Non-Coherent PSK-Based Dual-Function Radar-Communication Systems

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***Abstract*—Dual-function radar-communication (DFRC) systems enable information embedding into the radar signal emission. Existing methods for non-coherent phase-modulation DFRC employ multiple pairs of orthogonal waveforms and embed one communication symbol into each pair. The total number of symbols is equal to one-half of the number of waveforms. In this paper, we propose a new signaling strategy for embedding a higher number of communication symbols. The proposed method implements non-coherent phase-shift keying (PSK) by employing one of the orthogonal waveforms as a common reference and modulating the information in terms of the phase differences between all other waveforms and the reference waveform. The number of communication symbols that can be embedded equals the total number of waveforms minus one. We introduce two schemes for achieving a desired phase constellation. The proposed approach is shown to achieve a two-fold increase in the data rate compared to existing methods for a large number of waveforms.**

I. INTRODUCTION

Dynamic and shared spectrum access is emerging as effective means for mitigation of frequency congestions [1]–[4]. In particular, the coexistence of radar and communications has been the focus of intensive research [5]–[12]. Methods for spectral sharing between radar and communications include cooperative sensing and signal sharing [7]. The system concept of dual-function radar-communications (DFRC) treats communications as secondary to the primary radar function [13]–[19]. DFRC systems assign all available resources, e.g., the entire bandwidth and total transmit power, to the primary radar function. The secondary communication function is achieved by embedding signal symbols into the radar emission.

Recently, several techniques have been developed for embedding information into the radar emission, including waveform diversity-based method [13], sidelobe amplitude modulation (AM) method [14], multi-waveform amplitude shift keying (ASK) method [15], [17], and phase modulation (PM) method [16]. The waveform diversity technique employs a dictionary of waveforms, where each waveform represents a communication symbol, and achieves information embedding by radiating one of the waveforms during each radar pulse. Based on which waveform is radiated and reliably received, the receiver can decode the corresponding information. The sidelobe AM method uses time modulated arrays to introduce variations into the sidelobe levels (SLLs) of the instantaneous pattern towards the communication direction. Each SLL represents a specific communications symbol. Motivated by the recent advances in multiple-input multiple-output (MIMO) radar [20]–[25] and the successful exploitation of simultaneously transmitting multiple independent radar waveforms, the multi-waveform amplitude shift-keying (ASK) method [15] embeds information using bi-level sidelobe control. This technique enables embedding binary information via each waveform. The number of bits that can be embedded during each radar pulse is proportional to the number of orthogonal waveforms being used.

The essence of PM–based information embedding is to map the binary data into a phase symbol that belongs to the phase dictionary of an appropriate size. During each radar pulse, the method in [16] employs multiple pairs of orthogonal waveforms, where each pair represents one phase symbol. At the communication receiver, each embedded symbol is detected by estimating the phase difference between the signals associated with the two waveforms in each pair and, subsequently, deciphering the corresponding binary sequence. Compared to the AM and multi-waveform ASK methods, the PM-based method enables information embedding towards communication receivers located within and outside the main beam. It was also shown in [16] that the PM-based method has superior bit error rate (BER) compared to other methods.

In this paper, we develop a new DFRC system where the same transmit waveforms are used for both functions. In essence, co-existence is achieved using the same transmit resources in lieu of using separate signaling schemes. Specifically, a new multi-waveform PSK-based method is proposed for embedding information into the radar emission. During each radar pulse, the proposed method embeds K communication symbols, where each symbol represents L bits. To realize this scheme, a total number of $K+1$ orthogonal waveforms are transmitted simultaneously. One of the waveforms is taken as a common reference to all other waveforms, enabling

The work of A. Hassanien, M. G. Amin, Y. D. Zhang, and F. Ahmad was supported in part by a subcontract with Defense Engineering Corporation for research sponsored by the Air Force Research Laboratory under Contract FA8650-12-D-1376.

the formation of K waveform pairs. Each communication symbol is embedded in the phase difference between the two signal components associated with each waveform pair. The communication receiver employs a bank of matched filters to extract the signal components associated with each waveform and, then proceeds to estimate and decode the embedded communication symbols. The data rate that can be achieved is given by the product of the number of communication symbols per pulse, the number of binary bits per symbol, and the radar pulse repetition rate. Assuming that K is odd, the method in [16] employs only $(K+1)/2$ waveform pairs which can achieve a data rate $L(K+1)/2$. Therefore, the proposed method increases the data rate by a factor $\eta = 2-2/(K+1)$, without affecting the BER performance. Simulation examples are provided to verify the effectiveness of the developed method.

