# A Tutorial on Joint Radar and Communication Transmission for Vehicular Networks -Part I: Background and Fundamentals



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*(Invited Paper)*

***Abstract*—This letter is part of a three-part tutorial that focuses on joint radar-communication system design for vehicular networks. As Part I of the tutorial, this letter reviews the fundamental principles of dual-functional radar-communication from a technical perspective. First of all, we briefly introduce the background and the emerging applications of the technique. We then revisit the preliminaries of both radar and communication systems, with emphasis on the system modeling and key performance metrics. We then present the general design criteria and representative methodologies for joint radar-communication transmission in general settings, before finally focusing our study to the vehicular network which is the scope of Part II and III of this tutorial.**

***Index Terms*—Dual-functional radar-communication, spectrum sharing, joint transmission**

## I. Introduction

### A. Background

THE ever deteriorating spectrum scarcity together with key emerging applications have imposed a complete rethink of the utilization of spectral resources between radar and communication systems [1]. During the past few decades, both radar and communication systems have been deployed worldwide, spreading all over the spectrum from microwave to millimeter wave (mmWave) bands. Conventionally, most of the commercial communication systems are accommodated in the sub-6 GHz band, which are expected to achieve harmonious coexistence with a variety of existing radar systems, such as those for air traffic control (ATC), weather observation, geophysical monitoring and surveillance for defence and security [1], [2]. More recently, however, mmWave spectrum has been made accessible for 5G deployment, to leverage the large bandwidth resources available. This further leads to increasing congestion of the frequency spectrum in conjunction with the existing radar residents in this band, e.g., automotive radars and high-resolution imaging radars [1], [2]. With the aim of exploiting the under-utilized radar spectrum resources for wireless communication, both industry and academia have been looking towards radar-communication spectrum sharing, which has now attracted significant research interest and investment.

As a step beyond simply sharing the spectrum, the integration between the wireless communication and the radar sensing functionalities is highly sought after. It brings considerable gains in terms of spectral, energy, hardware and cost efficiency. This type of research, normally referred to as dual-functional radar-communication (DFRC), stands for a paradigm change where the previously competing radar and communication transmissions can be jointly designed, via the shared use of a single hardware platform and a joint signal processing framework. DFRC is envisioned as the key enabler for a number of emerging applications, including vehicular networks, indoor positioning as well as drone networks [1]. In what follows, we briefly motivate our tutorial by introducing emerging application scenarios for DFRC.

### B. Application Scenarios - The Need for DFRC

The DFRC system find applications in scenarios that require both sensing and transmission functionalities. This reduces the payload deployed on the platform, and benefits both functionalities by real-time cooperation.

*1) Vehicular network:* The next-generation vehicle-to-everything (V2X) network represents a globally growing market that is forecast to reach \$110.3 billion by 2026. Being equipped with numerous sensors and communication systems, today's vehicles have been shifted from conventional means of transportation into intelligent vehicles, where the V2X network is expected to play a significant role in ensuring the safety as well as in improving the driving and commuting experience. To address the critical performance requirements of the future connected vehicles, the V2X network should provide high-throughput communication service of Gbps transmission rate, and high-accuracy localization service at a centimeter-level resolution [3]. As a consequence, the need for DFRC design becomes indispensable as it can reduce the payload mounted on the vehicles or road infrastructures, with the potential of enhancing both the sensing and the communication performance via beneficial cooperation. To that aim, the dual-functional transmission schemes are yet to be explored under particular mobility constraints for vehicular scenarios.

*2) WiFi sensing:* Recent research interests has focused on exploiting WiFi signalling for indoor location-aware applications. These cover a large number of use cases, namely health-monitoring for the elderly, contextual awareness and security monitoring [4]. While in the outdoor environment, the GNSS is able to operate effectively with satisfactory performance, it fails to do so in the indoor scenario, which motivates the use of the WiFi-based positioning system (WPS) [4]. WPS is attractive due to its low cost and ubiquitous deployment, in which case no dedicated hardware is required. Depending

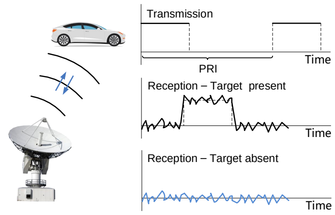


Fig. 1. Basic target probing operation for a pulsed radar.

on the sensing technique employed, WPS can be generally regarded as i) a passive radar, where the WiFi access point (AP) infers the position by processing the signal received from a cooperative user equipment (UE), or ii) an active radar, where the WiFi AP detects the human behavior by exploiting the micro-Doppler signature embedded in the reflected echoes, of other AP's transmission or via its own. Again, DFRC method tailored for WPS is key to implementing joint sensing and WiFi communication.

