Low-Rate Sampling Technique for Range-Windowed Radar/Sonar Using Nonlinear Frequency Modulation

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This paper proposes a new radar/sonar processing architecture that enables a low-rate time-sampling rate while still producing a high-resolution range profile over a narrow range-window. This new architecture generalizes the conventional “stretch processing” architecture (which employs linear frequency modulated (LFM) waveforms that suffer from high range-sidelobes), to nonlinear frequency modulated (NLFM) waveforms, which can lower the range-sidelobes without tapering. Computational results demonstrate both the estimation efficacy and the time-sampling efficiency of the proposed scheme.

I. INTRODUCTION

A radar/sonar system produces a range profile of the targets based on echoes received from its transmitted waveform. The radar/sonar receiver architecture generally consists of a matched filter whose impulse response is the conjugated time reversal of the transmitted waveform. This matched filter correlates the received echoes with variably delayed replicas of the transmitted waveform, in an operation known as pulse compression. This process results in a target range profile, wherein those time delays/range gates with energy levels above a threshold are declared to contain targets.

To attain fine resolution along the delay (i.e. range) coordinate, the radar/sonar system needs to transmit a broadband waveform [1], which in turn necessitates a high rate of time-sampling, in accordance with the Nyquist theorem. This requires expensive electronics for time-sampling, and generates an excessive amount of data.

The matched filter’s target-range profile spans range gates that may not all be of interest. The demanding time-sampling requirements for a broadband waveform may be reduced through the use of “stretch processing” [2, 4, 7, 9, 10, 14, 15, 18, 19], if prior information is available concerning the range-window in which targets of interest may exist. Stretch processing allows a low time-sampling rate in a homodyne architecture,1 but still achieves fine range-resolution for linear frequency modulations. Stretch processing converts the echo returns of a transmitted linear-frequency-modulated (LFM) signal (a.k.a. “chirp”, “linear chirp”, “sweep signal”) to a temporally windowed frequency-domain impulse which may be time-sampled at a rate much below that of the original broadband transmitted signal. To achieve this, the receiver mixes (i.e., multiplies for each time instant) the echo returns with another LFM waveform, so as to “de-ramp” the echo returns, thereby outputting one windowed frequency-domain impulse for each echo-return at a frequency corresponding to the range of the target. It should be noted that the term “stretch processing” may refer to any technique that converts delays from a small range to signals with a longer time duration. Additionally, there are other uses of the term that exist in the literature, see e.g. [11]. However, in this paper, stretch processing should be understood to refer explicitly to the common case in which LFM waveforms are used as described above.

In [18], it is analytically proven that LFM is the only polynomial phase modulation that will yield a windowed frequency-domain impulse after deramping. This constitutes a notable disadvantage, because LFM suffers from high sidelobes along the delay/range coordinate, due to the LFM power-spectrum’s nearly rectangular shape. Such high sidelobes could lead to the mis-identification of

1 Homodyne processing (a.k.a. synchrodyne, zero-IF receiver, direct-conversion receiver) suits applications with approximately prior known target range and velocity.
spurious targets and/or could obscure a weak target-return in the presence of a strong target-return. These sidelobes could be lowered by tapering, but only at the cost of increasing the mainlobe width and thus reducing the range resolution.

In contrast, nonlinear frequency modulated (NLFM) waveforms are recognized for their flexibility in power-spectrum shaping for low sidelobes [5], and do not require tapering, which would compromise range-resolution. Such NLFM waveforms have been investigated in the open literature (e.g. [3, 6, 8, 12, 13, 16, 17]) for the case of conventional matched filtering.

For such broadband NLFM waveforms, this paper proposes a new low-rate time-sampling homodyne radar/sonar architecture that generalizes conventional LFM waveform based stretch processing. This presently proposed processing architecture allows not only low-rate time-sampling, but also achieves fine resolution along the delay/range coordinate without tapering.

The rest of this paper is organized as follows. Section II formulates the problem and reviews the conventional architecture of stretch processing. Section III proposes the new homodyne architecture for radar/sonar processing, using NLFM. Section IV presents numerical examples to verify the efficacy of the proposed scheme and to characterize its performance.

