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Citation for published version (APA):

Document status and date:
Published: 01/10/2019

Document Version:
Accepted manuscript including changes made at the peer-review stage

Please check the document version of this publication:

• A submitted manuscript is the version of the article upon submission and before peer-review. There can be important differences between the submitted version and the official published version of record. People interested in the research are advised to contact the author for the final version of the publication, or visit the DOI to the publisher's website.
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Download date: 20. Nov. 2021
A Performance Enhancement Technique for a Joint FMCW RadCom System

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Abstract — The emerging trend of autonomously driving vehicles brings an increasing need for communication. In a setting comprising connected vehicles, information about the environment and vehicles themselves will be shared with other vehicles. Existing communication standards may not be able to meet this growing demand on communication bandwidth. Radar embedded communication can help overcome this bottleneck at the expense of degraded radar performance. In this work we introduce and demonstrate a novel radar signal processing technique to compensate for the self-interference due to the communication content on a chirp modulated radar signal. Recovery of the radar performance is demonstrated by simulations for the continuous phase modulation on the chirp modulation.

Keywords — Chirp Modulation, Connected Vehicles, Continuous Phase Modulation, FMCW, Radar Signal Processing.

I. INTRODUCTION

Frequency modulated continuous wave (FMCW) is the foremost waveform in automotive radar applications. Automotive radars utilize stretch processing at the receiver, which is also referred to as deramping, where the received waveform is mixed with the chirp generated by the radar synthesizer. As a result of this operation, ranges and velocities of scattering objects translate into frequencies.

In automotive systems, connected and automated driving requires not only sensing but also information exchange between vehicles to allow decision making in rapidly changing and complex environments. Different frequency bands are required to provide the transfer of sensor data besides traffic information [1]. Therefore, embedding communications into radar waveforms emerges as an alternative method to reduce congestion in the available spectrum.

Embedded communication results in self interference after deramping which can cause a masking of weak targets such as pedestrians or cyclists. To overcome this performance degradation, it is of interest to remove the communication signal before any subsequent radar processing steps. Joint radar and communication functionality of an FMCW radar which uses deramping in the receiver was demonstrated in [2]. Frequency shift keying was used to embed the data symbols in the radar signal. The proposed technique causes mixing products to appear in the baseband after deramping which could lead to false detection. In [3], embedded communication was realized by means of amplitude modulation of the radar signal. The resulting self interference of the mixing products were eliminated by choosing the carrier of the data signal such that the mixing products do not appear in the baseband of the radar receiver. The technique in [3] therefore requires additional bandwidth for the joint radar and communication signal. Moreover, non-linear power amplifiers limit the utility of amplitude modulation in radar signals.

In this paper, we propose the usage of continuous phase modulation (CPM) to embed communication into a radar which relaxes the power amplifier requirements and requires less additional bandwidth. In order to remove the interfering communication signal, the fact that the round trip time of the radar signal translates into a frequency is exploited. A frequency-dependent shift of the deramped signal is used to align the interfering communication signal. The aligned communication signal is removed before subsequent radar processing takes place. The performance recovery is demonstrated by means of simulations.

II. SYSTEM MODEL

An FMCW radar transmits a sequence of linear frequency modulated signals, also referred to as chirps. The radar signal serves as a carrier of the communication signal, which is embedded in the radar signal in form of phase modulation. The transmitted signal of the joint radar and communication signal of a single chirp is given as

\[ s(t) = \begin{cases} \sqrt{P} \text{Re} \{ x(t) \cdot \exp(j\phi(t)) \}, & 0 \leq t \leq T \\ 0, & \text{elsewhere} \end{cases} \]

where

\[ x(t) = \exp\left(j2\pi \left( f_0 t + \frac{\alpha}{2} t^2 \right) \right) \]

is the chirp with slope \( \alpha = B/T \). \( B \) and \( T \) are the chirp bandwidth and duration, respectively, see Fig. 1. The embedded CPM communication signal \( \phi(t) \) is conveying several symbols of duration \( T_{\text{symb}} \) per chirp. The transmitted signal \( s(t) \) gets reflected by a radar target which is located at \( t = 0 \) at a radial distance \( R \) and moves with a constant relative velocity \( v \). The radar receiver performs deramping by complex conjugate mixing of the received joint radar and communication signal and the transmitted chirp without communication as shown in [2].

The low pass filtered in-phase and quadrature components are sampled within an acquisition window of length \( T_{\text{acq}} \). To ensure that the baseband signal is present during the entire
therefore be considered constant, i.e. be neglected and the delay of the communication signal can window. The Doppler effect on the communication signal can phase term depending on the starting point of the acquisition frequency, \( f \) are the range and velocity dependent beat and Doppler frequency, \( f_s \) is the sampling frequency, and \( \psi \) is a constant phase term depending on the starting point of the acquisition window. The Doppler effect on the communication signal can be neglected and the delay of the communication signal can therefore be considered constant, i.e.

\[
\tau_0 = \frac{2f_sR}{c}.
\]

Since the deramped signal only occupies the negative part of the spectrum, image rejection and decimation is carried out before subsequent processing.

