Synergistic design of communications and radar systems with common spectral and hardware resources is heralding a new era of efficiently utilizing a limited radio-frequency (RF) spectrum. Such a joint radar communications (JRC) model has advantages of low cost, compact size, less power consumption, spectrum sharing, improved performance, and safety due to enhanced information sharing. Today, millimeter-wave (mm-wave) communications have emerged as the preferred technology for short distance wireless links because they provide transmission bandwidth that is several gigahertz wide. This band is also promising for short-range radar applications, which benefit from the high-range resolution arising from large transmit signal bandwidths. Signal processing techniques are critical to the implementation of mm-wave JRC systems. Major challenges are joint waveform design and performance criteria that would optimally trade off between communications and radar functionalities. Novel multiple-input, multiple-output (MIMO) signal processing techniques are required because mm-wave JRC systems employ large antenna arrays. There are opportunities to exploit recent advances in cognition, compressed sensing, and machine learning to reduce required resources and dynamically allocate them with low overheads. This article provides a signal processing perspective of mm-wave JRC systems with an emphasis on waveform design.

Spectrum sharing at mm-wave
In recent years, sensing systems (e.g., radar, lidar, or sonar) that share the spectrum with wireless communications (RF, optical, or acoustical) and continue to operate without any significant performance losses have captured significant research attention [1], [2]. The interest in spectrum-sharing systems is mostly because the spectrum required by the wireless media is a scarce resource, whereas the performances of communications and remote sensing systems improve from utilizing a wider spectrum. In this article, we focus on RF spectrum sharing between radar and communications.

Several portions of frequency bands—from very high frequency to terahertz—are allocated exclusively for different
radar applications [3]. Although a large fraction of these bands remains underutilized, radars must maintain constant access to these bands for target sensing and detection as well as to obtain more spectrum to accomplish missions such as secondary surveillance, multifunction integrated RF operations, communications-enabled autonomous driving, and cognitive capabilities. Conversely, the wireless industry’s demand for spectrum continues to increase, providing new services and accommodating a massive amount of users with high data-rate requirements. The present spectrum is used very inefficiently due to its highly fragmented allocation.

Emerging wireless systems such as commercial LTE communications technology, 5G, Wi-Fi, the Internet of Things (IoT), and Citizens Broadband Radio Services have long caused spectral interference to legacy military, weather, astronomy, and aircraft surveillance radars [1], [3]. Similarly, radar signals in adjacent bands leak into spectrums allocated for communications and deteriorate the service quality. Therefore, it is essential and beneficial for radar and communications to develop strategies that simultaneously and opportunistically operate in the same spectral bands in a mutually beneficial manner.

The spectral overlap of centimeter-wave (cm-wave) radars with a number of wireless systems at the 3.5-GHz frequency band led to the 2012 U.S. President’s Council of Advisors on Science and Technology report on spectrum sharing [4], and the changes in regulations for this band became a driver for the spectrum-sharing research programs of multiple agencies [3]. Today, it is the higher end of the RF spectrum, i.e., the mm-wave, formally defined by the frequency range of 30–300 GHz, which requires concerted efforts for spectrum management because its technologies are in an early developmental stage. Increasingly, mm-wave systems [5] are the preferred technology for near-field communications because they provide transmission bandwidth that is several gigahertz wide; however, this bandwidth is currently unlicensed. This enables applications that require very large data rates, such as 5G wireless backhaul, uncompressed high-definition video, in-room gaming, intra-large-vehicle communications, intervehicular communications, indoor positioning systems, and IoT-enabled wearable technologies [6].

The amount of novel sensing systems in the mm-wave band has also grown. Although these devices typically have short ranges because of heavy attenuation by physical barriers, weather, and atmospheric absorption, they provide high range resolution resulting from the wide bandwidth. Typical mm-wave radar applications include autonomous vehicles [7], gesture recognition [8], cloud observation [9], RF identification [10], indoor localization [11], and health monitoring [12]. In the following sections, we explain the distinct features and JRC challenges of mm-wave channels.

**The mm-wave channel**

Compared to cm-wave, the channel environment for mm-wave is characterized by unique challenges that motivate the ensuing specific design constraints.

**Strong attenuation**

Compared to sub-6-GHz transmissions envisaged in 5G, mm-wave signals encounter a more complex propagation environment characterized by higher scattering, severe penetration losses, and lower diffraction. These losses result in mm-wave communication links that are near line of sight (LOS) with fewer non-LOS (NLOS) clusters and smaller coverage areas. Similarly, lower diffraction results in poorer coverage around corners. High attenuation also implies that mm-wave radars are useful only at short ranges and, as a result, multipath is a less-severe problem.