II. TECHNICAL PRELIMINARIES

A. Radar/Sonar Data Model

Consider an arbitrary NLFM signal $s(t)$, transmitted for a duration of $T$:

$$s (t) = e^{+j2\pi \phi (t)} [u(t) - u(t - T)],$$  \hspace{1cm} (1)

where $\phi (t)$ denotes a time-dependent phase function, and $u(\cdot)$ represents the Heaviside step function.

The radar/sonar receiver’s data measurement consists of $K$ target echoes, which may be modeled as

$$r (t) = \sum_{k=1}^{K} \alpha_k s (t - \tau_k),$$  \hspace{1cm} (2)

where $\alpha_k$ denotes the amplitude of the $k$th echo, and $\tau_k$ symbolizes the corresponding time delay. The received signal $r(t)$ thus consists of scaled and delayed copies of $s(t)$.

Suppose that the range of interest corresponds to a delay window of $[\tau_1, \tau_K]$, wherein $\tau_1 < \tau_2 < \ldots < \tau_K$, without loss of generality. The goal is to estimate the echoes’ relative amplitudes and delays $\{\alpha_k, \tau_k, k = 1, \ldots, K\}$ from $r(t)$.

B. Review of Stretch Processing [2]

For conventional stretch processing, the transmitted radar/sonar signal must be a chirp (quadratic phase) waveform $s_{\text{LFM}}(t)$, with $\phi (t) = f_a t + \frac{\pi}{2} t^2$, where $\kappa = B/T$ is known as the “chirp rate”, sweeping a bandwidth of $B$ over a period of $T$. The above $s_{\text{LFM}}(t)$ has an instantaneous frequency of $f_a$ at time $t = 0$.

At the radar/sonar receiver, the received signal $r(t)$ is mixed with another chirp waveform $d_{\text{LFM}}(t)$, generated locally by the radar/sonar receiver itself, in order to de-ramp $r(t)$. Drawing upon the receiver’s prior knowledge that all targets should lie within a time delay window of $[\tau_1, \tau_K]$, the waveform $d_{\text{LFM}}(t)$ must start by $\tau_1$ and must end after $\tau_K + T$, in order for the receiver to accommodate the echoes from all targets. To ensure the capture of the entire range-window of interest, small guard intervals are added, such that $d_{\text{LFM}}(t)$ would span over $\tau_1$ (slightly before $\tau_1$) and $\tau_K$ (slightly after $\tau_K + T$), as shown in Fig. 1. Hence,

$$d_{\text{LFM}}(t) = e^{j2\pi \left[ f_a (t - \tau_1) + \frac{\pi}{2} (t - \tau_1)^2 \right]} \left[ u(t - \tau_1) - u(t - \tau_1) \right],$$  \hspace{1cm} (3)

Mixing the received signal $r(t)$ with the conjugate of the above deramping signal $d_{\text{LFM}}(t)$ results in the following:

$$p_{\text{LFM}}(t) = r(t) d_{\text{LFM}}^*(t) = \sum_{k=1}^{K} \alpha_k e^{j2\pi \left[ f_a (t - \tau_k - \tau_1) + \frac{\pi}{2} (t - \tau_k + T) \right]} \times e^{j2\pi \left[ f_a (t - \tau_k) + (\tau_k - \tau_1) - u(t - \tau_k - \tau_1) \right]} \times [u(t - \tau_k) - u(t - \tau_k - \tau_1)].$$  \hspace{1cm} (4)

In (4), a time-independent phase term has been absorbed into $\alpha_k$.

Stretch processing thus converts a target with strength $\alpha_k$ and delay $\tau_k$ into a windowed tone-signal of duration $T$, strength $\alpha_k$, and frequency $f_k$ that is directly related to the target range/delay $\tau_k$. The numerical values of $\{f_k, k = 1, \ldots, K\}$ may be evaluated by locating the spectral peaks of the Fourier transform $P_{\text{LFM}}(\tau)$ of $p_{\text{LFM}}(t)$.