### III. Proposed Technique

Range and Doppler information of the target is carried in the frequency and phase of the received signal, as indicated in (3). The phase attached communication signal interferes with this information; therefore, it is of interest to remove the communication signal prior to any subsequent radar processing. If the symbol time of the communication signal \( T_{\text{sym}} \) is larger than the round trip time of the radar signal \( \tau_0 \), complex conjugate mixing with the transmitted signal can be performed to remove the communication signal. This approach fails if \( T_{\text{sym}} \) is comparable to or less than \( \tau_0 \). In this paper, we address the latter case, which is relevant for communications at high symbol rates.

#### A. Frequency-dependent Time Shift

The maximum unambiguous detectable Doppler frequency is limited by

\[
f_{D,\text{max}} = \frac{1}{2 \cdot \text{PRI}}
\]

where PRI is the pulse repetition interval. The PRI is typically 3 orders of magnitudes larger than \( 1/f_s \). Therefore, \( f_b \) is the dominant part of the frequency in (3) and the intermediate frequency can be approximated by

\[
f_b + f_D \approx f_b.
\]

Moreover, comparing (4) and (6), it can be seen that

\[
f_b = \alpha \tau_0.
\]

Thus, the frequency of the deramped signal is proportional to the delay of the radar signal. This fact can be exploited to align the communication content with different delays by shifting each deramped signal in time according to its beat frequency. This frequency dependent shift technique was first introduced in [5] to compensate for slope non-linearities.

The shifting of a signal in the time corresponds to a multiplication with a linear phase term in the frequency domain. Therefore, the required time shift can be expressed as the phase response of a filter. The time shift property of a filter is determined by the phase and group delay \( \tau_{ph} \) and \( \tau_{gr} \), respectively. The group delay at a given angular frequency \( \Omega \) is the negative slope of the tangent of the phase response at this point and determines the time shift of a narrowband signal. The group delay is therefore given as

\[
\tau_{gr}(\Omega) = -\frac{\text{d} \angle H(j\Omega)}{\text{d} \Omega}.
\]

The beat frequency dependent time shifting is performed in the digital domain after all samples of the deramped chirp have been collected. Thus, (circularly) shifting forward and backward in time is possible. In the following, the alignment of the communication signal to the beginning of the acquisition window is demonstrated. Therefore, the signal must be shifted according to (9) by