The remainder of this paper is organized as follows. The signal model is described in Sec. II and the proposed information embedding scheme is presented in Sec. III. Sec. IV provides the simulation results and conclusions are drawn in Sec. V.

II. Signal Model

Consider a DFRC system with M transmit antennas arranged as a uniform linear array. It is assumed that appropriately designed transmit beamforming is used to satisfy the primary radar operation requirements, i.e., to focus the transmit power within the radar main beam, while keeping the sidelobe level below a certain value. Let $\mathbf{w}_k(\tau),\ k = 0, \ldots, K$, be the $M \times 1$ transmit beamforming weight vectors used to simultaneously form $K+1$ transmit patterns, where τ is the pulse index. Although the $K+1$ transmit beamforming weight vectors change from pulse to pulse, during each radar pulse they are selected from a larger set of weight vectors which can be designed off-line as described in the next section. The $M \times 1$ baseband transmit signal vector during the τth radar pulse is modeled as

$$\mathbf{s}(t;\tau) = \sum_{k=0}^{K} \mathbf{w}_k^*(\tau)\psi_k(t), \tag{1}$$

where t is the fast time index, $(\cdot)^*$ stands for the complex conjugate, and $\psi_0(t),\ \psi_1(t), \ldots, \psi_K(t)$ are $K+1$ radar waveforms. It is assumed that each waveform is normalized to have unit power, i.e. $\int_{T_p} |\psi_k(t)|^2 dt = 1$ for $k = 0, \ldots, K$, where T_p is the radar pulse duration. It is further assumed that the orthogonality condition

$$\int_{T_p} \psi_k(t)\psi_j^*(t)dt = 0, \quad k \neq j,\ k,j = 0, \ldots, K, \tag{2}$$

is satisfied. It is worth noting that the radar operation represents the primary function of the DFRC system, the spectral contents of all waveforms overlap in the frequency domain occupying the entire bandwidth of the joint system. Since the transmit waveforms, according to Equation (1), is a weighted sum of the orthogonal waveforms, the desired waveform properties of $\psi_k(t),\ k = 0, \ldots, K$, do not necessarily map to that of $\mathbf{s}(t;\tau)$. As such, constant modulus or low peak to average power ratio of $\mathbf{s}(t;\tau)$ must be eventually sought out in the final consideration of the proposed system. Note that, in practice, perfectly orthogonal coded waveforms cannot be achieved and, therefore, waveforms with low cross-correlations should be used. The problem of waveform design with desirable low cross-correlation properties has been extensively investigated in the literature (see [26]–[28], and references therein). However, the problem of waveform design is out of the scope of this paper.

Assume that a single-element communication receiver is located in direction θ_c. The baseband signal at the output of the communication receiver can be expressed as

$$r(t;\tau) = \alpha_{\mathrm{ch}}(\tau)\mathbf{a}^T(\theta_c)\mathbf{s}(t;\tau) + n(t;\tau), \tag{3}$$

where $\alpha_{\mathrm{ch}}(\tau)$ is the channel coefficient of the received signal which summarizes the channel gain between the transmit array and the communication receiver during the τth pulse, $\mathbf{a}(\theta_c)$ is the $M \times 1$ steering vector of the transmit array toward the spatial angle θ_c, $(\cdot)^T$ stands for the transpose operation, and $n(t;\tau)$ is the additive white Gaussian noise with zero mean and variance σ_n^2. Based on the signaling strategies used to construct the transmit signal vector $\mathbf{s}(t;\tau)$, communication symbols can be embedded into the radar emission. In the next section, we propose an efficient signaling strategy for simultaneously embedding multiple phase symbols during each radar pulse.