*3) Joint sensing and communication for unmanned aerial vehicle (UAV) network:* UAVs have been found emerging applications in critical tasks such as search and rescue, surveillance and reconnaissance and electronic countermeasure. In order to collect the useful data, UAVs are equipped with sensors, e.g., radars and cameras. To optimize the sensing performance of the UAV network, previous researches focus on the path planning/UAV placement with the aim of minimizing the sensing errors [5]. Very recently, the communication aspect of the UAV platform has been extensively studied, where the UAVs are employed as aerial base stations to provide communication services to data-demanding and disaster scenarios [6]. While the two functionalities have been individually investigated over the past few years, the joint sensing and communication design remains widely unexplored for the UAVs. This not only presents the advantage of minimizing the payload on the UAV, but also reduces its radar cross-section (RCS), which significantly improves the flexibility and covertness.

In this letter, we provide a technical overview of the DFRC transmission techniques. To begin with, we discuss the basic system model and the commonly used performance metrics for both radar and communication in Section II. Then, we introduce the design criteria and a design example in Section III, as well as a brief outlook of the radar-assisted vehicular communication. Finally, we conclude the paper in Section IV.

## II. Radar and Communication Basics

Before diving into the detailed DFRC models, let us revisit the basic principles for both radar and communication signal processing.

### A. Radar Signal Model and Performance Metrics

Fig. 1 illustrates the basic target probing operation of a pulsed radar, where a probing pulse is transmitted from the radar and hits the target. The target then reflects back an echo signal to the radar. The echo, often referred to as the target return, can be processed to extract the useful information such as the distance, velocity and azimuth angle of the target. By receiving and processing the returned echo, the radar is able to estimate the target's parameters, and thus to localize or to track the target of interest. Such a transmit-receive cycle is termed as a pulse repetition interval (PRI). As an example, let us consider a multi-antenna pulsed radar equipped with $N_t$ transmit antennas and $N_r$ receive antennas, which is detecting a point-like target in the far-field. The distance, azimuth angle and radial velocity of the target relative to the radar are given as $d$, $\theta$ and $v$, respectively. By denoting the radar probing signal as $\mathbf{s}_R(t) \in \mathbb{C}^{N_t \times 1}$, the echo signal can be mathematically expressed as

$$\mathbf{y}_R(t) = \beta e^{j2\pi f_D t}\mathbf{b}(\theta)\,\mathbf{a}^T(\theta)\,\mathbf{s}_R(t-\tau) + \mathbf{z}_R(t), \tag{1}$$

where $\beta \in \mathbb{C}$ contains both the round-trip path-loss and the RCS of the target, $f_D = \frac{2vf_c}{c}$ is the Doppler frequency, with $f_c$ and $c$ denoting the carrier frequency and the speed of the light respectively, $\mathbf{a}(\theta) \in \mathbb{C}^{N_t \times 1}$ and $\mathbf{b}(\theta) \in \mathbb{C}^{N_r \times 1}$ are the transmit and receive steering vectors, which are determined by the array geometry of the radar, $\tau = \frac{2d}{c}$ stands for the round-trip time delay, and finally $\mathbf{z}_R(t) \in \mathbb{C}^{N_r \times 1}$ denotes the additive Gaussian noise with a variance of $\sigma_R^2$.

*Remark 1:* Note that in addition to the noise, there is also interference that comes from other unwanted scatterers, e.g., trees and buildings. Such interference is referred to as "clutter", which should be suppressed to avoid affecting the processing of the target of interest. Given the page limit, we will not discuss the clutter interference in detail in this three-part tutorial.