Stretch processing can potentially reduce the radar/sonar receiver’s time-sampling requirement by orders of magnitude. For example, sub-meter-scale range resolution requires a waveform with bandwidth $B$ of hundreds of MHz, thereby necessitating time-sampling rates at or above the GHz range if matched filtering is used. In contrast, stretch processing for a relatively short range-window (and hence a narrow band of frequencies $\alpha |f_1 - f_k|$, e.g. for target identification applications) can have $P_{\text{LFM}}(\tau)$ spanning over a narrow band of only several tens of MHz, thereby significantly reducing the time-sampling rate. This concept is illustrated in Fig. 1.

LFM waveforms (and thus stretch processing), however, suffer from relatively high range-sidelobes. In contrast, NLFM waveforms can intrinsically offer lower range-sidelobes even without tapering. The next section
develops a low-rate time time-sampling homodyne architecture for the radar/sonar receiver to form a range-profile from target echoes of an NLFM transmitted waveform.

III. PROPOSED NLFM HOMODYNE PROCESSING

A. Development of the Proposed Processing Architecture

Define a "down-conversion" waveform,

\[ d(t) = e^{+j2\pi \phi_s(t)} [u(t - \tau_e) - u(t - \tau_r)]. \] (5)

that corresponds to the transmitted NLFM waveform \( s(t) \) of (1), where \( \phi_s(t) \) represents some phase function that may (or may not) be quadratic.

The receiver then mixes the received signal \( r(t) \) with \( d(t) \) to yield

\[ p(t) = r(t) d^*(t) \]

\[ = \sum_{k=1}^{K} \alpha_k \exp \left\{ j2\pi \left[ \phi_s(t - \tau_k) - \phi_s(t) \right] \right\} \]
\[ \times [u(t - \tau_k) - u(t - \tau_k - T)]. \] (6)

This mixing operation is mathematically equivalent to subtracting \( \phi_s(t) \) from the phase \( \phi_s(t - \tau_k) \) of the \( k \)th target return \( \forall k \). Because \( \phi_s(t) \) is generally a smooth function, \( \phi_p(t; \tau_k) = \phi_s(t - \tau_k) - \phi_s(t) \) undergoes only slow variability over time. Therefore, \( p(t) \) in (6) will generally be narrower in bandwidth than \( s(t) \) in (1), for a sufficiently narrow range-window. This is critical, because this bandwidth reduction [from \( r(t) \) to \( p(t) \)] enables the lowering of the required time-sampling rate. The proposed scheme requires a time-sampling rate proportional to only the maximum instantaneous frequency \( d \phi_p(t; \tau) / dt \), which roughly approximates the bandwidth resulting from a particular target's relative delay. Because this bandwidth increases with the target delay, the proposed scheme's bandwidth is approximately

\[ B_p = \max_{t \in [\tau_e, \tau_k + T]} \left| \frac{d\phi_p(t; \tau_k)}{dt} \right| \] (7)

In contrast, customary matched filtering requires a time-sampling rate proportional to \( B \). However, since \( B \gg B_p \) for a limited range-window, the proposed architecture can significantly lower the required time-sampling rate.

In order to estimate the parameters of interest \( \{\alpha_k, \tau_k, k = 1, \ldots, K\} \) from the low-bandwidth signal \( p(t) \) which represents the received signal after mixing with \( d(t) \), the signal \( p(t) \) is projected onto

\[ \psi(t; \tau) = e^{+j2\pi \phi_p(t; \tau)} \] (8)

where \( \psi(t; \tau) \) may be viewed as a canonical received signal after the aforementioned mixing. Such a projection will produce a range/delay profile \( P(\tau) \), the peaks of which provide an estimate of \( \{\alpha_k, \tau_k, k = 1, \ldots, K\} \).