**High path loss and large arrays**

Quite naturally, mm-wave signals suffer from higher path loss (PL) for fixed transmitter (Tx) and receiver (Rx) gains. According to the Friis transmission formula, compensating for these losses while keeping the same effective antenna aperture (or increasing the gain) imposes constraints on the transceiver hardware. Because the received power is contingent on the beams of the Tx and Rx being oriented toward each other, the same aperture is accomplished by using steerable antenna arrays whose elements are spaced by at most half the wavelength ($\lambda/2$) of the transmitted signal to prevent undesirable grating lobes. This interelement spacing varies between 0.5 and 5 mm for mm-wave carriers. Such narrow spacings impact the choice of RF and intermediate frequency (IF) elements because they should fit in a limited amount of space, which makes precise mounting difficult to accomplish, e.g., in vehicular platforms.

**Wide bandwidths**

The unlicensed, wide mm-wave bandwidth enables higher data rates for communications as well as range resolution in radar. In automotive radar, this ensures the detection of distinct, informative micromotions of targets such as pedestrians and cyclists [13]. Mm-wave Rx sampling at the Nyquist rate requires expensive, high-rate analog-to-digital converters (ADCs). Large bandwidths also imply that using low-complexity algorithms in Tx and Rx processing is critical [7]. Furthermore, mm-wave channels are sparse in both time and angular dimensions—a property that is exploited for its low-complexity, low-rate reconstruction by using techniques such as compressed sensing [11], [14]. It is crucial to consider whether relevant narrow-band assumptions hold in an mm-wave application; otherwise, the signal bandwidth is very broad with respect to the center frequency, and the steering vectors become frequency dependent.

**Power consumption**

The power consumption of an ADC increases linearly with the sampling frequency. At baseband, each full-resolution ADC consumes $15\text{–}795\text{ mW}$ at 36-MHz–1.8-GHz bandwidths. In addition, the power consumed by other RF elements such as power amplifiers and data interface circuits, in conjunction with the narrow spacing between antenna elements, renders it infeasible to utilize a separate RF–IF chain for each element. Thus, a feasible multiantenna Tx/Rx structure and its beamformers should
be analog or hybrid (wherein the potential array gain is exploited without using a dedicated RF chain per antenna and phase shifter) [15] because fully digital beamforming is infeasible.

**Short coherence times**

Mm-wave environments such as indoor and vehicular communications are highly variable with typical channel-coherence times in the range of nanoseconds [5]. The reliability and coverage of dynamic mm-wave vehicular links are severely affected by the use of narrow beams. The intermittent blockage necessitates frequent beam realignment to maintain high data rates. Additionally, mm-wave radar requires a wide Doppler range to detect fast vehicles and slow pedestrians [13]. Short coherence times impact the use of feedback and waveform adaptation in many JRC designs, where the channel knowledge may be invalid or outdated when transmit waveform optimization occurs.

**Communications channel**

Consider a Tx that employs an antenna array or a single directional antenna with carrier frequency $f$ and Tx/Rx antenna gain $G_{TX}$ ($G_{RX}$). The LOS communications channel with a delay spread comprising $L-1$ delay taps is $h_c(t, f) = G_c \sum_{k=0}^{L-1} a_k e^{-j2\pi f t} e^{-j2\pi v t}$, where $G_c$ is the large-scale communications channel gain at the reception, and $a_k$ is the PL coefficient of the $k$th path with time delay $\tau_k$ and Doppler shift $v$. The free-space attenuation model yields $G_c = (G_{TX}G_{RX}\lambda^2)/(4\pi^2 \rho_0^2)$, where $\gamma$ is the PL exponent. Furthermore, $\gamma = 2$ for mm-wave LOS in outdoor urban [5] and rural scenarios [16].