There are two fundamental problems in radar signal processing, i.e., detection and estimation. The detection problem, in its simplest form, is to determine whether a target is present or absent. This can be modeled as a binary hypothesis testing (BHT) problem as follows

$$\mathbf{y}_R(t) = \begin{cases} \mathcal{H}_0 : \mathbf{z}_R(t), \\ \mathcal{H}_1 : \beta e^{j2\pi f_D t}\mathbf{b}(\theta)\,\mathbf{a}^T(\theta)\,\mathbf{s}_R(t-\tau) + \mathbf{z}_R(t). \end{cases} \tag{2}$$

where the hypothesis $\mathcal{H}_0$ (null hypothesis) stands for the case that the radar receives nothing but the noise, while $\mathcal{H}_1$ represents that the radar receives both the returned echo and the noise. To solve the above BHT problem, one has to design a detector and apply it to the received signal $\mathbf{y}_R(t)$. Essentially, a detector $\mathcal{T}(\cdot)$ maps $\mathbf{y}_R(t)$ to a real number, and compares this number with a pre-defined threshold $\gamma$ to determine which hypothesis to assume as true, i.e.,

$$\mathcal{T}(\mathbf{y}_R) \underset{\mathcal{H}_0}{\overset{\mathcal{H}_1}{\gtrless}} \gamma. \tag{3}$$

Based on the specific scenarios and the prior knowledge, one can employ a collection of hypothesis testing schemes to design a detector, namely the *likelihood ratio test (LRT)*, *generalized likelihood ratio test (GLRT)*, *Rao test* and *Wald test*, etc [7].

There are a number of figures-of-merit to evaluate the performance of the designed target detector. Specifically, the *detection probability* $P_D$ characterizes the probability that a target is present and is detected by the radar, i.e., the probability that $\mathcal{H}_1$ is true and the radar chooses $\mathcal{H}_1$. The *false-alarm probability* $P_{FA}$ is the probability that a target is absent but is detected at the radar, i.e., the probability that $\mathcal{H}_0$ is true and the radar chooses $\mathcal{H}_1$. They can thus be expressed in the form

$$P_D = \Pr\left(\mathcal{H}_1 \left|\mathcal{H}_1\right.\right), P_{FA} = \Pr\left(\mathcal{H}_1 \left|\mathcal{H}_0\right.\right). \tag{4}$$

It should be noted that a high $P_{FA}$ can be very harmful to the radar, as the false alarm will result in unnecessary waste of the radar's hardware and signalling resources, and will cause significant disruption to the underpinning application. Therefore, one important criterion that a target detector must follow is the so-called *Neyman-Pearson criterion* [7], which aims at maximizing $P_D$ given a minimum tolerable $P_{FA}$.

Once a target is confirmed to be present, the radar should infer the target parameters, i.e., a selection of the range ($d$), angle/direction ($\theta$) and velocity ($v$), from the noisy echo $\mathbf{y}_R\left(t\right)$. This is normally done by designing an estimator $\mathcal{F}\left(\cdot\right)$, which maps $\mathbf{y}_R\left(t\right)$ to the parameter space to obtain satisfactory estimates, and can be given by

$$\mathcal{F}\left(\mathbf{y}_R\right) = \left[\hat{d}, \hat{\theta}, \hat{v}\right]^T, \tag{5}$$

where $\hat{d}, \hat{\theta}$ and $\hat{v}$ are the estimates of their true values. In practice, the estimators are individually developed for each of the parameters. To name a few, a straightforward techninque to estimate the distance (or the time-delay) is pulse compression, i.e., to matched-filter the echo with delayed counterparts of the probing signal. For estimating the radial velocity, the radar has to firstly collect a number of pulses in order to accumulate a significant Doppler shift, and then estimates the Doppler frequency by performing the Fast Fourier Transform (FFT). Finally, the angle of arrival can be estimated via a variety of subspace methods, such as MUltiple SIgnal Classification (MUSIC) [8].

The estimation performance is usually measured by the mean squared error (MSE) between the estimated parameter and its true value. Taking the angle $\theta$ for instance, the MSE can be computed as

$$\varepsilon_\theta = \mathbb{E}\left(\left(\theta - \hat{\theta}\right)^2\right). \tag{6}$$

Nevertheless, a closed-form MSE is difficult to obtain since the distributions of both estimated parameters and their true values are not tractable in most cases. One has to therefore resort to the *Cramér-Rao lower bound (CRLB)* as an optional performance metric. Simply speaking, CRLB is the lower-bound of the variances of all the unbiased estimators [1], which states that the estimation variance should be at least as high as the inverse of the Fisher Information [7], i.e.,

$$\operatorname{var}\left(\hat{\theta}\right) \ge \frac{1}{-\mathbb{E}\left(\dfrac{\partial^2 \ln p\left(\mathbf{y}_R;\theta\right)}{\partial \theta^2}\right)} \triangleq \operatorname{CRLB}\left(\theta\right), \tag{7}$$

where the denominator is the Fisher Information with respect to $\theta$, and $p\left(\mathbf{y}_R;\theta\right)$ is the likelihood function of $\theta$ given $\mathbf{y}_R$.

