Signaling Strategies for Dual-Function Radar-Communications: An Overview

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I. Introduction

The last decade witnessed a growing demand on radio frequency which is driven by technological advances benefiting the end consumer, but requiring new allocations of frequency bandwidths. Further, higher data rates for faster communications and wireless connections have called for an expanded share of existing frequency allocations. Concerns for spectrum congestion and frequency unavailability have spurred extensive research efforts on spectrum management and efficiency [1]–[4] within the same type of service, and have led to cognitive radio [5] and cognitive radar [6]. On the other hand, devising schemes for co-existence among different services have eased the competition for spectrum resources, especially for radar and wireless communication systems [7]–[14]. Both systems have been recently given a common portion of the spectrum by the Federal Communications Commission (FCC).

This paper discusses dual-function systems which are a special case of co-existence. In these systems, the same transmitter or receiver resources along with their temporal and/or spatial degrees of freedom are employed to achieve the objectives of two systems. Strategies for radar-embedded communication signals have successfully established a dual-function system that simultaneously performs both radar and communication functions. Fig. 1 provides an illustration of such a dual-function radar-communication (DFRC) system. Depending on the quality of services, error tolerance, and accuracy in signal estimation and target localization, these strategies may require additional resources and utilize advances in waveform design and multi-sensor transmit/receive configurations. In general, serving two different purposes from the same platform and striving to attain the desired performance for both radar and communications are likely to entail an increase in system complexity.

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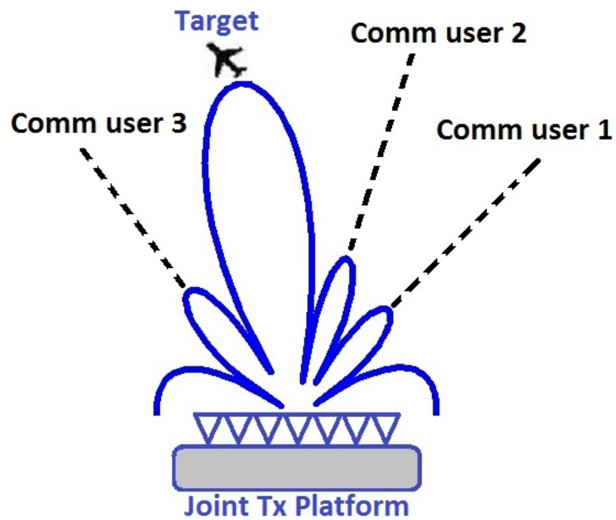


Fig. 1. Dual-function radar-communications.

The embedding of a communication signal into the radar emission for dual-functionality was reported in [15], wherein the radar transmit waveform is selected on a pulse-to-pulse basis from a bank of waveforms, each representing a communication symbol. The communication receiver deciphers the embedded information by determining which waveform was transmitted. Effective information embedding approaches have also been devised for DFRC systems with multi-sensor transmit/receive configurations [16], [17], [18], [19]. The use of time-modulated arrays to realize a dual-function array which enables performing the radar function in the mainlobe while realizing communications in the sidelobe was reported in [16]. The essence of this method is to use time-modulated array techniques to control the instantaneous pattern by using either sparse time-modulated array (STMA) or phase-only synthesis time-modulated array (POSTMA). Both STMA and POSTMA offer the ability to introduce variations in the sidelobe level (SLL) towards a certain spatial direction [16]. STMA is simple to implement by switching the transmit antennas on and off. However, because the number of antennas that need to be switched on is constant, STMA offers only a few degrees of freedom and, therefore, is not capable of achieving a large number of distinct SLLs. On the other hand, POSTMA offers enhanced capability to achieve more SLLs but requires computationally demanding nonlinear optimization. The ability to control the instantaneous pattern sidelobe and to achieve a number of distinct SLLs towards a certain direction has motivated the use of amplitude modulation (AM) to embed information as a secondary communication function to the primary radar function of the time-modulated array [16].

A multi-waveform DFRC system with multi-sensor transmit/receive configurations and bi-level sidelobe control was introduced in [17]. Motivated by the recent advances in multiple-input multiple-output (MIMO) radar and the successful exploitation of simultaneously transmitting multiple independent radar waveforms, the recently developed DFRC system enables embedding binary information via each waveform. In its simplest format, one binary bit is embedded for each radar pulse, enabling embedding of a number of binary bits that is equal to the number of employed orthogonal waveforms. The multi-waveform DFRC method uses convex optimization methods for embedding binary information and achieving a bi-level sidelobe control. That is, the communication embedding process is based on sidelobe binary amplitude shift keying (ASK). Convex optimizations have also allowed the design of transmit weight vectors satisfying constraints mandated by the radar operations while optimizing the transmit pattern. The DFRC system developed in [17], similar to the methods in [16], enables communications within the sidelobe region only. The inability to embed information within the main beam using AM or ASK based techniques is attributed to the fact that the main beam remains unchanged during the entire processing interval. For a DFRC system to be effective, the information embedding is inherently secure against eavesdroppers located in directions other than the intended communication directions [17].

A phase-modulation (PM) based method for embedding information into the radar emission under multi-sensor transmit/receive configurations was recently proposed in [19]. In order to deliver a finite number of binary bits per radar pulse, the PM based method maps the binary data into a phase symbol that belongs to a phase dictionary of an appropriate size. During each radar pulse, the PM based method embeds one phase symbol into the radar emission towards the intended communication direction. At the communication receiver, a phase detector is used to detect the embedded symbol and, subsequently, deciphers the corresponding binary sequence. Unlike the AM and multi-waveform ASK methods, the PM based method offers the ability to embed information towards communication receivers regardless of whether they are located within the sidelobe or the mainlobe. The PM based communications embedding can be coherent or noncoherent, and is suited for both directional communications and broadcasting.