**Radar channel**

The doubly selective (i.e., time- and frequency-selective) mm-wave radar channel is modeled after Tx/Rx beamforming using virtual representations obtained by uniform sampling in range dimensions [17]. Assume $L$ uniformly sampled range bins and that the $\ell$th range bin consists of a few $K$ virtual scattering centers. Each $\ell$, $k$th virtual scattering center is characterized by its distance $\rho_k$, delay $\tau_k$, velocity $v_k$, Doppler shift $\nu_k$, large-scale channel gain $G_{\ell,k}$, and small-scale fading gain $\beta_{\ell,k}$. Then, the multitarget radar channel model is $h_r(t, f) = \sum_{\ell=0}^{L-1} \sum_{k=0}^{K-1} G_{\ell,k} \beta_{\ell,k} e^{-j2\pi \rho_k t} e^{-j2\pi v_k t}$. The large-scale channel gain corresponding to the $\ell$, $k$th virtual target scattering center is $G_{\ell,k} = (\lambda^2 \sigma_{\ell,k})/(64\pi^2 \rho_k^3)$, where $\sigma_{\ell,k}$ is the corresponding scatterer’s radar cross section (RCS). The small-scale gain is assumed to be the superposition of a complex Gaussian component and a fixed LOS component leading to Rician fading. Similarly, the corresponding frequency-selective models can also include Rician fading. They capture, as a special case, the spiky model used in prior works on mm-wave communications/radar. In this case, the corresponding radar target models are approximated by the Swerling III/IV scatterers [18].

Furthermore, clustered channel models can be used to incorporate correlations and extended target scenarios, although they remain unexamined in detail. For instance, the conventional mm-wave automotive target model assumes a single, nonfluctuating (i.e., a constant RCS) scatterer based on the Swerling 0 model. This greatly simplifies the development and analysis of receive processing algorithms and tracking filters [7]. However, when the target is located within the close range of a high-resolution radar, the received signal is composed of multiple reflections from different parts of the same object. This extended target model is more appropriate for mm-wave applications and may also include a correlated RCS [13].

It is typical to assume a frequency-selective Rayleigh fading model for both communications and radar channels during the dwell time comprising $N_{CP}$ coherent processing intervals (CPIs). In radar terminology, this corresponds to the Swerling I/II target models. In each CPI with $M$ frames, the channel amplitude of each tap is considered to be constant, i.e., a block fading model is assumed. Moreover, constant velocity and quasi-stationarity conditions are imposed on the target model.

**Channel-sharing topologies**

Existing mm-wave JRC systems could be classified by the joint use of the channel [1], [23] (Figure 1). In the spectral coexistence approach, radar and communications operate as separate entities and focus on devising strategies that adjust transmit parameters and mitigate the interference adaptively for the other [3]. To this end, some amount of information exchange between the two systems, i.e., spectral cooperation, may only be allowed with minimal changes to the standardization, system hardware, and processing. In spectral co-design [1], [7], new, joint RF sensing and communications techniques are developed where a single unit is employed for both purposes while also accessing the spectrum in an opportunistic manner. New, fully adaptive, software-defined systems are attempting to integrate these systems into the same platform to minimize circuitry and maximize flexibility. Here, each Tx and Rx may have multiple antennas in a phased-array or MIMO configuration.

**JRC at mm-wave: Coexistence**

Interference management is central to the spectral coexistence of different radio systems. This typically requires sensing the state of the shared spectrum and adjusting Tx and Rx parameters so that the impact of interference is sufficiently reduced and individual system performance is enhanced. In the next section, we present figures of the merit-qualifying system performance and then discuss methodologies for mm-wave coexistence.

**Communications performance criteria**

Because the goal of communications systems is the error-free transfer of data at a high rate for a given bandwidth, commonly used performance criteria include quality-of-service (QoS) indicators such as spectral efficiency, mutual information (MI), channel capacity, pairwise error probability, bit/symbol error rates (BERs/SERs), and signal-to-interference-and-noise ratio (SINR). Given a communications signal model, the achievable spectral efficiency can be
used as a universal communications performance criterion. In practice, the achievable spectral efficiency $r$ is an upper bound, while the effective spectral efficiency $r_{\text{eff}}$ depends on the implemented Rx [e.g., minimum mean-square error (MMSE)] [24], decision feedback [25], or time-domain equalizer [26]) and is a fraction of the achievable spectral efficiency. The effective communications rate is then the product of the signal bandwidth $W$ and $r_{\text{eff}}$.

**Radar performance criteria**

By virtue of their use in both detection and estimation, radar systems lend themselves to a plethora of performance criteria depending on the particular task. Target-detection performance is characterized by probabilities of correct detection, misdetection, and false alarm. In parameter-estimation tasks, MSE, or variance in comparison to the Cramér–Rao lower bound (CRLB), is commonly considered. The CRLB defines the lower bound for estimation error variance for unbiased estimators. There are also several radar design parameters such as range/Doppler/angular resolution/coverage as well as the number of targets a radar can simultaneously resolve. In particular, the radar’s ability to discriminate in both range and velocity is completely characterized by the ambiguity function (AF) of its transmit waveform; it is obtained by correlating the waveform with its Doppler-shifted and delayed replicas.