Mathematically,

\[ P(\tau) = |\langle p(t), \psi(t; \tau) \rangle| \]
\[ = \left| \int_{\tau_e}^{\tau_k} p(t)e^{-j2\pi \phi_p(t; \tau)} dt \right| \] (9)
\[ \approx \left| \int_{\tau_e}^{\tau_k} \sum_{k=1}^{K} \alpha_k e^{+j2\pi \phi_p(t; \tau_k)} [u(t - \tau_k) - u(t - \tau_k - T)] \times e^{-j2\pi \phi_p(t; \tau)} dt \right| \]
The expression \( \mathbf{R}_s(\tau) \) represents the \( \tau_\ell \)-shifted autocorrelation \( \mathbf{R}_s(\tau) \) of the emitted NLFM waveform \( s(t) \). Thus, for the prespecified range-window \( [\tau_i,\tau_f] \) (wherein the targets are a priori known to lie), the above proposed algorithm gives a range/delay profile \( P(\tau) \) over the desired range-window that is in fact equivalent to that from conventional matched filtering. This process may be performed at the lower time-sampling rate allowed for by the relatively narrowband \( p(t) \). This proposed approach contrasts with conventional matched filtering, which projects the received signal \( r(t) \) onto \( s(t) \), thus requiring the signal processing to be performed at the higher time-sampling rate required by the relatively wideband \( s(t) \) and \( r(t) \).

From a practical standpoint, after \( p(t) \) is time-sampled at the receiver (for which it should again be emphasized that the required time-sampling rate is much lower than matched filtering since \( B > B_p \) the evaluation of (9) may be approximated numerically as

\[
P(\tau) \approx T_s \sum_{m=m_0}^{m_p} P(mT_s) e^{-j2\pi \phi_p(mT_s)}
\]

where \( m = m/T_s \), \( 1/T_s \), and \( M = m_0 - m_p \). If \( M = m_c - m_t \), then selecting for \( \tau \) a granularity of \( T_s \) results in a computational complexity of \( O(M^2) \). It should be noted that while this is a relatively high computational complexity, it is in fact exactly the same as stretch processing under the discrete Fourier transform (DFT) implementation. However, since the DFT is more efficiently implemented by the fast Fourier transform (FFT), this cost can be reduced to \( O(M \log M) \) in the case of stretch processing. The development of an efficient means to compute (12), however, is beyond the scope of this paper and is not discussed further.

In summary, the proposed sub-Nyquist homodyne architecture involves the following algorithmic steps.

1) Transmit an NLFM waveform \( s(t) \).
2) Mix the received signal \( r(t) \) with \( d(t) \) as in (5), to generate \( p(t) \).
3) Time-sample \( p(t) \) at a low rate (around \( 2B_p \)).

4) Compute (12) to obtain the range profile \( P(\tau) \).
5) Identify from \( P(\tau) \) the target range/delay \( \tau_\ell \) and the relative amplitude \( a_\ell, \forall \ell \).

B. Relation to Stretch Processing

Consider the scenario when \( s(t) \) is simply a standard LFM waveform, so that \( \psi(t) = \psi_{\text{LFM}}(t) = f_\alpha t + \frac{1}{2} t^2 \). Consequently,

\[
\psi(t;\tau) = e^{j2\pi \left[ f_\alpha (t-\tau) + \frac{1}{4} (t-\tau)^2 - (f_\alpha t + \frac{1}{2} t^2) \right]}
\]

and

\[
P(\tau) = \left| \langle p(t), \psi(t;\tau) \rangle \right| = \int_{t_i}^{t_f} p(t) e^{-j2\pi k\tau} dt = e^{j\left( \frac{1}{2} t^2 - f_\alpha \tau \right)} \int_{t_i}^{t_f} p(t) e^{-j2\pi k\tau} dt = \int_{t_i}^{t_f} p(t) e^{j\frac{\pi}{2} k\tau} dt.
\]

It may be seen from (13) that the projection degenerates to the customary Fourier transform. Hence, (9) would yield a result equivalent to conventional stretch processing.

The above development shows that stretch processing is a special case of the proposed processing architecture. Thus, stretch processing hardware (such as that described in [19]) can easily accommodate the proposed processing architecture by simply replacing the LFM waveform with the desired NLFM waveform.

