A Nonlinear Suppression Technique for Range Ambiguity Resolution in Pulse Doppler Radars

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Abstract — This paper reintroduces a novel range ambiguity resolution technique, invented by Carmen Palermo and first reported in 1962, called *nonlinear suppression*. A set of M unique modulations is selected which have desirable (mutually dispersive) cross-correlation properties. The transmitted pulse train is constructed by successively modulating each pulse with one of the M modulations. The received data are processed in M independent receiver channels with each channel corresponding to a single modulation. Ideally, the channels are designed such that the output response is only dependent upon one of the modulations (versus all M). This is accomplished to a useful extent with Palermo's technique by progressively removing unwanted contributions on a range cell by range cell basis. By passing the received data through an appropriate memoryless nonlinearity, the process preferentially removes energy from data responsive to each of the other $M-1$ codes.

I. INTRODUCTION

Pulse repetition frequency (PRF) selection for airborne pulse Doppler radar applications is normally based on the need to separate targets of interest from strong clutter returns. Since aircraft motion spreads the clutter in frequency, medium-to-high PRFs are chosen to ensure extended clutter-free zones exist in the Doppler spectrum. As a result, range ambiguity and clutter fold-over are introduced.

Figure 1 illustrates the effect of a uniform pulse train operating under range ambiguous conditions where the unambiguous range R_u is dependent upon the pulse repetition interval (PRI) T_r and may be expressed as:

$$R_u = \frac{cT_r}{2} \quad (1)$$

The true range profile of Figure 1 includes three targets at ranges of $0.5 R_u$, $2.25 R_u$, and $3.75 R_u$. Under range ambiguous conditions, the uniform pulse train produces apparent range results at the detector output of $0.5 R_u$, $0.25 R_u$, and $0.75 R_u$, respectively. In this case, the radar receiver is not able to unambiguously resolve the target locations and determine the true range of targets 2 and 3 — under normal operating conditions the receiver would simply declare three targets present at the apparent ranges and not actually know that an ambiguous condition exists. The typical technique for resolving range ambiguity is to utilize multiple PRFs. Many recent papers have attempted to

Figure 1. True Range and Apparent Range

optimize PRF selection and ambiguity resolution using several PRFs [1, 2].

The effect of range ambiguity on sidelobe and mainlobe clutter is often severe. Mainlobe clutter, often at ranges near the radar horizon, folds into the near range interval, while targets at ranges greater than R_u are buried in near-range sidelobe clutter. For targets with Doppler frequencies outside the clutter spectrum, such as high-speed targets moving toward the radar, Doppler filtering is normally sufficient to suppress the clutter (although the target range remains ambiguous). However, targets with low relative radial velocity, e.g., those encountered in a tail-chase scenario, possess Doppler frequencies falling within the sidelobe clutter spectrum and may be completely undetectable [3].

This paper reintroduces a novel range ambiguity resolution technique, invented by Carmen Palermo in 1962, called "nonlinear suppression" (NLS). Palermo, Leith, and Horgen first reported NLS in 1962, but the available technology at the time prevented further development [4]. A set of M unique pulse codes is selected based on orthogonality criteria or acceptable cross-correlation properties. Each radar pulse is coded with one of M pulse codes and transmitted. After M coded pulses have been transmitted, the process is repeated and another coded pulse train is transmitted. The received target return is simultaneously processed in M independent

Figure 2. Coded Pulse Train Using $M = 4$ Pulse Codes

channels with each channel corresponding to a distinct pulse code. Each channel is designed to suppress returns from all pulses except those containing the code corresponding to that channel. Ambiguity suppression is accomplished through a combination of matched filtering and nonlinear “hole-punching” designed to remove undesired, compressed pulse returns. The channel outputs are subsequently combined, providing the detector with an input signal having ambiguities reduced by a factor of M .