**Interference mitigation**

The mm-wave radar and communications Tx and Rx can use all of their degrees of freedom (DoF), e.g., different antennas, frequency, coding, transmission slots, power, or polarization to mitigate or avoid mutual interference. Interference may also be caused by the leakage of signals from adjacent channels as a result of reusing identical frequencies in different locations. In general, the higher the frequency in mm-wave bands, the weaker the multipath effects. The Txs can adjust their parameters so that the level of interference is reduced at the Rx. To this end, awareness about the dynamic state of the radio spectrum and interference experienced in different locations, subbands, and

![Diagram of Coexistence and Co-Design Systems](image_url)

*FIGURE 1.* (a) The spectral coexistence system where radar and communications subsystems are independently located and access the associated radio channels such as radar target channel $h_r$, communications channel $h_c$, radar-to-communications interference $h_s$, and communications-to-radar interference $h_d$ [19]. (b) The co-design system where only Rxs are shared. In this joint multiple-access channel, the radar operates in monostatic mode, and both systems transmit different waveforms that are orthogonal in spectrum, code, or time [20]. (c) In a Tx-shared co-design, the monostatic radar functions as a communications Tx, emitting a common JRC waveform [21]. (d) A bistatic broadcast co-design with a common Tx, Rx, and joint waveform [7]. The joint waveform transmitted by the Tx vehicle bounces off targets (e.g., T1 and T2) and is received by the Rx vehicle. A variant is the in-band, full-duplex system with different waveforms but common Txs and Rxs [22]. BS: base station.
time instances is desired. This may be in the form of feedback provided by the Rxs to the Tx about the channel response and SINR. Both the Tx and Rx can be optimized such that the SINR is maximized at the Rxs for both subsystems.

**Rx techniques**

Interference mitigation may be performed only at the Rx that renders the channel state information (CSI) exchange optional. Typically, this requires multiple antennas at the Rx, a common feature at the mm-wave, and processing of the received signals in the spatial and/or temporal domains. These techniques employ the receive-array covariance matrix $\Sigma$ (or its estimate $\hat{\Sigma}$) in certain interference-canceling Rx structures. Here, the received signal space spanned by eigenvectors of $\Sigma$ is divided into two orthogonal subspaces of signal and interference plus noise. The received signal is then projected onto a subspace orthogonal to the interference-and-noise subspace to enable processing of practically interference-free signals. If the interference impinges the Rx from angles different than that of the desired signal, Rx beamforming is commonly used [23]. The beampattern design ensures high gains toward the desired signals and steers nulls toward the interference. Common solutions include minimum-variance distortion-less response, linearly constrained minimum variance, and diagonal loading [27].

Advanced interference cancellation Rxs estimate CSI, use feedback about channel response, or sense other properties of the state of the radio spectrum. These estimates are later used to cancel interference contribution from the overall received signal. The coherence time of the channels should be sufficiently long enough so that the feedback or channel estimates are not outdated during the interference cancellation process. These techniques either require knowledge of modulation schemes employed by coexisting radio systems or are applied to digital modulation methods only. A prime example of this is the successive interference cancellation method, which decodes and subtracts the strongest signal first from the overall received signals and then repeats the same procedure by extracting the next-weakest signal from the residual signal and so on [1]. In the absence of CSI, nontraditional radar interference models are used for robust communications signal decoders [28].

**Tx techniques**

Adapting Tx and optimizing transmit waveforms may be used to minimize the impact of interferences in coexistence systems. In a radar communications coexistence scenario, e.g., the optimization objective could be maximizing the SINR at each Rx while respecting the desired data rate for each communications user and target Neyman–Pearson detector performance for radar users. Designing a precoder for each Tx and/or decoders for each Rx achieves this goal by steering the interferences to different spaces than the desired signals.

One such example design in the context of MIMO communications and MIMO radar is the switched small singular-value space projection method [29] in which the interference is steered to space spanned by singular vectors corresponding to zero or negligible singular values. This method requires information exchange between the radar subsystem and communications base stations. Another example of a precoder/decoder design for interference management in radar communications coexistence is via interference alignment (IA) [30] where IA coordinates coexisting multiple Txs such that their mutual interference aligns at the Rxs and occupies only a portion of the signal space. The interference-free signal space is then used for radar and communications purposes.