C. Performance in the Presence of Additive Noise

The signal-to-noise ratio (SNR) for the proposed architecture is now considered. For a single target echo received in the presence of additive noise, the received signal may be modeled as

\[
r(t) = \alpha_1 s(t - \tau_1) + n(t),
\]

where \( n(t) \) denotes a zero-mean additive white noise process, with variance \( N_0/2 \). Therefore,

\[
p(t) = \left[ \alpha_1 s(t - \tau_1) + n(t) \right] e^{-j2\pi \phi_p(t)}
\]

\[
\times \left[ u(t - \tau_1) - u(t - \tau_1) \right],
\]

which may be substituted into (9) to yield

\[
P(\tau) = \int_{t_i}^{t_f+T} \alpha_1 s(t - \tau_1) s^*(t - \tau) dt + \int_{t_i}^{t_f+T} n(t) e^{-j2\pi \phi_p(t;\tau)} dt = \alpha_1 R_s(\tau - \tau_1) + \int_{t_i}^{t_f+T} n(t) e^{-j2\pi \phi_p(t;\tau)} dt.
\]
The signal power peaks at \( \tau = \tau_1 \), resulting in a value of
\[
|S(0)|^2 = |\alpha_t R_s(0)|^2 = \alpha_t^2 T^2 = TE,
\]
where \( E \) denotes the energy received from the single target-echo, that is,
\[
E = \int_{\tau_1}^{\tau_1+T} |\alpha_k s(t - \tau_1)|^2 dt = \alpha_k^2 T = TE.
\]

The noise variance equals
\[
E \{N(\tau)^2\} = \int_{\tau_1}^{\tau_1+T} \int_{\tau_1}^{\tau_1+T} E \{ n(t) n^*(\xi) \} e^{j2\pi[\phi_s(t) - \phi_s(t - \tau)]} dt d\xi = \int_{\tau_1}^{\tau_1+T} \frac{N_0}{2} dt = \frac{N_0 T}{2}.
\]

Putting the above expressions together, the output SNR equals
\[
\text{SNR} = \frac{|S(0)|^2}{E \{N(\tau)^2\}} = \frac{2E}{N_0}.
\]
This above output SNR is identical to the matched filter bound, and thus retains the SNR performance of the conventional matched filter implementation.\(^2\)

IV. NUMERICAL EXAMPLES

This section presents sample computational results to illustrate the performance of the proposed processing architecture.

Recall that the proposed architecture allows usage of any NLFM waveform, in contrast to stretch processing which is restricted to an LFM waveform. Consider the following highly nonlinear phase function:
\[
\phi_s(t) = \frac{\pi B}{T} t^2 - \pi BT \sum_{m=1}^{M} \frac{K_m}{\pi m} \cos \left(2\pi \frac{m}{T} t\right)
\]
where \( K_m \) represents a set of constant values [6]. The delay profile of the corresponding NLFM waveform with phase function as in (15) is compared in Fig. 2 against that from a standard LFM waveform for \( B = 100 \text{ MHz} \) and \( T = 50 \mu s \). While both delay-profiles peak at the same delay/range, the NLFM waveform offers a –40 dB sidelobe, whereas the LFM waveform exhibits significantly higher sidelobes at –13 dB. The lower
\(^2\) A similar observation is made in [2] that conventional stretch processing also retains the SNR performance of the matched filter.

sidelobe levels of the NLFM waveforms highlights one advantage over conventional LFM waveforms.

Suppose that the NLFM waveform whose range profile is shown in Fig. 2 is used as the transmitted waveform to interrogate an environment that consists of two equal-strength scatterers at relative delays of \( \tau_1 = 3 \mu s \) and \( \tau_2 = 3.025 \mu s \), so that their spacing is 25 ns (3.75 m) apart. It should be noted that the reference points \( \tau_1 \) and \( \tau_2 \) are certainly important from a practical standpoint, since they govern the captured range-window. However, from a simulation perspective, their values need only to match the desired range-window, which in this case captures the two closely spaced scatterers. Hence, their values can be selected so that \( \tau_1 = \tau_1 \) and \( \tau_e = \tau_2 + T \).