II. NONLINEAR SUPPRESSION EXAMPLE

To illustrate the concept of nonlinear suppression, consider the case with $M = 4$ and assume binary phase-shift keyed (PSK) modulation – any pulse compression method may be used provided the codes have desirable cross-correlation properties. Each code may be represented as a sequence of ± 1 's, with each element representing a “chip” of duration T_c , the compressed pulse length. Codes are then specified by:

$$\begin{aligned} p_1[n] &= \pm 1, n = 0, 1, \dots, \text{PCR}-1 \\ p_2[n] &= \pm 1, n = 0, 1, \dots, \text{PCR}-1 \\ p_3[n] &= \pm 1, n = 0, 1, \dots, \text{PCR}-1 \\ p_4[n] &= \pm 1, n = 0, 1, \dots, \text{PCR}-1 \end{aligned} \quad (2)$$

where PCR is the *pulse compression ratio* which is a function of the pulse duration T_p and given by:

$$\text{PCR} = \frac{T_p}{T_c} \quad (3)$$

For this example, the coded transmit pulse sequence is $p_1, p_2, p_3, p_4, p_1, p_2, p_3, p_4, p_1, \dots$, etc. For a coherent

Figure 3. Four Channel NLS Receiver Architecture

Figure 4. Nonlinear “Hole Punch” Function

processing interval (CPI) consisting of N pulses with N/M an integer value, there are N/M “bursts” consisting of M pulses each, as shown in Figure 2.

As shown in Figure 3, the NLS receiver processing for this example consists of using four channels with each channel specifically designed to suppress all but one distinct pulse code. Letting $z[n]$ denote the discrete complex baseband waveform obtained from the IF processor, each NLS channel accepts $z[n]$ as an input and produces complex output $y_m[n]$ – subscript m denotes the particular pulse code the channel is designed for. Each channel output is then formatted and combined with the other channel outputs to form the NLS processor output $y[n]$.

The processing flow of a single NLS receiver channel, Channel 1, is shown in Figure 5 and proceeds as follows. First, the complex input data is filtered with h_4 , a filter “matched” to coded pulse p_4 :

Figure 5. NLS Channel 1

$$h_4[n] = p_4[n_{\max} - n] \quad (4)$$

The value of $n_{\max}$ depends on the length of $z[n]$. A convenient processing interval is one PRI, which results in $n_{\max}$ equaling T_r/T_s , where T_s is the sample time. The convolution of $z[n]$ with $h_4[n]$ effectively compresses portions of the return containing p_4 pulses while simultaneously "spreading" other data further in time, thus reducing their amplitude.

A nonlinearity ("hole puncher") is next introduced to eliminate the compressed data response. The specific nonlinear function used for this work is shown in Eq (5).

$$\Gamma_{\alpha}(\sigma) = \begin{cases} 0 & |\sigma| > \alpha \\ \sigma & |\sigma| \leq \alpha \end{cases} \quad (5)$$

The function of Eq (5) is plotted in Figure 4 where α is the threshold and σ is a real number. Since radar data is generally complex, the nonlinearity is applied separately to the In-phase (I) and Quadrature (Q) data. The selection of threshold value α is discussed in Section V.

After applying the nonlinearity, the remaining data is filtered using the conjugate of h_4 , which is identically p_4 since real-valued pulse codes are used. The compression/puncturing process is then sequentially repeated for p_3 and p_2 . The final step in Channel 1 processing is to compress the p_1 pulses by filtering with h_1 . The overall NLS processing of Channel 1 may be expressed concisely as:

$$y_1[n] = h_1[n] * \Lambda_2(\Lambda_3(\Lambda_4(z[n]))) \quad (6)$$

where

$$\Lambda_m = p_m[n] * \Gamma_{\alpha_m}(h_m[n] * z[n])$$

To suppress larger amplitude responses first, some consideration must be given to the order in which the pulses are suppressed within each channel. If input data $z[n]$ only consists of data from one PRI, then the largest amplitude returns are most likely from the most recently transmitted pulse (near-range targets). The next largest amplitude returns are likely from the pulse transmitted prior to the most recent pulse, and so on. Therefore, if the most recent transmitted pulse is p_1 , it is desirable for each channel to suppress pulses in the following order:

- Channel 1: p_4, p_3, p_2
- Channel 2: p_1, p_4, p_3
- Channel 3: p_2, p_1, p_4
- Channel 4: p_3, p_2, p_1

Clearly, the order of suppression changes with each PRI.