In references [20] and [21], low probability of intercept (LPI) communication methods are developed by utilizing radar-embedded communications where each incident radar pulse is remodulated into different waveforms corresponding to a set of symbol constellations. Such methods achieve covertness as the communication signals are masked by the strong radar signal and similar to the ambient radar scattering (clutter). The covert communication emitter may be made purely passive by harvesting energy from the radar signals. Such operation is similar to passive radio frequency identification (RFID) tags. By implementing intra-pulse remodulations, the resulting overall data rate is on the same order of the radar pulse repetition frequency (PRF), which is typically a few kHz. In order to minimize the symbol estimation errors, a communication waveform must be sufficiently separated from other waveforms and from the clutter. A reasonable choice is that the communication signals reside in, or very close to, the passband of the incident radar illumination, yet are temporally coded in fast time to possess a manageable level of correlation with the clutter. While this approach also implements communication functionality in an operational radar platform, there are fundamental differences to the DFRC systems described in this article. A covert communication system is usually not a part of the radar system design and may not be cooperative to the radar system.

In the following sections, we highlight the different approaches for DFRC functionalities. But, first, we present some definitions implied in this work.

- *Radar and Communications Co-existence:* The co-existence of radar and communications refers to the situation where radar and communications are allowed to utilize the same bandwidth either simultaneously or non-simultaneously. The latter is achieved by following some form of bandwidth allocation/scheduling. In either case, the radar and communication systems need not be combined on a joint platform.
- *Dual-Function Radar-Communications:* A dual-function radar-communication system is equipped with a joint platform that enables performing a primary radar function and a secondary communication function simultaneously. The secondary function of the system can be performed only while the primary function is being performed.
- *Directional Radar-Embedded Communications:* In directional radar-embedded communications, the communication process is direction-dependent, i.e, the embedded information message is delivered to one or multiple communication receivers located in specific spatial directions, and it cannot be detected by receivers located in directions other than the intended receivers.
- *Radar-Embedded Broadcasting:* Radar-embedded broadcasting refers to the case when the embedded information can be received by communication receivers located in any spatial direction.
- *Coherent Radar-Embedded Communications:* In coherent radar-embedded communications, the successful detection of the embedded information requires phase synchronization between the transmitter and the receiver.

II. System Configuration and Signal Model

We consider a joint radar-communication platform equipped with M transmit antennas arranged as a uniform linear array (ULA). The radar receiver employs an array of N receive antennas with an arbitrary linear configuration. Without loss of generality, a single-element communication receiver is assumed to be located in direction θ_c, which is known to the transmitter. The bandwidth for joint radar-communication functions is denoted as B and the total transmit power budget is P_t. Let $\psi_k(t), k = 1, 2, \ldots, K$, be a set of K orthogonal waveforms, each occupying the same bandwidth B. In other words, the spectral contents of all waveforms overlap in the frequency domain. It is assumed that each waveform is normalized to have unit power, i.e., $\int_{T_p} |\psi_1(t)|^2 dt = \ldots = \int_{T_p} |\psi_K(t)|^2 dt = 1$, where T_p is the radar pulse duration and t is the fast time index. It is further assumed that the orthogonality condition $\int_{T_p} \psi_k(t)\psi_{k'}^*(t)dt = 0$ is satisfied for $k \neq k'$, where $(\cdot)^*$ stands for the complex conjugate.

Let $\mathbf{s}(t;\tau)$ be the $M \times 1$ baseband transmit signal vector during the τth radar pulse. The baseband signal at the output of the communication receiver can be expressed as

$$y_{\text{com}}(t;\tau) = \alpha_{\text{ch}}(\tau)\mathbf{a}^T(\theta_c)\mathbf{s}(t;\tau) + n(t;\tau), \tag{1}$$

where $\alpha_{\text{ch}}(\tau)$ is the channel coefficient of the received signal which summarizes the propagation environment between the transmit array and the communication receiver during the τth pulse, and $\mathbf{a}(\theta_c)$ is the steering vector of the transmit array toward the communication direction θ_c. In addition, $n(t;\tau)$ is the additive white Gaussian noise with zero mean and variance σ_c^2. Based on the manner in which the transmit signal vector $\mathbf{s}(t;\tau)$ is formed, information can be embedded into the radar emission as will be discussed in details in the following section.

III. Signaling Strategies for Radar-Embedded Communications

This section presents an overview of four signaling strategies for radar-embedded communications, i.e., the waveform diversity based method [15], the SLL AM based method [16], the SLL ASK based method [17], and the PM based method [19]. For all these methods, the waveform dictionary used at the transmitter is assumed to be known to the communication receiver.

A. Waveform Diversity Based Information Embedding

The waveform diversity scheme embeds N_b bits of information per pulse by selecting the radar waveform on a pulse-to-pulse basis from a set of $K = 2^{N_b}$ waveforms [15]. Each waveform represents a distinct N_b-bit communication symbol. The communication receiver can obtain the embedded information in each pulse by determining which waveform was transmitted. Assuming that the kth communication symbol is embedded during the τth pulse, the transmit signal vector can be expressed as $\mathbf{s}(t;\tau) = \mathbf{1}_M\psi_k(t)$, where $\mathbf{1}_M$ is the $M \times 1$ vector of 1's. This signaling strategy is illustrated in Fig. 2.

As different waveforms are employed over a coherent processing interval (CPI), the range sidelobe response at the output of the radar receive filter (typically implemented as nominally matched to the

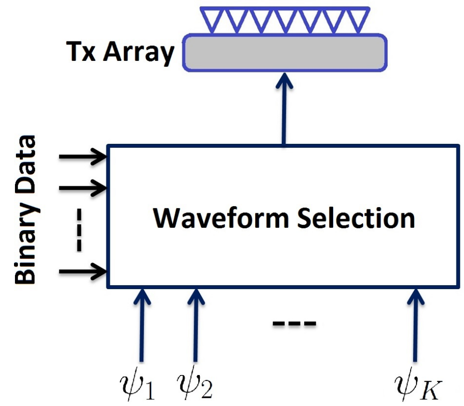


Fig. 2. Waveform diversity based signaling.

transmit waveform) can vary from pulse to pulse. This results in a loss of coherency within the CPI, leading to a degradation of the radar clutter cancellation performance. In order to maintain acceptable radar performance under the waveform diversity embedding scheme, the set of K waveforms and the associated set of K radar receive filters should be designed such that all waveform-filter pair responses are driven to be identical.