**JRC at mm-wave: Co-Design**

Central to facilitating the co-design of radar and communications systems are waveform design and their optimization exploiting available DoF (e.g., spatial, temporal, and spectral polarization). The optimization is based on the system performance criteria and availability of CSI, awareness about the target scene, and the levels of unintentional or intentional interference at the Rxs.

**JRC performance criteria**

In co-design, JRC waveforms are modeled to simultaneously improve the functionalities of both subsystems with a quantifiable tradeoff. In [31], a radar round-trip delay-estimation rate is developed and coupled with the communications information rate. This radar estimation, however, is not drawn from the same class of distributions as that of communications data symbols; therefore, it provides only an approximate representation of the radar performance. However, the potential invalidity of some assumptions limits the extension of this to the estimation of other target parameters.

The mm-wave designs for single- and multiple-target scenarios in [32] and [33] suggest an interesting JRC performance criterion, which attempts to parallel the radar CRLB performance with a new, effective communications symbol MMSE criteria as a function of effective maximum-achievable communications spectral efficiency, $r_{\text{eff}}$. The MMSE communications criteria presented in this section is analogous to the MSE distortion in the rate distortion theory. Let $\text{MMSE}_{c}$ be the MMSE of a communications system with spectral efficiency $r$. Then, $\text{MMSE}_{c}$ and $r$ are related to each other through the equation $(1/N)\text{Tr}[\log(2\text{MMSE}_{c})] = -r$, where $N$ is the code length. Therefore, the effective communications distortion MSE (DMSE) that satisfies $(1/N)\text{Tr}[\log(2\text{DMSE}_{\text{eff}})] = -r_{\text{eff}} = -\delta \cdot r$ can be defined as $\text{DMSE}_{\text{eff}} \triangleq \text{MMSE}_{c}^{\delta}$, where $\delta$ is a constant fraction of communications symbols transmitted in a CPI with the channel capacity $C$. The performance tradeoff between communications and radar is quantified in terms of a weighted combination of the scalar quantities $(1/N)\text{Tr}[\log(2\text{DMSE}_{\text{eff}})]$ and $(1/Q)\text{Tr}[\log(2\text{CRLB})]$, respectively, where the log scale is used to achieve proportional fairness between the communications distortion and radar CRLB values, and $Q$ is the number of detected targets. Pareto-optimal solutions that assign weights to different design goals have also been explored [34].
MI is also a popular waveform-optimization criterion. At the radar Rx, depending on whether the communications signal reflected off the target is treated as useful energy or interference or is ignored altogether, a different MI-based criterion results. Although MI maximization enhances the characterizing capacity of a radar system, it does not maximize the probability of detection. The optimal radar signals for target characterization and detection tasks are generally different [3, 19].

Radar-centric waveform design

In this section, we first consider the appropriate radar-centric waveforms. These range from conventional signals to emerging multicarrier waveforms.

Conventional continuous-wave and modulated waveforms

A simple continuous-wave (CW) radar provides information about only Doppler velocity. To extract range information, either the frequency/phase of the CW signal is modulated or very short-duration pulses are transmitted. In practice, the well-known frequency-modulated CW (FMCW) and phase-modulated CW (PMCW) radars are used. A typical FMCW radar transmits one or multiple chirp signals wherein the frequency increases or decreases linearly in time, and the chirps reflected off the targets are captured at the Rx. A chirp bandwidth of a few gigahertz may be used to provide a range resolution of a few centimeters, e.g., a 4-GHz chirp achieves a range resolution of 3.75 cm. For PMCW radar, binary pseudorandom sequences with desirable autocorrelation/cross-correlation properties are typically used. The PMCW signal is easier to implement in hardware, and its AF has lower sidelobes than that of FMCW [7].

A general, bistatic uniform linear array (ULA) PMCW-JRC system [7] follows the topology shown in Fig. 1(d). The Tx sends \( M \) repetitions of the PMCW code of length \( L \) from each of its \( N_t \) transmit antennas. The Doppler shift and flight time for the paths are assumed to be fixed over the CPI. The reflections from \( Q \) targets that impinge on \( N_r \) receive antennas. Let \( t_r \) represent chip time (the time needed for transmitting one element of one PMCW code sequence, i.e., fast time). The Doppler shifts and flight time for every path are assumed to be fixed over a coherent transmission time \( M t_b \), where \( t_s = L c \) is the time taken to transmit one block of code, i.e., slow time. The transmit waveform takes the form

\[
x_q(t) = \sum_{m=0}^{M-1} \sum_{l=0}^{L-1} a_m e^{jk_q} e^{j\phi_q} s(t - t_{c} - m t_b),
\]