III. TEST RESULTS

A nonlinear suppression 'proof-of-concept' test was conducted using the GP-3 radar [5], a low-power developmental radar owned by Air Force Research Laboratory (AFRL) and currently operated by the Air Force Institute of Technology (AFIT). Table 1 lists the major parameters of the GP-3 and Table 2 lists the specific NLS test parameters used.

TABLE 1
GP-3 RADAR PARAMETERS

Parameter	Value
Carrier Frequency Range	9.43 GHz to 9.93 GHz
Instantaneous Bandwidth	3.5 MHz
First IF Frequency	900 MHz
Second IF Frequency	30 MHz
Third IF Frequency	5 MHz
Sampling Frequency	20 MHz
Number of Transmit Channels	4
Number of Receive Channels	4

TABLE 2
NLS TEST PARAMETERS

Parameter	Value
Sample Time (T_s)	0.05 μ s
Chip Length (T_c)	0.8 μ s
Pulse Length (T_p)	101.6 μ s
Pulse Compression Ratio (PCR)	127
Number of Pulse Codes (M)	4
Pulse Coding	Gold Codes
Pulse Repetition Interval (PRI)	406.4 μ s
Unambiguous Range (R_u)	60.96 km
Target 1 Range	30.48 km (0.5 R_u)
Target 2 Range	137.16 km (2.25 R_u)
Target 3 Range	228.6 km (3.75 R_u)

The target ranges are the same as depicted in Figure 1. Each target return was formatted, delayed, and broadcast over one of the four separate GP-3 transmit channels at X-Band. Four 127-chip Gold code sequences comprised the pulse coding. For this test, all target returns had identical IF signal-to-noise ratios (SNR) of approximately 10 dB. The receiver output consisted of 12-bit sampled IF data centered at 5 MHz with a bandwidth of 3.5 MHz.

Figure 6 shows the result when using a single Gold code for all pulses. The dashed vertical grid lines mark the unambiguous range (R_u) for this waveform. The ambiguities appear exactly as previously illustrated in Figure 1.

Figure 7 shows the NLS results when using four distinct gold codes – the true target ranges are clearly evident. The sidelobes appearing in the NLS output are a result of the

Figure 6. Test Results for Single Gold Code

Figure 7. Test Results for NLS Using Four Gold Codes

pulse code cross-correlation, autocorrelation sidelobes, and distortion induced by the nonlinearity.

IV. CLUTTER SIMULATION RESULTS

Nonlinear suppression may provide a means of detecting weak targets buried in clutter due to range ambiguity. To investigate this possibility, a Matlab[®] simulation was conducted with parameters shown in Table 3. The clutter was generated using a backscatter coefficient generator based on the Weibull distribution [6]. The received signal plus clutter return is shown in Figure 8. The dominant peaks are the specular return from the ground directly beneath the radar. Although this simulation did not incorporate Doppler effects due to platform motion, each clutter cell was assigned a uniformly distributed random phase value between 0 and 2π . Four 127-chip maximal length sequences (m -sequences) were used for pulse coding.

The output from each NLS channel is shown in Figure 9 and the resolved unambiguous output is shown in Figure 10.

Figure 8. Received Signal Plus Clutter

Figure 9. Received Signal Plus Clutter (Simulation)

TABLE 3

CLUTTER SIMULATION TEST PARAMETERS

Parameter	Value
Sample Time (T_s)	0.8 μ s
Chip Length (T_c)	0.8 μ s
Pulse Length (T_p)	101.6 μ s
Pulse Compression Ratio (PCR)	127
Number of Pulse Codes (M)	4
Pulse Coding	Maximal Length
Pulse Repetition Interval (PRI)	406.4 μ s
Unambiguous Range (R_u)	60.96 km
Target 1 Range	45.72 km (1.5 R_u)
Target 2 Range	137.16 km (2.25 R_u)
Target 3 Range	228.6 km (3.75 R_u)
Radar Altitude	5000 km
Radar Velocity	240 m/s
Antenna Mainbeam Gain	34 dB
Antenna Sidelobe Gain	-10 dB

Data were processed over twelve PRIs with the results of Figure 10 obtained by integrating three pulses in each PRI. As indicated, Target 2, which is most affected by clutter, is

Figure 10. Unambiguous NLS Output from Clutter Simulation

clearly detectable while Target 3 barely peaks above the clutter.