Let the K arbitrary waveforms, $\psi_k(t), k = 1, 2, \ldots, K$, be generated from binary codes of length-L chips. Assuming a sampling rate of P samples per chip, the discretized version of the kth waveform can be represented by an $LP \times 1$ vector, Ψ_k. Denote the receive filters as $\mathbf{h}_1, \mathbf{h}_2, \ldots, \mathbf{h}_K$, each of length $\tilde{L}P$ with $\tilde{L}$ being an integer multiple of L. Then, the coherency of the range sidelobes at the radar receiver can be ensured by jointly designing $\mathbf{h}_1, \mathbf{h}_2, \ldots, \mathbf{h}_K$ using a modified least squares mismatch filtering approach [15]. That is,

$$\mathbf{h}_k = (\mathbf{B}_k^H \mathbf{B}_k + \delta \mathbf{I})^{-1} \mathbf{B}_k^H \bar{\mathbf{r}}, \qquad k = 1, \ldots, K, \tag{2}$$

where $\delta \mathbf{I}$ is a diagonal loading term, $\bar{\mathbf{r}} = \sum_{k=1}^{K} \mathbf{B}_k (\mathbf{B}_k^H \mathbf{B}_k + \delta \mathbf{I})^{-1} \mathbf{B}_k^H \mathbf{e}$ with its element corresponding to the match point set to a unit value, $\mathbf{e}$ is an elementary vector of length $\tilde{L}P + LP - 1$ whose non-zero element corresponds to the match point, and the matrix $\mathbf{B}_k$ is obtained from the $(\tilde{L}P + LP - 1) \times \tilde{L}P$ convolution matrix containing the kth waveform Ψ_k by zeroing out P neighboring rows above and below the row corresponding to the match point. It is noted that the filter design approach works for low PRF radars, since it does not accommodate multiple-time-around clutter.

B. Sidelobe Amplitude-Modulation Based Communications

The essence of the sidelobe AM based method is to embed information into the radar emission via modulating the SLL towards the intended communication direction [16]. In order to satisfy the primary radar operation requirements, the radar mainlobe is kept unchanged during the entire CPI. Therefore, the sidelobe AM based method enables information delivery to a communication receiver located within the sidelobe region, but does not enable communications within the main beam of the radar. The N_b information bits are mapped into a dictionary of $K = 2^{N_b}$ amplitude symbols denoted as $\mathbb{D}_{\mathrm{AM}} = \{\Delta_1, \ldots, \Delta_K\}$. Each symbol Δ_k is represented by a specific SLL. Therefore, the implementation of this method requires the use of a single radar waveform and K distinct SLLs. This signaling strategy is illustrated in Fig. 3.

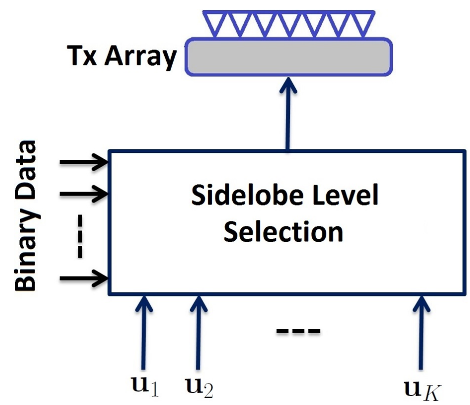


Fig. 3. Sidelobe amplitude-modulation based signaling.

Here, we summarize two approaches to achieve this number of distinct SLLs.

1) Time-Modulated Array Based Implementation: The time-modulated array based approach optimizes the average array factor within the radar integration time, which is divided into P time intervals. The average factor for a ULA is defined as [16]

$$\mathrm{AF} = \frac{1}{P}\sum_{p=1}^{P}\sum_{m=0}^{M-1} w_{m,p} \exp\left\{-j\frac{2\pi}{\lambda} d \sin\theta_0\right\}, \tag{3}$$

where $w_{m,p}$ is the complex weight associated with the mth transmit antenna during the pth time interval, λ is the wavelength, and θ_0 is the spatial direction of the main radar beam. The complex weight is defined as $w_{m,p} = \zeta_{m,p} e^{j\varpi_{m,p}}$, where $\zeta_{m,p}$ is the amplitude and $\varpi_{m,p}$ is the phase. The problem of designing the weights can be solved using one of the following two methods [16]:

i) By using an STMA, i.e., by switching the antennas on and off. This corresponds to choosing $\varpi_{m,p} = 0$

and $\zeta_{m,p} \in \{0, 1\}$. However, due to the reduced number of degrees of freedom, STMA is capable of achieving only a few distinct SLLs, thereby limiting the number of communication symbols that can be embedded.

ii) The second method utilizes POSTMA by fixing the amplitude $\zeta_{m,p} = 1$ and optimizing the AF over the phase parameters $\varpi_{m,p} \in [0, 2\pi)$. The AF optimization is carried over a total of MP variables and genetic algorithms can be used to solve the associated optimization problem.

2) Convex Optimization Based Implementation: The second approach to achieving K distinct SLLs employs convex optimization to design K transmit beamforming weight vectors [17], [22]. The SLL associated with each transmit weight vector represents one communication symbol. During each radar pulse, one of the K weight vectors is employed. While designing the K weight vectors, the radar operation mandates that the transmit power radiation pattern within the main beam of the radar remains unchanged during the entire CPI. Assuming that the radar operation takes place in a wider spatial sector $\mathbf{\Theta}$, one way to design the transmit beamforming weight vectors is to minimize the difference between the desired and actual transmit power radiation patterns under the constraints that the sidelobes be bounded by certain pre-defined levels. This weight vector design criterion is formulated as the following optimization problem [22]

$$\begin{aligned} \min_{\mathbf{u}_k} \quad & \max_{\theta} \left| |G_{\mathrm{d}}(\theta)| - \left| \mathbf{u}_k^H \mathbf{a}(\theta) \right| \right|, \quad \theta \in \mathbf{\Theta} \\ \text{subject to} \quad & \left| \mathbf{u}_k^H \mathbf{a}(\theta) \right| \leq \varepsilon, \quad \theta \in \bar{\mathbf{\Theta}}, \\ & \mathbf{u}_k^H \mathbf{a}(\theta_c) = \Delta_k, \end{aligned} \tag{4}$$

where $\mathbf{u}_k$ is the kth $M \times 1$ transmit beamforming weight vector, $G_{\mathrm{d}}(\theta)$ is the desired transmit beampattern, $\bar{\mathbf{\Theta}}$ is the sidelobe region, ε is a positive number of user's choice used for controlling the SLLs, and Δ_k corresponds to the kth SLL towards the communication direction θ_c. Optimization problems in the form of (4) are convex and can be efficiently solved using second-order cone programming.