[1]Note that for unbiased estimators, the MSE equals to the variance, i.e., $\varepsilon_\theta = \operatorname{var}\left(\hat{\theta}\right)$.

### B. Downlink Communication Model and Performance Metrics

In this subsection, we briefly review the basic model of multi-user multi-input multi-output (MU-MIMO) system and the corresponding performance metrics. Let us consider a MU-MIMO downlink, where an $N$-antenna base station (BS) is communicating with $K$ single-antenna UEs. The received signal vector at the UEs is known to have the form of

$$\mathbf{y}_C\left(t\right) = \mathbf{H}\mathbf{x}\left(t\right) + \mathbf{z}_C\left(t\right), \tag{8}$$

where $\mathbf{x}\left(t\right) \in \mathbb{C}^{N\times 1}$ is the transmit signal vector, $\mathbf{H} = \left[\mathbf{h}_1, \mathbf{h}_2, ..., \mathbf{h}_K\right]^T \in \mathbb{C}^{K\times N}$ is the MIMO channel matrix, with $\mathbf{h}_k \in \mathbb{C}^{N\times 1}$ being the channel vector of the $k$th user, and finally $\mathbf{z}_C\left(t\right)$ is the additive white Gaussian noise (AWGN) with a variance of $\sigma_C^2$.

For the sake of ensuring the downlink communication performance, $\mathbf{x}\left(t\right)$ should be carefully designed given both the channel $\mathbf{H}$ and the data streams intended for the $K$ UEs. By denoting the $K$ data streams as $\mathbf{s}_C\left(t\right) \in \mathbb{C}^{K\times 1}$, which is assumed to be temporally white with zero mean and unit variance, $\mathbf{x}\left(t\right)$ can be computed by

$$\mathbf{x}\left(t\right) = \mathcal{P}\left(\mathbf{s}_C\left(t\right), \mathbf{H}\right). \tag{9}$$

The above mapping operation is defined as *precoding*, which can be based on a linear or nonlinear operation in general. Linear precoding maps $\mathbf{s}_C\left(t\right) \in \mathbb{C}^{K\times 1}$ to $\mathbf{x}\left(t\right) \in \mathbb{C}^{N\times 1}$ using a linear operation defined by

$$\mathbf{x}\left(t\right) = \mathbf{W}\mathbf{s}_C\left(t\right), \tag{10}$$

where $\mathbf{W} = \left[\mathbf{w}_1, \mathbf{w}_2, ..., \mathbf{w}_K\right] \in \mathbb{C}^{N\times K}$ is a precoding matrix with $\mathbf{w}_k$ being the precoding vector for the $k$th user. To suppress or even to mitigate the multi-user interference (MUI), closed-form schemes, including *maximum ratio transmission (MRT)*, *zero-forcing (ZF)* and *minimum mean squared error (MMSE) precoders*, are popular choices for implementing $\mathbf{W}$ given their low computational complexities [9]. To further improve the communication performance under controllable constraints, optimization based linear precoding has been extensively studied in the last decade, where an important performance metric is the signal-to-interference-plus-noise ratio (SINR). For the $k$th user, this can be expressed as

$$\mathrm{SINR}_k = \frac{\left|\mathbf{h}_k^T \mathbf{w}_k\right|^2}{\sum_{i=1, i\ne k}^{K} \left|\mathbf{h}_k^T \mathbf{w}_i\right|^2 + \sigma_C^2}, \tag{11}$$

where the numerator represents the useful signal power, and the first term in the denominator represents the MUI. Accordingly, the achievable sum-rate can be obtained as

$$R = \sum\nolimits_{k=1}^{K} \log_2\left(1 + \mathrm{SINR}_k\right). \tag{12}$$

By relying on the communication metrics, one can readily formulate various optimization problems, where a classic formulation is the following SINR balancing problem [10]

$$\max_{\mathbf{w}_k} \min_{k}\ \mathrm{SINR}_k,\ s.t.\ \sum\nolimits_{k=1}^{K} \left\|\mathbf{w}_k\right\|^2 \le P_T, \tag{13}$$

where $P_T$ is the transmit power budget. The above optimization problem aims for maximizing the minimum SINR among $K$ users, and thus guarantee the level of quality-of-service

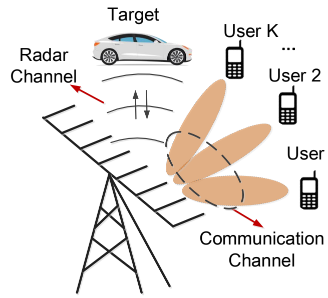


Fig. 2. Dual-functional radar-communication system.