III. Proposed Multi-Waveform PSK Based Information Embedding

Let us consider $\psi_0(t)$ as a common reference to all other waveforms by using the same weight vector $\mathbf{w}_0$ instead of $\mathbf{w}_0(\tau)$ for all radar pulses. Then, substituting (1) in (3), the communication receiver output signal is rewritten as

$$\begin{aligned} r(t;\tau) &= \alpha_{\mathrm{ch}}(\tau)\Bigg(\mathbf{w}_0^H\mathbf{a}(\theta_c)\psi_0(t) \\ &\qquad + \sum_{k=1}^{K}\mathbf{w}_k^H(\tau)\mathbf{a}(\theta_c)\psi_k(t)\Bigg) + n(t;\tau) \\ &= \alpha_{\mathrm{ch}}(\tau)\left(G_0\psi_0(t) + \sum_{k=0}^{K} G_k(\tau)\psi_k(t)\right) + n(t;\tau), \end{aligned} \tag{4}$$

where $(\cdot)^H$ is the Hermitian operation, $G_0 \triangleq \mathbf{w}_0^H\mathbf{a}(\theta_c)$ and $G_k(\tau) \triangleq \mathbf{w}_k^H(\tau)\mathbf{a}(\theta_c),\ k = 1, \ldots, K$, are the complex transmit gains towards the communication direction θ_c during the τth radar pulse. Making use of the special structure of (4), a set of K phase rotations can be defined as follows

$$\phi_k(\tau) = \mathrm{angle}\left(\frac{G_k(\tau)}{G_0}\right), \quad k = 1, \ldots, K, \tag{5}$$

where $\mathrm{angle}(\cdot)$ is the phase angle of a complex number, measured in radians. A number of K communication symbols can be embedded in each radar pulse via modulating the phase rotations given by (5), i.e., each phase rotation $\phi_k(\tau)$ can be selected from a pre-specified discrete phase constellation. The following two subsections present a method for constellation realization using beamformer design and the proposed signaling strategy for embedding information at the transmitter and reliably detecting it at the receiver.

A. Phase Constellation Realization via Beamformer Design

Let $\mathbf{\Omega} = \{\Omega_1, \ldots, \Omega_{2^L}\}$ be a desired constellation of 2^L phase symbols which can be chosen from the interval $[0,\ 2\pi]$ using a uniform or non-uniform grid. Each symbol represents L bits of binary information. Assume that the beamforming weight vector $\mathbf{w}_0$ associated with the reference waveform $\psi_0(t)$ is already designed[1] and let us denote this vector as the principal weight vector. Consider a bank of 2^L weight vectors, denoted as $\mathbf{u}_1, \ldots, \mathbf{u}_{2^L}$, such that the transmit radiation pattern associated with each of these vectors is the same as that of the principal weight vector except for some phase rotation towards the intended communication direction. The phase constellation $\mathbf{\Omega}$ can be realized by taking the phase symbols as

$$\Omega_\ell = \text{angle}\left(\frac{\mathbf{u}_\ell^H \mathbf{a}(\theta_c)}{G_0}\right), \quad \ell = 1, \ldots, 2^L. \tag{6}$$

Given the principal weight vector $\mathbf{w}_0$, the weight vectors $\mathbf{u}_1, \ldots, \mathbf{u}_{2^L}$, which satisfy (6), can be designed using one of the following two methods.

1) Convex Optimization Based Design: One way to design $\mathbf{u}_\ell$ such that it has the same pattern as that of $\mathbf{w}_0$ while satisfying (6) is by minimizing the norm of the difference between the two weight vectors, i.e., by minimizing the deviation of $\mathbf{u}_\ell$ from $\mathbf{w}_0$ while enforcing one linear constraint to satisfy the phase requirements. This can be formulated as the following optimization problem

$$\min_{\mathbf{u}_\ell} \left\| \mathbf{w}_0 - \mathbf{u}_\ell \right\| \quad \text{subject to} \quad \mathbf{u}_\ell^H \mathbf{a}(\theta_c) = G_0 e^{-j\Omega_\ell}. \tag{7}$$

The optimization problem (7) is convex and, therefore, can be efficiently solved using the interior point methods [30]. Another advantage to this convex optimization based design is that the desired phase rotations (6) are satisfied with equality. However, the obtained solutions yield transmit weight vectors with very similar, but not identical, patterns to that of $\mathbf{w}_0$.