For example, consider the case of a transmit ULA with $M = 10$ antennas spaced half wavelength apart. The main radar operation is assumed to take place within the sector $\mathbf{\Theta} = [-10^\circ\ 10^\circ]$. A single communication direction of $\theta_c = -50^\circ$ is also assumed. Four transmit beamforming weight vectors are designed by solving (4) for $\varepsilon = 0.1$. The values $\Delta_1^2 = 0.01$, $\Delta_2^2 = 0.0033$, $\Delta_3^2 = 0.0066$, and $\Delta_4^2 = 10^{-4}$ are used. Fig. 4 shows the normalized transmit power distribution patterns versus the spatial angle for the four transmit weight vectors. Clearly, the main radar beam pattern is the same for all four vectors, while the SLLs towards the communication direction are different to enable AM based information embedding.

In order to embed the kth communication symbol Δ_k during the τth pulse, the beamforming weight

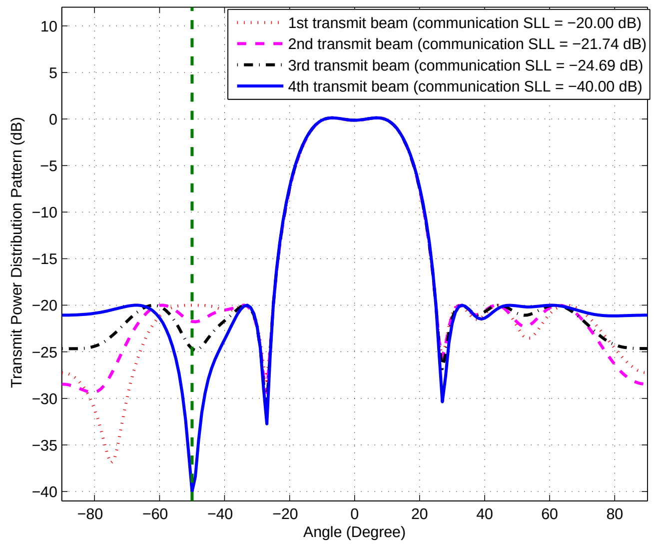


Fig. 4. Transmit power distribution versus spatial angle. (Figure adapted from [22])

vector associated with that symbol, i.e., $\mathbf{u}_k$, should be used. Then, the SLL AM based method models the transmit signal during the τth pulse as

$$\mathbf{s}(t;\tau) = \sqrt{P_{\mathrm{t}}}\mathbf{u}_k\psi(t). \tag{5}$$

Assuming a single-antenna communication receiver, the matched-filter output is given as

$$\begin{aligned} y_{\mathrm{c}}(\tau) &= \sqrt{P_{\mathrm{t}}}\alpha_{\mathrm{ch}}(\tau)\left(\mathbf{u}_k^H\mathbf{a}(\theta_c)\right) + n(\tau) \\ &= \sqrt{P_{\mathrm{t}}}\alpha_{\mathrm{ch}}(\tau)\Delta_k + n(\tau). \end{aligned} \tag{6}$$

The signal strength $\eta_{\mathrm{SLL}}(\tau) = |y_{\mathrm{c}}(\tau)|$ can be measured at the receiver. Then, by comparing $\eta_{\mathrm{SLL}}(\tau)$ to a set of $K-1$ thresholds $T_k,\ k = 1,\ldots,K-1$, that divide the K SLLs in an appropriate manner, the embedded communication symbol can be estimated. The estimated symbol can be subsequently converted to the corresponding binary sequence.

C. Multi-Waveform ASK Based Information Embedding

In this subsection, we briefly review the multi-waveform ASK method for information-embedding [17], [22]. This method employs multiple waveforms and two transmit beamforming weight vectors denoted as $\mathbf{u}_{\mathrm{H}}$ and $\mathbf{u}_{\mathrm{L}}$. This signaling strategy is illustrated in Fig. 5. The SLL associated with $\mathbf{u}_{\mathrm{H}}$ in the spatial direction of the communication receiver is assumed to be higher than the SLL associated with $\mathbf{u}_{\mathrm{L}}$. The optimization problem (4) can be used to design the aforementioned two weight vectors. In this respect, $\Delta_k = \Delta_{\mathrm{H}}$ is used while designing $\mathbf{u}_{\mathrm{H}}$ and $\Delta_k = \Delta_{\mathrm{L}}$ is used for designing $\mathbf{u}_{\mathrm{L}}$, where $\Delta_{\mathrm{H}} > \Delta_{\mathrm{L}}$.

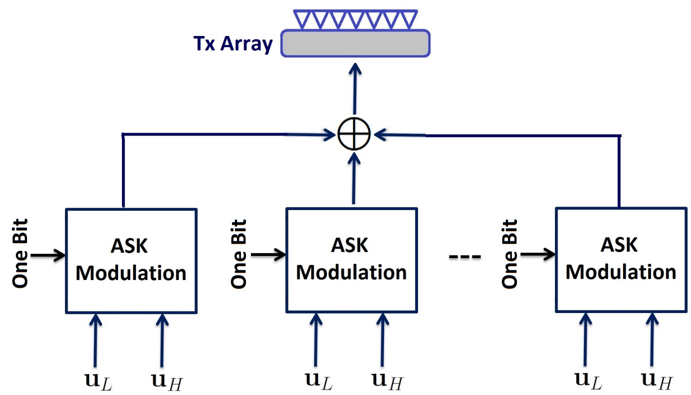


Fig. 5. Waveform diversity based signaling.