Fig. 3 shows the time-frequency plane representation of \( r(t) \) and \( p(t) \) for the first scatterer, which are denoted, respectively, as \( r_1(t) \) and \( p_1(t) \). Note that \( r(t) \) and \( p(t) \) for the second scatterer are not shown for space considerations, since they are very similar to \( r_1(t) \) and \( p_1(t) \). For reference, \( r_{1,\text{LFM}}(t) \) and \( p_{1,\text{LFM}}(t) \), which correspond to \( r(t) \) and \( p(t) \) for the first scatterer when an LFM waveform (conventional stretch processing) is used, are also shown. Recall that \( p(t) \) results from mixing the measurement \( r(t) \) with the down-conversion waveform \( d(t) \). Fig. 3 shows that \( p(t) \) spans a bandwidth of only \( B_p = \ldots \)
20 MHz, even though \( r(t) \) spans the transmitted waveform’s entire bandwidth of \( B = 100 \) MHz. This verifies the significant reduction of the time-sampling rate by the proposed scheme.

For the same scenario as in Fig. 3, Figs. 4–6 show the target delay-profile of the two closely spaced targets using the proposed method (9), using matched filtering, and using stretch processing, at various time-sampling rates. In the following discussion, the term “Nyquist rate” is used to refer to the full sampling rate based on the bandwidth \( B = 100 \) MHz of the transmitted waveform \( s(t) \).

In Fig. 4, Nyquist rate sampling is used, and it can be seen that the proposed method and matched filtering yield very similar target profiles. At half of the Nyquist rate, Fig. 5 shows that the proposed method retains its Nyquist rate delay-profile, but the matched filter delay-profile is severely degraded. At one-fifth of the Nyquist rate, Fig. 6 shows that the proposed method’s target delay-profile remains unperturbed, but the matched filter delay-profile degrades still further. The trend observed in Figs. 4–6 is due to the fact that the proposed scheme’s \( p(t) \) has a bandwidth significantly less than that of \( r(t) \) [and hence that of \( s(t) \)]; the proposed scheme can thus tolerate significantly sub-Nyquist rate sampling, without introducing any distortion/aliasing, even as the matched filter sub-Nyquist rate delay-profiles start to exhibit elevated sidelobes and great loss of target resolution. For reference, Figs. 4–6 also show the delay-profiles for stretch processing. Like the proposed method, it can be seen from Figs. 4–6 that the delay-profile of stretch processing also remains invariant across sampling rates due to the deramping operation described in Section II-B. However, because stretch processing is restricted to LFM waveforms, the sidelobe levels exhibited in Figs. 4–6 are very high. Thus, the proposed method can offer significant savings in time-sampling requirements, while simultaneously lowering sidelobes without tapering, thereby preserving the mainlobe’s width.

The time-sampling reduction may be quantified via (7), for any \( \phi_s(t) \). Fig. 7 shows (7) for the \( \phi_s(t) \) of (15). Beyond a relative delay of roughly 12 \( \mu s \), the bandwidth \( B_p \) of \( p(t) \) exceeds the bandwidth \( B = 100 \) MHz that characterizes \( s(t) \) and \( r(t) \), thereby indicating that matched filtering would be sufficient. Below this relative delay of roughly 12 \( \mu s \), however, \( p(t) \) would have a bandwidth under that of \( s(t) \) and \( r(t) \), thereby allowing a time-sampling rate below that of the matched filtering. As such, the proposed scheme’s maximum range window is around 2 km in this example.
V. CONCLUSION

The conventional homodyne radar/sonar architecture of stretch processing reduces the time-sampling rate required by conventional matched filtering. However, stretch processing requires the usage of LFM waveforms, which suffer from high range-sidelobes. The presently proposed radar/sonar architecture generalizes conventional stretch processing to exploit the much wider class of NLFM waveforms, which offer lower range-sidelobes. This new architecture significantly reduces the time-sampling rate, while lowering the range-sidelobes without the need for tapering. A numerical example was given to show that if an NLFM waveform with a bandwidth of 100 MHz is used to illuminate two targets spaced around 4 m apart, the proposed architecture lowers the required sampling rate by a factor of five compared with the transmitted waveform’s Nyquist rate. Moreover, the range-sidelobes are reduced from the usual −13 dB for an LFM waveform to −40 dB for the NLFM waveform.

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