, at the moment, to define which one is preferable.

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Gaspare Galati was with the company Selenia from 1970 till 1986, where he was involved in radar systems analysis and design. From March 1986 he has been associate professor (and full professor from November 1994) of radar theory and techniques at the Tor Vergata University of Rome, where he also teaches probability, statistics, and random processes. He is senior member of the IEEE, member of the IEE, and member of the AICT; within the AICT he is the chairman of the Remote Sensing, Navigation and Surveillance Group. He is the chairman of the SP and AES chapter of the IEEE. His main scientific interests are in radar, surveillance, navigation, and ATC/ASMGCS.

Gabriele Pavan received the Electronics Engineering Degree (Laurea) from the Tor Vergata University of Rome in 1993. After the Laurea he continued research on radarmeteorology. In 2001 he received a Ph.D. in Environmental Engineering. From 2007 he has been a researcher at the Tor Vergata University,where he currently teaches Probability Theory and Signal Processing. His focus of research is on the observation of atmospheric phenomena by radar systems.

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Технология Multiple-Input-Multiple-Output (MIMO) является новым направлением в радиолокации, которое имеет значительный потенциал для улучшения характеристик современных радиолокационных систем. В отличие от обычных радиолокаторов с фазированной антенной решеткой, система MIMO может излучать через свои антенны несколько сигналов, которые могут быть коррелированы или не коррелированы друг с другом. Для лучшего разделения при приеме сигналы обязательно должны обладать свойством ортогональности. Хотя ортогональность может проявляться во временной, частотной областях или при кодировании сигналов, для предотвращения изменений в поперечнике рассеяния цели и нежелательных эффектов Доплера, сигналы должны передаваться одновременно и на одной и той же несущей частоте. Как следствие, ортогональность в сигнальной области наиболее предпочтительна, и для успешного использования такой системы ключевую роль играет качество построения сигналов. Хорошие результаты для радиолокационных систем с технологией MIMO показывают ортогональные сигналы со случайной фазой. В статье, после введения в MIMO технологии радиолокационных систем, представлено статистическое описание основных характеристик таких сигналов, в том числе анализ автокорреляционных, кросс-корреляционных и спектральных свойств. Наконец, предложено два новых метода для генерации сигналов со случайной фазой: рекурсивный и нерекурсивный (закрытая форма). Предварительные результаты будут представлены.

Ключевые слова: MIMO, ортогональные сигналы, фазовый шум.

Табл. 1. Ил. 12. Библиогр.: 17 назв.

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Побудова радіолокаційних сигналів для шумового MIMO радара / Г. Галаті, Г. Паван // Прикладна радіоелектроніка: наук.-техн. журнал. – 2013. – Том 12. – № 1. – С. 3-10.

Технологія Multiple-Input-Multiple-Output (MIMO) є новим напрямком у радіолокації, який має значний потенціал для поліпшення характеристик сучасних радіолокаційних систем. На відміну від звичайних радіолокаторів з фазованими антенними решітками, система MIMO може випромінювати через свої антени кілька сигналів, які можуть бути корельовані або некорельовані один з одним. Для кращого поділу при прийомі сигнали обов'язково повинні мати властивість ортогональності. Хоча ортогональність може проявлятися в тимчасовій, частотної областях або при кодуванні сигналів, для запобігання змін у поперечнику розсіювання цілі і небажаних ефектів Доплера, сигнали повинні передаватися одночасно і на одній і тій самій несучій частоті. Як наслідок, ортогональність у сигнальній області найкраща, і для успішного використання такої системи ключову роль відіграє якість побудови сигналів. Хороші результати для радіолокаційних систем з технологією MIMO показують ортогональні сигнали з випадковою фазою. У статті, після введення в MIMO технології радіолокаційних систем, подається статистичний опис основних характеристик таких сигналів, в тому числі аналіз автокореляційних, крос-кореляційних і спектральних властивостей. Нарешті, запропоновано два нових методи для генерації сигналів з випадковою фазою: рекурсивний і нерекурсивний (закрита форма). Попередні результати будуть представлені.

Ключові слова: MIMO, ортогональні сигнали, фазовий шум.

Табл. 1. Іл. 12. Бібліогр.: 17 найм.