V. DISCUSSION

Many aspects of nonlinear suppression for radar applications are currently under investigation. In addition to evaluating NLS performance in distributed clutter, two problems stand out: determining the correct amplitude threshold for the nonlinearity and selecting the appropriate NLS pulse code set.

V.1 AMPLITUDE THRESHOLD DETERMINATION

Setting the correct threshold value (α) in Eq (5) is key to achieving adequate suppression and avoiding unwanted distortion or signal loss. Figure 11 illustrates the situation where the compressed pulse is large in amplitude and distant from the desired data (far left-side of figure) – perhaps a least challenging, more desirable case. Setting the threshold above α_1 ensures that only the compressed pulse response is suppressed. However, setting it at or below α_2 will suppress desired data as well.

Given the situation depicted in Figure 11, the appropriate threshold may be found by estimating the data peak using a simple distance metric (e.g., Euclidean distance). To ensure this situation is encountered frequently, a large pulse compression ratio should be chosen, as well as, a code with low sidelobes.

The situation illustrated in Figure 12 represents a more challenging case. The compressed data contains several predominant peaks, one due to a compressed pulse (near sample #2700) and others due to sidelobes of high-power pulses that have been previously suppressed. The desired data is spread between samples 6000 and 8000. Due to the large undesirable sidelobes and low peak power of the

Figure 11. Amplitude Threshold Determination

Figure 12. Impact of High Sidelobes on Threshold Determination

desired signal, setting the threshold near 10^5 would eliminate much of the strong sidelobes as well as suppress the compressed target while leaving most of the desired data intact.

Several approaches to automatic threshold selection are under development, including:

1. Fusing data from all channels to set the threshold in any single channel. A strong peak in one channel provides information about target location that may be exploited by others.
2. Choosing "evaluation windows" to localize the threshold. The data size may be too large to set an effective threshold. For testing and simulation results in this paper, a single PRI was used. Smaller data blocks, on the order of a few pulse lengths, would allow for independent thresholding and suppression.

3. Using statistical models. Classical detection and estimation theory may be employed to estimate the target data levels.

V.2 NONLINEAR SUPPRESSION CODE SELECTION

The initial NLS evaluation employed two linear FM (LFM) codes, one with a positive slope and one with a negative slope, as a form of chirp diverse (variable slope) waveform coding. Although relatively simple and easy to implement, only two orthogonal codes are available with this technique and achievable ambiguity reduction is limited.

As revealed by examples in the previous section, high power targets may produce high-level sidelobes that hinder effective ambiguity suppression. Unlike spread-spectrum communication techniques, which normally use a continuously repeating spreading code, the envisioned ambiguity suppressing radar waveform is a finite sequence of coded pulses. This results in correlation occurring over partial code periods, rather than the full periodic correlation offered by a continuous code. On the other hand, typical radar pulse compression codes are designed to exhibit optimal autocorrelation properties, given a single code is used for all pulses. The partial period cross-correlation properties, and their affects on correlation based processing, are not well understood for either radar codes or spread-spectrum codes.

Although convenient for evaluation and testing, using well-known codes such as m -sequences and Gold Codes will not necessarily yield optimal results for NLS radar applications, since these codes are designed for improved correlation characteristics. Fundamentally, the code family best suited for meeting radar NLS requirements should 1) have a large partial period autocorrelation peak with low integrated sidelobe levels, and 2) provide maximum signal dispersion when cross-correlating pulse codes within the family (low partial period cross-correlation).

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