2) Transmit Radiation Pattern Invariance Based Design: Following the guidelines in [29], the $M \times 1$ principal weight vector $\mathbf{w}_0$ can be used to generate a population of 2^{M-1} weight vectors of the same dimensionality, all having the same transmit power radiation pattern as that of $\mathbf{w_0}$. Let us denote the aforementioned population as $\mathbf{V}_{\text{pop}} = \{\mathbf{v}_1, \ldots, \mathbf{v}_{2^{M-1}}\}$. The required beamforming weight vectors $\mathbf{u}_1, \ldots, \mathbf{u}_{2^L}$ can be selected from $\mathbf{V}_{\text{pop}}$ while trying to satisfy the phase requirements in (6). Here, we propose to select the desired weight vector according to the following criterion

$$\begin{aligned} \min_{\mathbf{w_0}, \mathbf{u}_1, \ldots, \mathbf{u}_{2^L}} \quad & \sum_{\ell=1}^{2^L} \left| \Omega_\ell - \text{angle}\left(\frac{\mathbf{u}_\ell^H \mathbf{a}(\theta_c)}{G_0}\right) \right| \\ \text{subject to} \quad & \{\mathbf{w_0}, \mathbf{u}_1, \ldots, \mathbf{u}_{2^L}\} \in \mathbf{V}_{\text{pop}}. \end{aligned} \tag{8}$$

[1]A variety of methods can be used from the literature to design a transmit beamforming weight vector that satisfies the requirements dictated by the radar function (see [29] and references therein).

The formulation (8) is non-convex and, therefore, is difficult to solve in a computationally efficient manner. We use exhaustive search to select the best weight vector combination. It is worth noting that the principal weight vector is also selected jointly with $\mathbf{u}_1, \ldots, \mathbf{u}_{2^L}$ to makes sure that the closest phase constellation to the desired one is achieved. However, the replacement of the principal weight vector with another vector from $\mathbf{V}_{\text{pop}}$ does not change the transmit radiation pattern. Note that for a small sized constellation, i.e., for small L, it is easy to satisfy (6) by solving (8). However, as L becomes larger, the computational complexity associated with performing exhaustive search to solve (8) grows exponentially.

B. Proposed Information Embedding Strategy

Let $\{b_1(\tau), \ldots, b_{LK}(\tau)\}$ be a binary sequence of length KL bits that needs to be embedded during the τth radar pulse. The sequence is then partitioned into K segments of L bits each, i.e., the kth segment contains the binary bits $b_{(k-1)L+1}, \ldots, b_{kL}$. Each segment is mapped into the corresponding communication symbol. Let us define the indices

$$I_k(\tau) = \mathcal{D}(b_{(k-1)L+1}, \ldots, b_{kL}), \quad k = 1, \ldots, K, \tag{9}$$

where $\mathcal{D}$ is the operator that maps a set of binary bits into the corresponding integer $I_k(\tau)$, which ranges between 0 and $2^L - 1$. During the τth radar pulse, the indices $I_1(\tau), \ldots, I_K(\tau)$ are used to determine the K phase symbols from the phase constellation that need to be embedded and the corresponding transmit weight vectors. Therefore, the signal model (4) can be rewritten as

$$\begin{aligned} r(t;\tau) = \alpha_{\text{ch}}(\tau) \Bigg(& \mathbf{w}_0^H \mathbf{a}(\theta_c) \psi_0(t) \\ & + \sum_{k=1}^{K} \mathbf{e}_k^T(\tau) \mathbf{U}^H \mathbf{a}(\theta_c) \psi_k(t) \Bigg) + n(t;\tau), \end{aligned} \tag{10}$$

where $\mathbf{U} \triangleq [\mathbf{u}_1, \ldots, \mathbf{u}_{2^L}]$, and $\mathbf{e}_k(\tau)$ is the $2^L \times 1$ vector of all zeros except for the $(I_k(\tau)+1)$th entry which equals one.