(QoS) for all users, subject to the power budget of the BS. On the other hand, one can also minimize the power given a required SINR threshold for each user. It has been proved that both problems can be solved in polynomial time via convex optimization techniques [10].

Complementary to the above linear precoding approaches, nonlinear precoding techniques involve designing the precoded signal $\mathbf{x}(t)$ without an explicit precoding matrix $\mathbf{W}$. They are able to further boost the performance or even achieve the capacity of the Gaussian broadcast channel. Such examples include dirty paper coding (DPC), Tomlinson-Harashima precoding (THP) and vector perturbation (VP), among which the DPC is known to be capacity-achieving [9].

## III. DFRC Transmission: Merging Communication and Radar Signalling

The DFRC design aims at combining the communication and radar in a single operation, which brings forward exciting opportunities and challenges in the joint signal design. In the following, we provide insights on the DFRC system by shedding light on the criteria of joint transmission. Such designs can be realized either on a radar or a communication basis, which we term as "radar-centric design" and "communication-centric design", respectively.

### A. Radar-centric DFRC

It is straightforward to see that the pure radar waveform is unlikely to be used for communication, given that it contains no signalling information. As such, to design a radar-centric DFRC waveform is equivalent to embedding communication data into the radar waveform, without unduly degrading the sensing performance. This can be in general represented as

$$\mathbf{s}_D(t) = \mathcal{C}(\mathbf{s}_R(t), \mathbf{s}_C(t)), \tag{14}$$

where $\mathcal{C}(\cdot)$ represents the operation which embeds the communication stream $\mathbf{s}_C(t)$ into the radar waveform $\mathbf{s}_R(t)$ to generate a dual-functional waveform $\mathbf{s}_D(t)$. We list some representative data embedding approaches as follows:

- *Time-frequency embedding:* The communication symbols can be modulated on the temporal or frequency domains by using radar signals as carriers, where numerous combinations have been explored, e.g., amplitude, phase or frequency shift keying (ASK/PSK/FSK) using chirp carriers [11].
- *Code-domain embedding:* Binary/poly-phased codes are commonly used for radar, which can be modulated using direct spread spectrum sequences (DSSS), typically employed for covert communications.
- *Spatial embedding:* Radar can formulate multiple beams to serve for both purposes of target sensing and communication. Alternatively, communication symbols can also be modulated on the sidelobes of the radar beampattern, where the main-beam is dedicated to target sensing [12].

### B. Communication-centric DFRC

In general, any known communication waveform can be exploited for target sensing, even if it consists of unknown data. This is because the targets' information can always be extracted from the echoes by correlating with a known reference waveform. Nevertheless, the sensing performance critically hinges on some intrinsic features of the waveform. As such, the communication waveform needs to be carefully tailored in order to satisfy the sensing constraints. This can be mathematically represented as

$$\mathbf{s}_D(t) = \mathcal{R}(\mathbf{x}(t)), \tag{15}$$

where $\mathcal{R}(\cdot)$ stands for some radar-specific shaping constraints, for which we name a few as follows:

- *Low sidelobe level:* To minimize the interference generated from unwanted detection bins and clutter, the radar should guarantee that the sidelobes of the transmit signals (usually generated by correlating the echoes with the reference waveform), including those on the temporal (range), spectral (velocity) and spatial (angle) domains, are at a minimum level, which also corresponds to a "thumbtack-like" ambiguity function of the radar.
- *Constant-modulus transmission:* To ensure that the SINR of the reflected echo is significant enough, the radar is often required to transmit at the maximum available power budget. Typically, high-power radar transmission favors the constant-modulus waveforms to avoid signal distortion, when operating at the saturation region of the power amplifier in the transmit RF chain.
- *Clutter suppression:* As mentioned in Remark 1, the clutter interference is generated by the scattered signals from unwanted targets. Note that the strength of the clutter is also dependent on the sensing waveform, which is referred to as the "signal-dependent clutter". One may pre-suppress the clutter by carefully designing the transmit waveform.