where \( i \in [1, N_i] \) and \( a_m = e^{jk_m} \) denote differential phase-shift keying (DPSK) symbols over slow time (time for sending one code sequence). The DPSK modulation is robust to constant phase shifts. Furthermore, \( s(t) \) is the elementary baseband pulse shape, \( \xi_q \in [0, \pi] \) is the binary phase code, \( e^{j(\pi - 1)k q} \) is the beamsteering weight for the \( q \)th antenna, \( k = 2\pi f / c \) is the wavenumber, and \( \beta \) is the angle between the radiating beam perpendicular to the ULA (for simplicity, we consider only azimuth and ignore common elevation angles). The Tx steers the beam in multiple transmissions from \([ -\pi / 2, \pi / 2] \), each time with angle \( \beta \). As shown in Fig. 2, the communications and radar waveform for PMCW-JRC are combined in analog hardware.

Let \( \Delta V_q^{(1)} \) be the radial-relative velocity between the Tx and \( q \)th path, where the superscript in \( (\cdot)^{(1)} \) in the formula at the beginning of this paragraph refers to the Tx-target path, and the corresponding Doppler shift is \( f_{D_1} = (\Delta V_q^{(1)}) / c \). Where \( c = 3 \times 10^8 \) m/s is the speed of light. The signal that impinges on the \( q \)th scatterer is

\[
z_{q,a}(t) = \sum_{m=0}^{M-1} \sum_{l=0}^{L-1} h_{q,a}^{(1)} e^{j\beta_q} s(t - t_{c} - m t_b - \tau_q^{(1)} - \tau_q^{(2)}) e^{j2\pi f_{D_1} t},
\]

(2)

where \( t_{c} \) and \( h_{q,a} \) are the point time delay and propagation loss for each path, respectively. We exploit the standard narrow-band assumption to express the received signal as a phase-Doppler-shifted version of the transmit signal. Assume \( \tau_q = \tau_q^{(1)} + \tau_q^{(2)} \) to be the total flight time corresponding to a bistatic range \( R_q = c \tau_q \), where the superscript in \((\cdot)^{(2)}\) in the formula presented earlier in this paragraph denotes variable dependency on the target-Rx path. Assume \( f_{c,1} = f_{D_1} + f_{D_2} \) to be the bistatic Doppler shift and \( \psi_q \) to be the angle between the \( q \)th scatterer and perpendicular line to receive the ULA. After Tx/Rx beamforming and frequency synchronization, the received signal at antenna \( p \), obtained as a superposition of these reflections, takes the form shown in (3)

\[
\hat{y}_p(t) = \sum_{q=1}^{Q} \sum_{n=1}^{N_t} h_{p,a}^{(2)} z_{q,a}(t - \tau_q^{(2)}) e^{j(2\pi f_{D_2} t)} + \tilde{N}_p(t)
\]

(3)

In (3), \( e^{j\phi_q} = e^{-j2\pi f_{D_2} (\tau_q) / c} e^{-j2\pi f_{D_2} (\tau_{c,1}^{(2)})} \) is a static phase shift, \( h_{p,a}^{(2)} \) accumulates the effect of the \( q \)th Tx-target-Rx point scatterer as well as the PL and RCS of the target, and \( \tilde{N}_p(t) \) is complex circularly symmetric white Gaussian noise with variance \( \sigma^2 \). An extended target is modeled as a cluster of points. This, combined with the superposition of reflections from the independent scatterer, renders the model in (3) applicable for extended targets. After downconversion to baseband and ignoring RCS dependency on Tx and Rx antennas, i.e., \( \sum_{n=1}^{N_t} h_{p,a}^{(2)} e^{j\phi_q} = \sum_{n=1}^{N_t} d_{q,a} \) as \( d_{q} = d_{q,1} + d_{q,2} \), the received signal is

\[
y_p(t) = \sum_{q=1}^{Q} \sum_{m=0}^{M-1} \sum_{l=0}^{L-1} d_q a_m e^{j2\pi f_{D_1} t} e^{j(\xi + \phi_q)} s(t - t_{c} - m t_b - \tau_q) + \tilde{N}_p(t), \quad p \in [1, N_r],
\]

(4)