\[
\tau_{gr}(\Omega) = \frac{\Omega B}{2\pi \alpha v}.
\]
\[ \omega \text{ is the normalized angular frequency:} \]
\[ \phi(t) \text{ and } \phi_2(t) \text{, respectively. This operation corresponds to} \]
\[ \frac{1}{\pi f_s} \text{quadratic phase expression from (13)} \]
\[ \text{the frequency domain by multiplying the spectrum with the} \]
\[ B \text{. Recovery of the Radar Signal} \]
\[ \text{where } \Omega_B = 2\pi f_B \text{ is the angular beat frequency. From this} \]
\[ \text{we can derive the required phase response } \angle H(j\Omega) \text{ of the} \]
\[ \text{corresponding filter using (10) which results in} \]
\[ \angle H(j\Omega) = \frac{\Omega^2}{4\pi\alpha}. \] (12)
\[ \text{The resulting phase response is non-linear with respect to the} \]
\[ \text{frequency and thus causes a dispersion of the signal. This fact} \]
\[ \text{limits the usable communication bandwidth and suggests the} \]
\[ \text{use of bandwidth efficient modulation techniques. CPM is not} \]
\[ \text{only bandwidth efficient, it also has constant envelope which} \]
\[ \text{makes it robust against non-linear distortions. The usage of} \]
\[ \text{CPM as a phase attached communication signal in a radar} \]
\[ \text{was proposed in [6].} \]
\[ \text{The corresponding phase response in the discrete time} \]
\[ \text{domain } \angle H(e^{j\omega}) \text{ is given by using the relation } \Omega = \omega f_s \text{ with} \]
\[ \omega \text{ is the normalized angular frequency:} \]
\[ \angle H(e^{j\omega}) = \omega^2 f_s^2 \frac{1}{4\pi\alpha}. \] (13)
\[ \text{B. Recovery of the Radar Signal} \]
\[ \text{The frequency-dependent time shift could be done in the} \]
\[ \text{frequency domain by multiplying the spectrum with the} \]
\[ \text{quadratic phase expression from (13)} \]
\[ \tilde{y}[n] = \mathcal{F}^{-1} \left\{ \mathcal{F} \{y[n]\} \cdot \exp \left( j\frac{\omega^2 f_s^2}{4\pi\alpha} \right) \right\} \] (14)
\[ \text{where } \mathcal{F} \text{ and } \mathcal{F}^{-1} \text{ denote the discrete Fourier transform (DFT)} \]
\[ \text{and its inverse, respectively. This operation corresponds to} \]
\[ \text{a fractional shift of the signal. The resulting signal reads therefore} \]
\[ \tilde{y}[n] = \exp \left( -j2\pi \left( f_B n + \frac{2f_s R}{c} \right) + j\phi[n] + \Theta_B \right), \] (15)
\[ \text{where } \phi[n] \text{ is the shifted communication signal and } \Theta_B \text{ is a} \]
\[ \text{constant phase term depending on the beat frequency. In the} \]
\[ \text{dispersion-free case, } \phi[n] \text{ equals } \phi[n]. \]
\[ \text{A snapshot of the phase of two differently delayed versions of} \]
\[ \text{the same communication signals } \phi(t) \text{ are presented in} \]
\[ \text{Fig. 3a. The signals are delayed by } \tau_1 \text{ and } \tau_2 \text{ with } \tau_1 < \tau_2 \text{ and} \]
\[ \text{have corresponding beat frequencies of } f_{B_1} \text{ and } f_{B_2}. \text{Fig. 3b} \]
\[ \text{shows the signals after the frequency-dependent time shift. It} \]
\[ \text{can be clearly seen that the signal undergoes a longer time} \]
\[ \text{shift due to its higher beat frequency.} \]
\[ \text{The shifting operation aligns the two phase modulation} \]
\[ \text{components in time. After the alignment, the resulting} \]
\[ \text{signal is multiplied by the complex conjugate phase of the} \]
\[ \text{communication signal} \]
\[ \hat{y}[n] = \tilde{y}[n] \cdot \exp(-j\phi[n]). \] (16)
\[ \text{The resulting signal reads therefore} \]
\[ \hat{y}[n] = \exp \left( -j2\pi \left( f_B n + \frac{2f_s R}{c} \right) + j\epsilon[n] \right) \] (17)
\[ \text{where } \epsilon[n] \text{ is the residual phase error due to dispersion which} \]
\[ \text{is shown in Fig. 3c. The effect of the dispersion can also be} \]
\[ \text{seen in Fig. 4, which represents the two power spectrums of the} \]
\[ \text{two differently delayed signals. Ideally, the spectrum of both} \]
\[ \text{signals should be multiplied by the linear phase according to} \]
\[ \text{the slope of the tangent at their angular beat frequencies as} \]
\[ \text{indicated. This requirement is not met due to the quadratic} \]
\[ \text{nature of the phase response and causes different shifts of the} \]
\[ \text{single frequency components.} \]
\[ \text{C. Constraints} \]
\[ \text{The proposed compensation technique imposes constraints in the} \]
\[ \text{time and frequency domain properties of the joint radar} \]
\[ \text{and communication signal.} \]
Firstly, the sampled signal must encompass the entire communication signal, independent of its delay. This is ensured by introducing guard intervals at the beginning and end of the chirp depending on the starting and end points of the acquisition window. The guard intervals can be seen in Fig. 1 while Fig. 5 shows their effect. Secondly, due to the discontinuity of the phase response in $\omega = 0$, the entire spectrum of the deramped signal must lie within the interval $(-f_s, 0]$. Hence, the bandwidth of the communication signal is limited by

$$B_{\text{com}} = 2 \cdot \min (f_B + f_D, f_s - (f_B + f_D))$$

(18)

which can be seen from Fig. 4 that shows the flipped spectrum for sake of simplicity.

IV. NUMERICAL SIMULATIONS

Fig. 6a illustrates a range-Doppler plot for two targets at a radial distance of 85 m and 95 m with relative velocities of the of -5 m/s and 5 m/s, respectively. The radar emits 256 chirps with a PRI = 30.6 $\mu$s, $B = 250$ MHz and $T = 25.6$ $\mu$s at a carrier frequency $f_c = 77$ GHz. The sampling frequency after decimation is $f_s = 10$ MHz. In the resulting range-Doppler plot both targets can be clearly distinguished.

As an example for a joint radar and communication signal, Gaussian minimum shift keying (GMSK) modulation with a symbol time of $T_{\text{symb}} = 1$ $\mu$s and a time bandwidth product of 0.3 is used to embed the communication signal in the radar waveform. The resulting interference from the embedded communication signal is shown in Fig. 6b. The embedded communication signal causes a broadening of the signal in the range domain and any Doppler information is lost. As a result, the two targets cannot be detected. The novel mitigation technique leads to significant recovery of the radar performance. Both targets can be detected again as it can be seen in Fig. 6c.

V. CONCLUSION

A novel radar signal processing technique has been introduced in this paper. It removes the embedded communication signal from the deramped radar waveform by shifting the communication content according to its delay. The concept was demonstrated using a numerical simulation in which only the propagation delay had to be compensated. For practical implementation, all group delays caused by the individual components of the receiver must be known and compensated. Depending on the selected system parameter, the dispersion of the communication signal can influence the detection and estimation performance of the radar. Further research will focus on mitigation techniques for this case.

VI. ACKNOWLEDGEMENT

This research is supported by the Dutch Technology Foundation TTW which is part of the Netherlands Organisation for Scientific Research (NWO), and which is partly funded by the Ministry of Economic Affairs. The authors acknowledge discussions with NXP colleagues on the method proposed here.

REFERENCES