### C. A Design Example

For further clarity, we overview a representative technique proposed by [13] for communication-centric design in this subsection. Following the MU-MIMO model in Sec. II-B, we consider an $N$-antenna DFRC BS, serving for $K$ single-antenna UEs while detecting targets. Let us rewrite (8) in the matrix form as follows

$$\mathbf{Y}_C = \mathbf{S}_C + \underbrace{(\mathbf{H}\mathbf{X} - \mathbf{S}_C)}_{\text{MUI}} + \mathbf{Z}_C, \tag{16}$$

where we consider a block of symbols in time, with length $L$. Accordingly, $\mathbf{Y}_C, \mathbf{S}_C, \mathbf{Z}_C \in \mathbb{C}^{K\times L}$, and $\mathbf{X} \in \mathbb{C}^{N\times L}$ are the matrix versions of $\mathbf{y}_C(t)$, $\mathbf{s}_C(t)$, $\mathbf{z}_C(t)$ and $\mathbf{x}(t)$, respectively. It can be seen that the second term at the right-hand side represents the MUI. If the MUI equals to zero, the $\mathbf{H}$ matrix vanishes and the channel degenerates to an AWGN channel, where the data symbols can be detected via directly demodulating $\mathbf{Y}_C$. The MUI is favorable as a least-squares (LS) cost function given its smoothness and convexity.

A typical optimization problem for DFRC that optimizes the communication performance given a well-designed radar waveform is expressed as [13]

$$\min_{\mathbf{X}} \|\mathbf{HX} - \mathbf{S}_C\|_F^2 \tag{17a}$$

$$s.t \;\; |x_{n,l}| = \sqrt{P_T/N},\ \forall n, \forall l, \tag{17b}$$

$$\|\mathrm{vec}\left(\mathbf{X} - \mathbf{X}_0\right)\|_\infty \le \varepsilon, \tag{17c}$$

where $P_T$ denotes the transmit power budget, and $\mathbf{X}_0 \in \mathbb{C}^{N\times L}$ represents a given reference radar signal matrix. We minimize the communication MUI subject to the following shaping constraints: i) Constant-modulus constraint (CMC) in (17b) and ii) similarity constraint (SC) in (17c) guaranteed by a similarity threshold $\varepsilon$ in an $L_\infty$-norm sense. As stated above, the CMC is typically required at the radar to avoid signal distortion. Moreover, the SC is enforced in the optimization to ensure that the designed DFRC signal matrix $\mathbf{X}$ possesses some useful properties of the benchmark signal $\mathbf{X}_0$, such as low sidelobe levels and robustness to the clutter interference. While problem (17) is non-convex due to the CMC involved, it can be efficiently solved via a specifically tailored branch-and-bound (BnB) algorithm proposed in [13], where a globally optimal solution can be readily obtained.

### D. Vehicular DFRC: Radar-assisted Communication

We end this section by discussing a novel line of research direction in DFRC that has recently attracted great interests within the V2X framework discussed in our introduction. That is radar-assisted communication, or communication served by radar sensing. While this research area remains widely open, it is particularly sought after in the scenarios with highly dynamic V2X communication channels.

In the future vehicular network having both capabilities of high-throughput communication and high-accuracy localization, the mmWave spectrum will play a significant role due to the large bandwidth available. Conventionally, the beam training and tracking between mmWave transceivers are implemented over a communication-only protocol. The transmitter firstly sends pilot symbols, which are used to estimate the angle at the receiver. Then, the estimated angle is fed back to the transmitter, and is employed to formulate spatial beams, such that a high-quality communication link can be established. While this method is effective in static or low-mobility scenarios, the high mobility of the vehicles together with the ultra-low latency requirements of traffic applications impose critical challenges in the beam training/tracking design. Indeed, the angle estimation accuracy can be improved by using more pilot symbols, but is at the price of increased overheads and latency, which are vital for the autonomous vehicle applications. By exploiting the side information provided by an extra radar sensor for beam tracking, the feedback loops between the transmitter and receiver could be replaced by the echo signals reflected by the vehicles, thus reducing the overheads [14]. As a step beyond, it would be more advantageous to transmit DFRC signals from a single transceiver both for vehicle sensing and communication, where a dedicated radar device is no longer needed. Some initial works on this topic can be found in [15]. Owing to the increasing interest in this challenging research area, we dedicate our focus to this topic in Part II and Part III of this tutorial.

## IV. Conclusion

In this letter, we have provided an overview on the fundamental concepts and basic principles of the DFRC design for vehicular network. Starting from the motivation and key applications, we have overviewed the basics of radar and communication signal processing, and their joint design through DFRC processing. In the sequence of this tutorial, Part II focuses on the state of the art in vehicular DFRC, and Part III focuses on a novel proposal of predictive beamforming for vehicular DFRC.

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