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FIGURE 2. A simplified block diagram that shows the major steps of transmit and receive processing for a general mm-wave JRC system. In case of PMCW-JRC, the radar and communications waveforms are combined in the analog hardware before the RF stage. On the other hand, the information bits from these two subsystems are mixed digitally in OFDMA-JRC. The multiplexing of radar-only and radar-communications frames for both PMCW- and OFDMA-JRC are depicted in the transmit portion. The receive processing for both systems is largely similar. OFDMA: orthogonal frequency-division multiplexing access; LO: local oscillator.
where \( c_q = e^{-j\omega_0 q \sin(\phi)} \). Collecting the Nyquist time samples for antenna \( p \) and rearranging them accordingly to slow/fast time, we form the matrix

\[
Y_p^\text{PMCW-JRC} = \sum_{q=1}^{Q} c_q^{-1} d_q \text{Diag}([b_q^\text{c} \otimes s]^T \otimes e_q] + N_p \in \mathbb{C}^{M \times L},
\]

(5)

where vectors \( e_q = [e^{j \omega_0 q \sin(\phi)}]_{m=1}^{M} \) and \( b_q = [e^{j \omega_0 q \sin(\phi)}]_{l=1}^{L} \) collect Doppler samples in slow and fast time, respectively, \( s = [e^{j \omega_0 q \sin(\phi)}]_{l=0} \) contains \( L \) chips of code sequence, and \( P_k \) is a cyclic permutation matrix for a shift of \( k_q \)

\[
P_{k_q} = \begin{bmatrix} 0_{K_q \times L-K_q} & I_{K_q \times K_q} \\ I_{L-K_q \times L-K_q} & 0_{L-K_q \times K_q} \end{bmatrix} \in \mathbb{C}^{L \times L},
\]

(6)

where \( k_q \in \{0, \ldots, L-1\} \) is determined by the range of the \( q \)-th scatterer. If there is no delay between the Tx and Rx for all of the paths, then \( k_q = 0 \) for all \( q \) and \( P_k \) becomes identity matrix.

In a PMCW-JRC system, the communications symbols and Doppler parameters are coupled, thus leading to a nonidentifiable model. This is resolved by a multiplexing strategy through which unknown parameters in the received signal are uniquely identified. The PMCW-JRC system adopts time-division multiplexing between radar-only (X) and JRC (Xc) frames, which are transmitted for \( \mu \) and \( 1 - \mu \)% of the CPI, respectively. The value of \( \mu \) depends on the amount of prior knowledge about the target scene. For instance, when the scene is stationary, such as driving a straight path on a highway, we may not need full sensing capacity and may scale up the allocated time appropriately for communications. A coarse estimate of radar target parameters (e.g., range, angle, and Doppler) is obtained from \( Y_p^\text{PMCW-JRC} \) of radar-only frames \( X_c \), while communications symbols are extracted from the received signal samples of the \( X_c \) frame. After extracting communications symbols from \( X_c \), the residual signal is exploited to further improve the radar target estimates using low-complexity JRC superresolution algorithms [7].

**Multicarrier waveforms**

Multicarrier waveform radars provide additional DoF to deal with dense spectral use and demanding mm-wave target scenarios such as drones, low-observable objects, and many moving vehicles in automotive scenarios. Different DoF can be used in an agile manner to achieve optimal performance depending on the radar task, nature of the targets, and state of the radio spectrum. A general drawback of multicarrier radar waveforms is their time-varying envelope, leading to an increased peak-to-average-power ratio (PAPR) or peak-to-mean-envelope-power ratio, which makes it difficult to use the amplifiers efficiently when high-transmit powers are needed. However, in mm-wave radars, the transmit powers tend to be small, and surveillance ranges are short. The PAPR reduction is achieved by not allocating all of the subcarriers or by using an appropriate coding/ waveform design. Hence, the PAPR issue in mm-wave may be less severe.

Multicarrier complementary phase-coded (MCPC) waveforms [35], wherein each subcarrier is modulated by a pseudorandom code sequence of a specific length, is also a viable mm-wave JRC candidate. The MCPC design exploits DoF in spectral and code domain. In a sense, it is related to orthogonal frequency-division multiplexing (OFDM) because after each subcarrier is modulated by a code in the time domain, the subcarriers remain orthogonal without intercarrier interference. If the subcarriers are un-coded, the waveform is exactly OFDM. The intercarrier spacing in MCPC must accommodate spreading the signals in frequency due to phase codes such as Barker, P3, or P4 polyphase codes [18]. This is achieved by choosing the intercarrier spacing to be the inverse of chip duration. In OFDM, intercarrier spacing is smaller. A generalized multicarrier radar (GMR) waveform devised in [36] and [37] subsumes most of the widely used radar waveforms, such as pseudorandom frequency hopping, MCF, OFDM, and linear-step approximations of linear FM signals (as special cases). A matrix model of Txs and Rxs is developed for GMR and allows for the defining of waveforms and codes, spreading in the time and frequency domains, power allocations, and active subcarriers using a compact notation. Different waveforms are obtained by choosing the dimensions of the matrix model and filling the entries appropriately. This approach allows for relaxing the perfect orthogonality requirement, which may lead to a better resolution of target delays and Doppler velocities at the mm-wave.