Matched filtering the received data in (10) to $\psi_0(t)$ yields

$$y_0(\tau) = \alpha_{\text{ch}}(\tau) \mathbf{w}_0^H \mathbf{a}(\theta_c) + \tilde{n}_0(\tau), \tag{11}$$

where $\tilde{n}_0(\tau)$ is the noise at the output of the matched filter whose variance is the same as that of $n(t;\tau)$. Similarly, matched filtering the received data in (10) to each of the other K transmitted orthogonal waveforms yields the virtual data sets $y_k(\tau),\ k = 1, \ldots, K$, expressed as

$$y_k(\tau) = \alpha_{\text{ch}}(\tau) \mathbf{e}_k^T(\tau) \mathbf{U}^H \mathbf{a}(\theta_c) + \tilde{n}_k(\tau), \quad k = 1, \ldots, K, \tag{12}$$

where $\tilde{n}_k(\tau)$ is the noise at the output of the kth matched filter with variance σ_n^2. To detect the embedded communication symbols, the receiver estimates the phase difference between each of the virtual received signals in (12) and the reference signal in (11), that is

$$\hat{\phi}_k(\tau) = \text{angle}\left(\frac{y_k(\tau)}{y_0(\tau)}\right), \quad k = 1, \ldots, K. \tag{13}$$

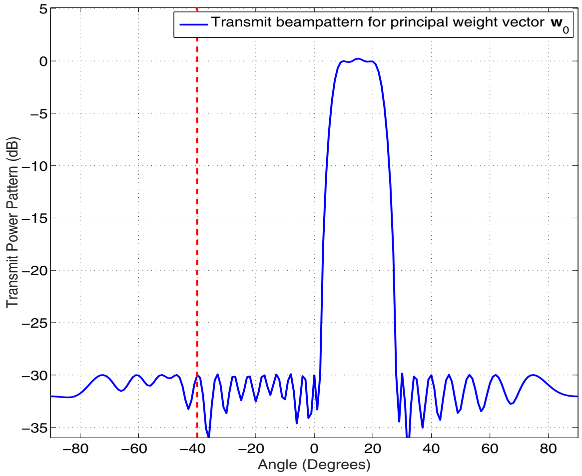

Fig. 1. Transmit power distribution versus spatial angle.

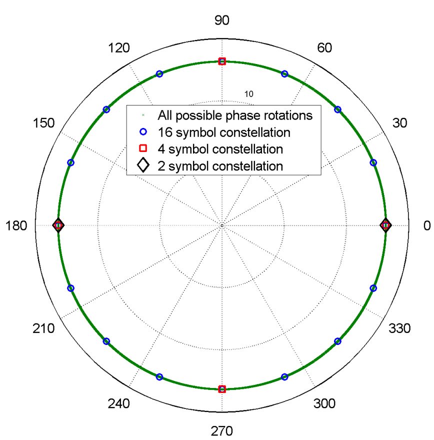

Fig. 2. Phase-rotation distribution versus spatial angle for a population of 2^{12} and phase constellations for $L = 1, 2$, and 4.

The receiver then proceeds to map the estimated phase differences into the corresponding binary format to retrieve the binary sequence.

IV. Simulation Results

In the simulation, we consider a uniform linear transmit array consisting of $M = 25$ antennas spaced one-half wavelength apart. We assume that the main radar operation takes place within the sector $\mathbf{\Theta} = [10° \ 20°]$. Three examples are considered. In the first example, we investigate the ability to design the principal weight vector that yields a certain beampattern and use it to build the desired constellations. In the second and third examples, we use the designed weight vector and the obtained constellations to test and analyze the performance of the proposed technique for information embedding towards the receivers located in the sidelobe and the main beam, respectively.