In order to embed N_b bits per radar pulse, the method requires N_b orthogonal waveforms, i.e., the number of waveforms used equals the number of transmit bits. In this respect, N_b waveforms are transmitted simultaneously where the total transmit power P_{t} is divided equally among the N_b waveforms. Every transmitted waveform is used to deliver one information bit to the communication receiver. During each radar pulse, the waveform $\psi_n(t),\ n = 1, \dots, N_b$, is radiated either via $\mathbf{u}_{\mathrm{H}}$ for $b_n(\tau) = 0$ or $\mathbf{u}_{\mathrm{L}}$ when $b_n(\tau) = 1$. The transmit signals are modeled as

$$\mathbf{s}(t;\tau) = \sqrt{\frac{P_{\mathrm{t}}}{N_b}} \sum_{n=1}^{N_b} \Big(b_n(\tau)\mathbf{u}_{\mathrm{L}}^* + (1 - b_n(\tau))\mathbf{u}_{\mathrm{H}}^* \Big) \psi_n(t). \tag{7}$$

At the communication receiver, the received signal is given by

$$\begin{aligned} y_c(t;\tau) =& \sqrt{\frac{P_{\mathrm{t}}}{N_b}} \alpha_{\mathrm{ch}}(\tau) \sum_{n=1}^{N_b} \Big(b_n(\tau)\mathbf{u}_{\mathrm{L}}^H \mathbf{a}(\theta_c) \\ & + \big(1 - b_n(\tau)\big) \mathbf{u}_{\mathrm{H}}^H \mathbf{a}(\theta_c) \Big) \psi_n(t) + \mathbf{n}_c(t;\tau) \\ =& \sqrt{\frac{P_{\mathrm{t}}}{N_b}} \alpha_{\mathrm{ch}}(\tau) \sum_{n=1}^{N_b} \Big(b_n(\tau)\Delta_{\mathrm{L}} + (1 - b_n(\tau))\Delta_{\mathrm{H}} \Big) \psi_n(t) + n_c(t;\tau). \end{aligned} \tag{8}$$

In order to extract the embedded binary bits, the received signal component associated with each transmitted orthogonal waveform is obtained using matched filtering. The nth matched filter output $y_n(\tau),\ n =$

$1, \ldots, N_b$, can be expressed as

$$y_n(\tau)=\begin{cases}\sqrt{\frac{P_{\rm t}}{N_b}}\alpha_{\rm ch}(\tau)\Delta_{\rm H}+n_n(\tau), & b_n(\tau)=0,\\ \\ \sqrt{\frac{P_{\rm t}}{N_b}}\alpha_{\rm ch}(\tau)\Delta_{\rm L}+n_n(\tau), & b_n(\tau)=1,\end{cases} \tag{9}$$

where $n_n(\tau)$ is the additive noise term which has the same statistics as that of $n_c(t;\tau)$.

Measuring the signal strength at the output of each matched filter, the transmitted bits can be extracted by performing the following ratio test

$$\hat{b}_n(\tau) = \begin{cases} 0, & \text{if } |y_n(\tau)| \geq T_0,\\ \\ 1, & \text{if } |y_n(\tau)| < T_0,\end{cases} \tag{10}$$

where T_0 is a threshold. Note that the embedding and detection of each bit are performed independently from other bits due to the orthogonality of the transmitted waveforms.

D. Phase-Modulation Based Communications

The PM based method embeds information by controlling the phase of the signal radiated towards a predetermined direction where the communication receiver is located. The N_b sequence of binary bits are mapped into a dictionary of $K = 2^{N_b}$ phase symbols denoted as $\mathbb{D}_{\rm PM} = \{\Omega_1, \ldots, \Omega_K\}$, where Ω_k denotes the kth phase symbol. This signaling strategy is illustrated in Fig. 6. Ideally, the phase symbols in $\mathbb{D}_{\rm PM}$ can be chosen to be uniformly distributed on the unit circle. For coherent communications, this method requires the use of one waveform and a band of K transmit beamforming weight vectors. For noncoherent communications, two waveforms and a bank of K pairs of transmit beamforming weight vectors are needed. To ensure that the radar operation is not affected by information embedding, all weight vectors should have the same transmit radiation pattern. The following subsection shows how to generate a population of weight vectors that have the same transmit radiation pattern.

1) Radiation Beampattern Invariance Property: Consider an $M \times 1$ principal transmit beamforming weight vector $\mathbf{w}$, which satisfies a certain desired transmit power radiation pattern as mandated by the radar operation. One way to design such a principal weight vector is by solving (4). By using the so-called radiation pattern invariance property, the principal weight vector $\mathbf{w}$ enables generating a population of 2^{M-1} weight vectors, denoted as $\mathbf{W} = \{\mathbf{w}_1, \ldots, \mathbf{w}_{2^{M-1}}\}$, of the same dimensionality, all having the same transmit power radiation pattern as that of $\mathbf{w}$ [23].

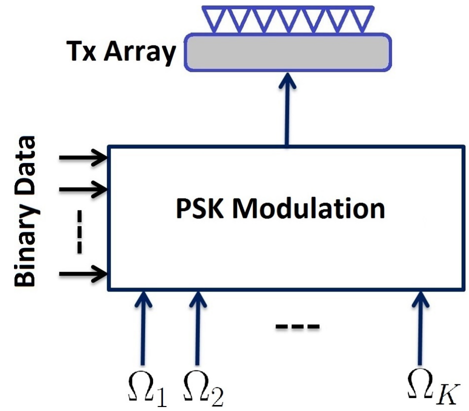


Fig. 6. Waveform diversity based signaling.

A bank of K pairs of transmit beamforming weight vectors denoted as $\{\mathbf{u}_1, \mathbf{v}_1\}, \ldots, \{\mathbf{u}_K, \mathbf{v}_K\}$, is selected from the population $\mathbf{W}$ and the kth phase symbol Ω_k is taken as the phase rotation associated with the kth pair, that is [19]

$$\Omega_k = \text{angle}\left(\frac{\mathbf{u}_k^H \mathbf{a}(\theta_c)}{\mathbf{v}_k^H \mathbf{a}(\theta_c)}\right). \tag{11}$$

For example, to embed $N_b = 2$ bits, a phase constellation of 4 symbols is assumed, e.g., $\Omega_k \in \{-\frac{\pi}{2}$, 0, $\frac{\pi}{2}$, $\pi\}$. The principal weight vector is taken as the vector associated with the red dotted curve in Fig. 4. Following the guidelines in [23], the principal vector is used to generate a population of $2^{M-1} = 512$ weight vectors which have identical transmit power radiation patterns. The weight population is used to build 256 pairs of vectors and the phase rotations associated with the communication direction $\theta_c = -50°$ for all available pairs are plotted in Fig. 7. The figure depicts that the 256 phase rotations cover the entire phase domain between $0°$ and $360°$. This enables choice of a phase-rotation dictionary of $K = 4$ symbols that are the same as the desired phase constellation. One realization for this case is indicated by the red circles in Fig. 7.

2) PM Information Embedding: Let $\psi_{\mathrm{u}}(t)$ and $\psi_{\mathrm{v}}(t)$ be two orthogonal radar waveforms of unit power each. During each radar pulse, one of the K phase symbols is embedded in order to deliver the N_b bits of information. Assuming that the kth phase symbol is embedded and the total transmit power is divided equally between the two radar waveforms, the $M \times 1$ transmit signal vector is given as

$$\mathbf{s}(t,\tau) = \sqrt{\frac{P_{\mathrm{t}}}{2}}\Big(\mathbf{u}_k^* \psi_{\mathrm{u}}(t) + \mathbf{v}_k^* \psi_{\mathrm{v}}(t)\Big), \tag{12}$$

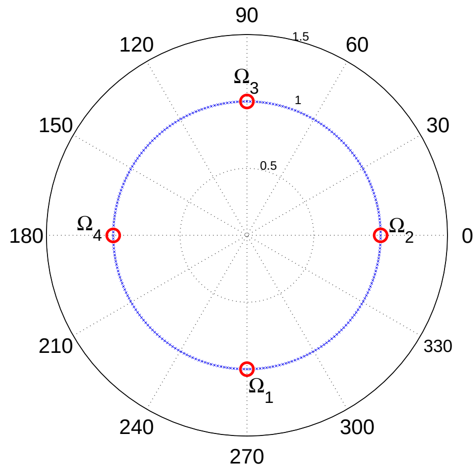


Fig. 7. Four symbol phase constellation.

Then, the output signals of the two matched filters at the communication receiver can be expressed as [19]

$$y_{\mathrm{u}}(\tau) = \sqrt{\frac{P_{\mathrm{t}}}{2}}\alpha_{\mathrm{ch}}(\tau)\Big(\mathbf{u}_k^H\mathbf{a}(\theta_c)\Big) + n_{\mathrm{u}}(\tau), \tag{13}$$

and

$$y_{\mathrm{v}}(\tau) = \sqrt{\frac{P_{\mathrm{t}}}{2}}\alpha_{\mathrm{ch}}(\tau)\Big(\mathbf{v}_k^H\mathbf{a}(\theta_c)\Big) + n_{\mathrm{v}}(\tau). \tag{14}$$

The embedded phase symbol can be estimated as

$$\hat{\varphi}(\tau) = \mathrm{angle}\left(\frac{y_{\mathrm{u}}(\tau)}{y_{\mathrm{v}}(\tau)}\right). \tag{15}$$

Then, the actual embedded binary sequence can be determined by comparing $\hat{\varphi}(\tau)$ obtained from (15) to the phase dictionary $\mathbb{D}_{\mathrm{PM}}$.

It is worth noting that both waveforms $\psi_{\mathrm{u}}(t)$ and $\psi_{\mathrm{v}}(t)$ are radiated simultaneously and, therefore, propagate through the same environment. Therefore, the initial phase at the transmit array and the reference phase at the receive array represent a common phase component in the received signals associated with both waveforms. Since the phase symbol detection depends on the difference between the phases associated with $\psi_{\mathrm{u}}(t)$ and $\psi_{\mathrm{v}}(t)$, any common phase term is canceled out. As a result, implementing a phase detector at the communication receiver does not require phase synchronization and, therefore, the communication process, in this case, is non-coherent. However, if phase synchronization between the transmit array and the communications receiver is possible and an accurate estimate of the complex channel coefficient $\alpha_{\rm ch}$ is available, then two different phase symbols can be embedded via $\psi_{\rm u}(t)$ and $\psi_{\rm v}(t)$ and detected independently. In this case, the communication process is coherent and the data rate can be doubled.

It is also worth noting that the phase rotation in (11) is direction-dependent, i.e., Ω_k is a function of the communication direction θ_c. Therefore, the actual embedded phase symbol can be detected by a communication receiver that is located in the direction θ_c, while any other communication receiver or eavesdropper located in a direction other than the intended communication direction will not be able to decode the embedded phase. Therefore, in this case, the communication process is directional.

However, in practice, there are situations where the direction of the communication receiver is either unknown or rapidly changing. Other scenarios may involve multiple intended receivers that are distributed over different directions. In such cases, information embedding in broadcast mode is recommended. One way to achieve broadcasting is by selecting $\mathbf{v}_k$ as a rotated version of $\mathbf{u}_k$, that is,

$$\mathbf{v}_k = \mathbf{u}_k e^{-j\Omega_k}, \qquad k = 1, \ldots, K. \tag{16}$$

In this case, the kth pair of transmit beamforming weight vectors will be $\{\mathbf{u}_k,\ \mathbf{u}_k e^{-j\Omega_k}\},\ k = 1, \ldots, K$. Therefore, for any arbitrary direction θ, the phase difference between the signals associated with the kth broadcast vector pair is given by

$$\varphi_k \,=\, \text{angle}\left(\frac{\mathbf{u}_k^H \mathbf{a}(\theta)}{\mathbf{u}_k^H \mathbf{a}(\theta) \cdot e^{-j\Omega_k}}\right) = \Omega_k. \tag{17}$$

Eq. (17) implies that the phase difference between the two signals is constant regardless of the direction at which the communication receiver is located. Therefore, the broadcast message can be detected from any arbitrary direction. However, it is worth noting that the transmit processing gain $G(\theta) = |\mathbf{u}_k^H \mathbf{a}(\theta)|$ is direction-dependent. Therefore, the detection performance depends on whether the communication receiver is located within the radar mainlobe or the sidelobe region. Interestingly, because of the high transmit gain within the radar main beam, receivers located in the main beam will be able to decode the embedded phase with much higher quality as compared to receivers located in the sidelobe region.