Example 1: Beampattern Synthesis and Phase Constellation Realization

We design the principal transmit beamforming weight vector $\mathbf{w}_0$ by solving the following optimization problem

$$\min_{\mathbf{w}_0} \max_{\theta} \left| |\mathbf{w}_0^H \mathbf{a}(\theta)| - 1 \right|, \quad \theta \in \Theta \tag{14}$$

$$\text{s.t.} \left| \mathbf{w}_0^H \mathbf{a}(\bar{\theta}) \right|^2 \leq \varepsilon, \quad \bar{\theta} \in \bar{\mathbf{\Theta}}, \tag{15}$$

$$\left| \mathbf{w}_0^H \mathbf{a}(\theta_c) \right|^2 = \varepsilon, \tag{16}$$

where $\bar{\mathbf{\Theta}}$ represents the out-of-sector region. We set $\varepsilon = 10^{-3}$ because it is assumed that the radar operation requires the power level emitted in the sidelobe areas to be at least 30 dB lower than the mainlobe. A single communication direction towards the spatial direction $\theta_c = -40°$ is assumed. Note that the constraint (16) is needed only when the communication direction is located in the sidelobe region. Fig. 1 shows the transmit beampattern obtained by solving the optimization problem (14)–(16). It is clear from the figure that the sidelobe attenuation with respect to the main beam is larger than 30 dB for all directions within the out-of-sector region. The figure also shows that the SLL towards the communication direction attains the maximum value allowable to ensure that the communication receiver receives the highest possible power within the sidelobe of the radar.

Following the guidelines in [29], a polynomial of order 24 whose coefficients are given by the entries of the principal weight vector is constructed and its 24 roots are calculated. This number of roots enables the generation of a population of 2^{24} weight vectors which have exactly the same transmit power patterns as that of the principal weight vector. In this simulation, we select 12 roots and use them to generate $\mathbf{V}_{\text{pop}}$ of 2^{12} weight vectors. Using the generated population of vectors, (8) is solved three times for $L = 4$, $L = 2$, and $L = 1$ in order to generate phase constellations of 16, 4, and 2 symbols, respectively. Fig. 2 shows all possible phase rotations associated with these vectors in the population $\mathbf{V}_{\text{pop}}$. The figure also depicts the three different phase constellations obtained by solving (8). It can be observed from the figure that the generated phase constellations are uniformly distributed on the unit circle, i.e., within the $[0, \ 2\pi]$ domain.

Example 2: Communications Performance within Sidelobe Region

In the second example, we investigate the performance of the proposed multi-waveform PSK method in terms of the BER. The performance of the proposed method is compared with the techniques proposed in [15] and [16]. We assume that the average transmit power from each transmit antenna is normalized to 1, i.e., total transmit power is fixed to $P_{\text{total}} = M$. For every method considered, the total transmit power is divided evenly among the number of waveforms used. For the method of [15], two transmit beamforming weight vectors with the same beampattern within the main radar beam and different SLLs towards the communications direction are used. In addition, 8 waveforms are used to embed 8 bits

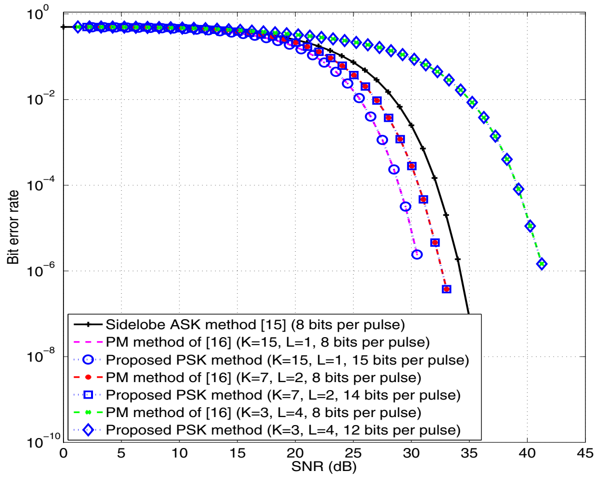

Fig. 3. BER versus SNR; second example.

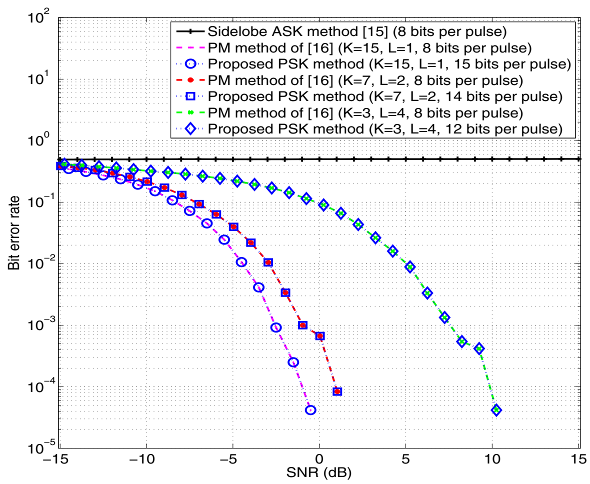

Fig. 4. BER versus spatial angle of the communication receiver; second example.

of information per pulse. For the method of [16] and the proposed method, the three phase constellations obtained from the previous example along with the associated weight vectors are used. For the case of $L = 1$, i.e., constellation of 2 phase symbols, the method of [16] employs 8 waveform pairs (16 waveforms) to embed 8 bits of information per pulse. Using the same number of waveforms, the proposed method employs one waveform as a reference and utilizes the remaining 15 waveforms to embed 15 bits of information per pulse. For the case of $L = 2$, the constellation contains 4 phase symbols, i.e., 2 bits per symbol. The method of [16] employs 4 pairs of waveforms to embed 8 bits of information per pulse. The proposed method, using the same number of 8 waveforms, employs one waveform as a reference and embeds 14 bits of information per pulse via the other 7 waveforms. The last case of $L = 4$ corresponds to a phase constellation of 16 phase symbols, i.e., it enables embedding 4 bits of information per symbol. The method of [16] uses 2 pairs of waveforms to embed 8 bits of information per pulse while the proposed method uses 4 waveforms and embeds 12 bits per pulse, i.e., 4 bits per waveform after reserving one of the waveforms as a reference.

To compute the BER, 10^6 pulses are considered, i.e., the process of information embedding is repeated independently for 10^6 times. The propagation coefficient α_c is modeled as a random variable with unit magnitude and uniformly distributed random phase. Fig. 3 shows the BERs for the three methods tested versus the signal-to-noise ratio (SNR). The figure shows that the BER curves for the proposed method coincide with the BER curves for the method of [16] for the three constellations used. This was expected because both methods use the same number of waveforms and assign the same amount of power to each waveforms. However, the proposed method achieves a higher data rate as compared to the method of [16] due to the novel structure of the transmit signaling used to embed the information. The figure also shows the BER for the sidelobe ASK method of [15] which exploits a constellation of two amplitude symbols. It can be seen from the figure that the BER performances for the proposed PSK based method and the PM based method of [16] using a constellation of two phase symbols are superior to the BER for the sidelobe ASK method.

Example 3: Communications Within Main Radar Beam

In the last example, we simulate the case of radar-embedded communications toward a receiver located within the main beam of the radar. The communication receiver is assumed to be located in the direction $\theta_c = 17°$. The sidelobe ASK functions within the sidelobe region only. However, we include it in the comparison in this example to show that a communications receiver located in the main radar beam cannot detect/decode a sidelobe ASK based radar-embedded message. The proposed PSK based method as well as the method of [16] enable communications within the main beam with higher accuracy due to the high transmit power gain within the main beam. Fig. 4 depicts the BER versus the SNR for all methods. As expected, it can be observed from the figure that the sidelobe ASK method of [15] fails to decode any message in the main beam region. It can also be confirmed from the figure that the BER curves for the proposed PSK based method and the method of [16] mirror the corresponding curves in Fig. 3 but at a much lower SNR. The improvement in SNR is attributed to the high gain within the main radar beam. This enables the use of PSK based methods to deliver information towards a communication receiver located at much longer ranges from the transmitter when they are spatially positioned within the main beam of the radar.

V. Conclusion

A new multi-waveform PSK based approach for dual-function radar-communication system has been developed. In this approach, a sequence of LK bits of information per radar pulse can be delivered and requires $K + 1$ orthogonal waveforms, each is used to embed L bits of information. Compared with previous PM based methods, the proposed method enables an increase in the data rate by a factor of $2 - \frac{2}{K+1}$. For large values of K, the achieved data rate is almost twice that of existing methods. Simulation examples were carried out to validate the effectiveness of the proposed approach.

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