IV. Performance Comparisons and Discussions

The four signaling strategies for embedding information into the radar emission presented in the previous section share the common advantage that all resources available to the joint transmit platform, including the entire bandwidth and the total transmit power, are used to satisfy the primary radar function

TABLE I

COMPARISON BETWEEN DIFFERENT INFORMATION EMBEDDING TECHNIQUES.

	Waveform diversity [15]	Sidelobe AM [16]	Multi-waveform ASK [17]	PM method [19]
No. of waveforms	2^{N_b}	One	N_b	One pair
SSLs change from pulse to pulse	No	Yes	Yes	No
Comm. within sidelobe/mainlobe	Both	Sidelobe	Sidelobe	Both
Directional comm./broadcasting	Broadcasting	Directional	Directional	Either

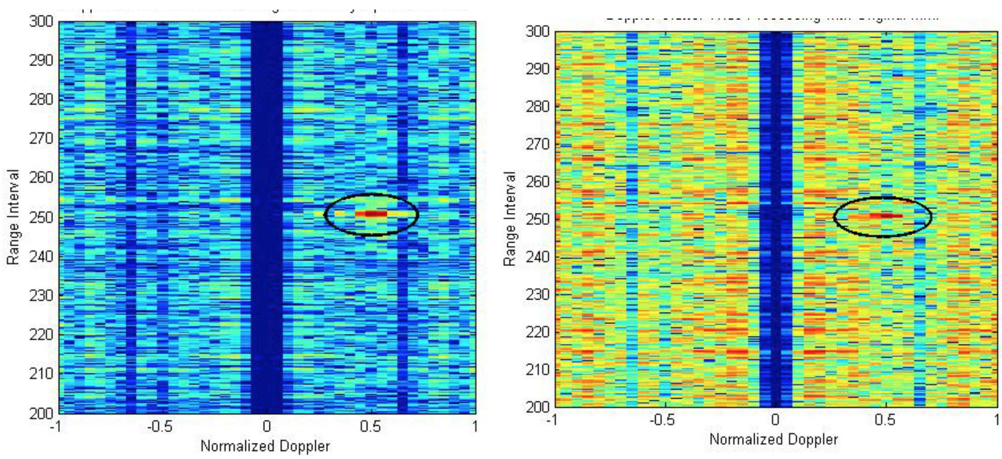
of the system. However, every method differs in terms of how it achieves the secondary communication function of the system. One of the differences is the number of waveforms needed. The waveform diversity method [15] employs a much higher number of waveforms as compared to the other three methods. Although the waveform diversity method utilizes one waveform at a time, the transmitted waveform changes from pulse to pulse. On the other hand, the three modulation based methods transmit simultaneously the same set of waveforms during all pulses. Another difference lies in the pulse to pulse SLL variation which, in turn, modulate the clutter and/or interference signal components observed by the radar receiver in the sidelobe region, especially in the vicinity of the intended communication directions. In this respect, the sidelobe AM method [16] and the multi-waveform ASK based methods introduce SLL variations while the radiation patterns associated with the waveform diversity method [15] and the PM based method [19] remain unchanged during the entire processing interval. In addition, the latter two methods enable communication delivery to communication receivers located at arbitrary directions anywhere in the angular domain while the former two methods enable communications within the sidelobe region only. Assuming that, for all methods being considered, the same number of bits N_b is embedded during each radar pulse, the comparison between the different aspects, requirements, and particularities of all presented methods can be concisely summarized as in Table I.

A few comments are in order with regards to the data rates that can be achieved using the presented information embedding methods. The data rate is given as the product of the number of bits per pulse and the PRF. Modern pulsed radar systems support PRFs in the kHz range [24], [25]. By embedding few bits per pulse, the waveform diversity method can achieve an overall data rate in the range of kbps [20]. However this method is limited to small values of N_b because the number of required waveforms grows exponentially with N_b. The sidelobe AM based method is also limited to small values of N_b because the number of required SLLs grows exponentially with N_b. Increasing the number of SLLs yields a closer constellation distance, leading to deterioration in the detection performance. For the multi-waveform ASK and the PM based methods, higher values of N_b can be used to obtain a higher data rate, especially when the number of transmit array elements is large. For all the presented methods, incorporating additional types of diversity, e.g., polarization, offers the potential for achieving even higher data rates.

We now provide some illustrative examples of the four signaling strategies for radar-embedded communications. First, we demonstrate the effectiveness of the joint receive filter design, which ensures identical waveform-filter pair responses at the radar receiver under the waveform diversity based method. A single target with 15 dB signal-to-noise ratio (SNR) is considered to be masked by clutter with a clutter-to-noise ratio of 50 dB. The clutter is modeled as a superposition of a white stationary Gaussian process and two white Gaussian processes that provide a small amount of clutter Doppler spread. The CPI consists of 20 pulses, with the waveform for each pulse randomly chosen from a set of $K = 4$ waveforms generated from binary codes of length $L = 50$ chips. The associated receive filters are obtained using the modified least squares mismatch filtering approach discussed in Section III-A. Simple clutter cancellation is employed using a wide notch, comprising the 0 and ± 1 Doppler bins, to suppress the Doppler-spread clutter [15]. Fig. 8(a) depicts the corresponding range-Doppler map in which the target is clearly visible since the clutter cancellation capability of the radar is maintained. For comparison, a range-Doppler map for the case of applying traditional receive filters is provided in Fig. 8(b). Although the target is visible, it is unlikely to be detected by the subsequent detector due to the inferior clutter suppression resulting from the lack of range sidelobe coherency.

Next, we draw some comparisons in terms of BER performance for the sidelobe AM based method, the multi-waveform ASK method, and the PM based method, all of which use the same waveforms during all pulses. For the simulation examples presented below, a uniform linear transmit array consisting of $M = 10$ antennas spaced half wavelength apart is considered, and $N_b = 2$ bits are assumed to be embedded during each radar pulse.

1) Information Delivery within the Sidelobe Region: The first example compares the BER performance for information delivery within the sidelobe region. In this respect, a long sequence of 2×10^7 bits is randomly generated. During each radar pulse, one communication symbol representing two bits is embedded, i.e., the process of embedding/detecting two bits at a time is independently repeated 10^7 times. The channel coefficient α_c is modeled as a deterministic random variable with a constant magnitude and uniformly distributed random phase. Fig. 9 shows the BERs for the three methods tested versus the SNR. It can be observed that the sidelobe AM based method has the highest BER which can be attributed



(a) (b)

Fig. 8. Clutter cancellation performance using (a) optimized receive filters and (b) traditional receive filters. (Figure adapted from [15])

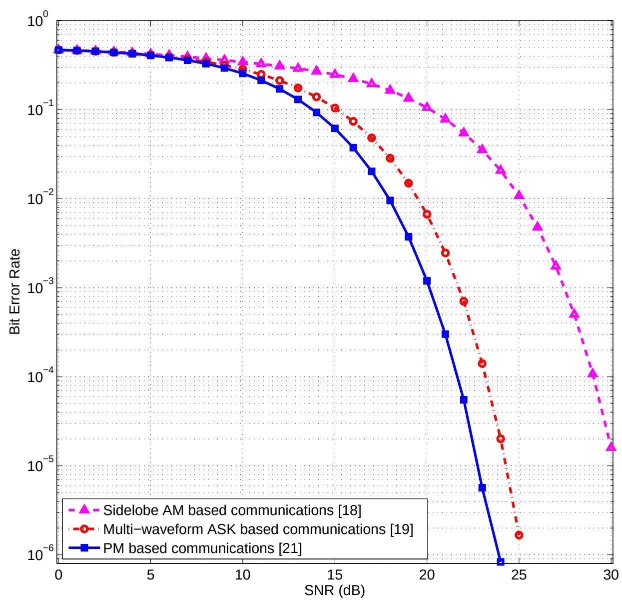


Fig. 9. BER versus SNR for sidelobe communications ($\theta_c = -50^{\circ}$).

to the small constellation separation. Also, the multi-waveform ASK method outperforms the sidelobe AM based method. Additionally, it can be observed that the PM based method achieves the best BER performance as compared to the other two aforementioned methods.

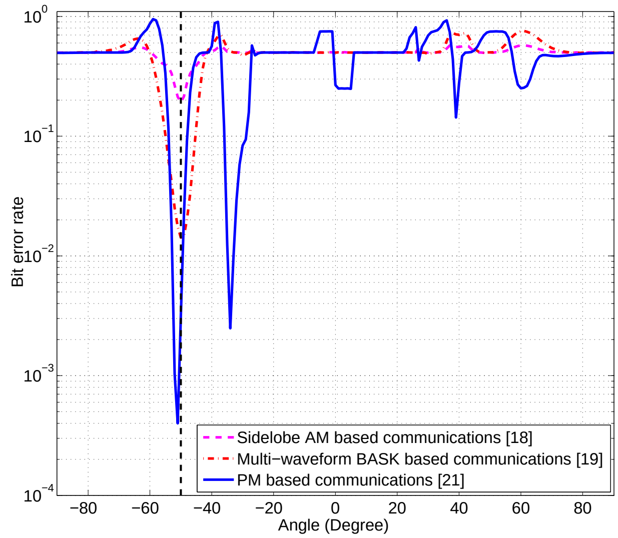


Fig. 10. BER versus angle for sidelobe communications ($\mathrm{SNR} = 10$ dB).

Fig. 10 shows the BER versus the spatial angle with the SNR fixed to 10 dB for all methods. It can be observed from the figure that the PM based method achieves a better BER towards the communication direction as compared to the methods of [16] and [17]. The figure also shows that the three methods have inherent security against information interception from directions other than the intended communication direction. The security feature against eavesdroppers located in directions other than the intended communication direction becomes more robust as the size of the constellation increases.

2) Information Delivery within Main Beam of the Transmit Power Pattern: To test the ability to communicate within the main radar beam, a communication receiver is assumed to be located in direction $\theta_c = 7.5°$. All setup parameters are chosen to be the same as the previous example. Fig. 11 shows the BER versus SNR for the sidelobe AM, the multi-waveform ASK, and the PM based methods. It can be observed from the figure that the sidelobe AM and the multi-waveform ASK based methods totally fail to decode the embedded messages simply because they are designed to communicate within sidelobe region only. On the other hand, the PM based method mirrors the curve for the same method in Fig. 9 except for a 20 dB shift on the SNR axis. This is attributed to the fact that the transmit processing gain within the main radar beam is 20 dB higher than the SLL. Therefore, communications within the main radar beam using this method can be achieved with a much higher accuracy.

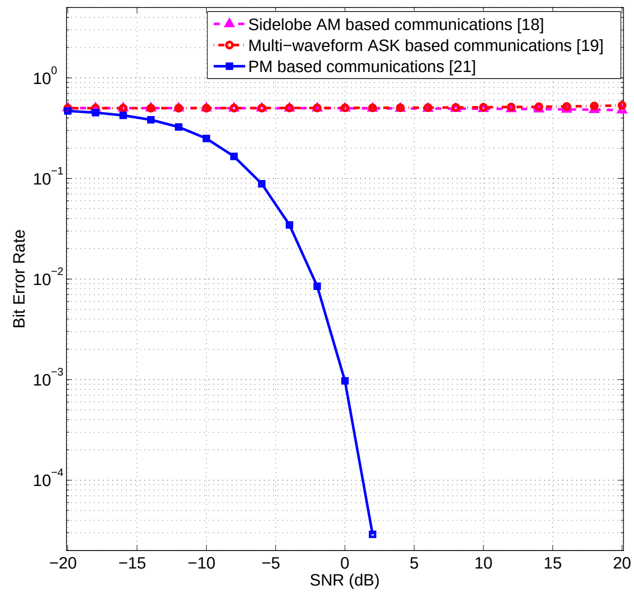


Fig. 11. BER versus SNR for main beam communications ($\theta_c = 7.5°$).

V. Conclusion

We have considered dual-function radar communication systems which are a special case of radar-communications co-existence. We have presented an overview of different strategies for radar-embedded communication signals. Such strategies are key to establishing dual-functions systems that permit simultaneous execution of both radar and communication functions from a shared platform. We have provided a balanced and complete account of existing methods and discussed their respective advantages and disadvantages.

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