**Spatial DoF and multiple waveforms**

A few different solutions use the same waveform for both subsystems but make use of radar’s spatial DoF for communications symbols. For instance, in [38], the radar array beampattern sidelobes are modulated by communications messages along the user directions. In [39], the communications symbols are represented by a different pairing of antennas and waveforms in an MIMO configuration. Spatial DoF are also useful for adaptively canceling specific users. A joint beamforming method is suggested in [40] for dual-function radar communications, which comprise MIMO radar and communications systems, assuming full-duplex transmission. The downlink communications signal is embedded into the transmit radar waveform, and uplink communications takes place when the radar is in listening mode. This necessitates accurate synchronization among the subsystems. The technique utilizes spatial diversity by enforcing the spatial signature of the uplink signals to be orthogonal to the spatial steering vectors associated with the radar target returns. The Rx beamformer employs adaptive and nonadaptive strategies to separate the desired communications signal from echoes of targets, clutter, and noise even if they impinge the array from the same direction. Other solution paths consist of finding spatial filters that mitigate in-band MIMO communications interference through optimization of the sidelobe and cross-correlation levels in
MIMO radar systems [41], [42], by exploiting coarray processing with multiple waveforms [43] and designing precoders/decoders through IA [44].

However, for mm-wave JRC systems, full-resolution ADCs at the baseband signal result in an unacceptably high power consumption. This makes it infeasible to utilize an RF chain for each antenna element, implying that the prevailing MIMO systems that employ fully digital beamforming are not practical for mm-wave systems. Thus, the benefits of using multiple waveforms for spatial mitigation in mm-wave JRC systems are yet to be carefully evaluated. Currently, a single data stream model that supports analog beamforming with frequency-flat Tx/Rx beamsteering vectors is more common [17]. The use of large antenna arrays in mm-wave suggests that one feasible JRC approach may be to simply partition the arrays for radar and communications functionalities [14].

**Communications-centric waveform design**

The most popular communications signal for mm-wave JRC because it provides a stable performance in multipath fading and relatively simple synchronization [22]. Also, frequency division in duplexing has an added advantage; unlike time-division duplexing, the former employs different bands for uplink and downlink so that the impact on the interference in radar systems is less severe. Some solutions [7], [22] also employ the related OFDM (OFDM-A) waveform for a JRC system. Although OFDM users are allocated only on time domain, OFDMA users can be differentiated using time and frequency. The latter, therefore, provides DoF in both temporal and spectral domains. Although OFDM-JRC offers high dynamic range and efficient RF processing implementation based on fast Fourier transform (FFT), it requires additional processing to suppress high sidelobes in Rx processing and to reduce the PAPR. Moreover, the OFDM cyclic prefix (CP) used to transform frequency-selective channels to multiple frequency-flat channels, leading to a simplified equalizer, may be a nuisance in the radar context. The CP may adversely affect the radar’s ability to resolve ambiguities in radar ranging. Its length depends on the number of channels, particularly the maximum excess delay that the radar signal may experience (i.e., the time difference between the first- and last-received component of the signal). For radar applications, the CP duration should be equal to or longer than the total maximum signal travel time between the radar platform and target. Other communications waveforms proposed for mm-wave automotive JRC include spread spectrum, noise OFDM, and multiple encoded waveforms [7].

**OFDM-JRC**

Consider the same bistatic scenario in Figure 1(d) that was previously analyzed for the PMCW-JRC system. The OFDM-JRC Tx (Figure 2) sends \(N_t\) OFDM symbols from \(N_t\) transmit antennas and reflections from \(Q\) targets impinging on \(N_r\) receive antennas. Assume that \(\beta\) is the angle of departure. The Doppler shift and flight time for the paths are assumed to be fixed over a CPI, i.e., \(N_s T_{sym}\), where \(T_{sym}\) is the duration of one OFDM symbol, and \(a_{nm}\) are multiplexed communications/radar DPSK on the \(n\)th carrier of the \(m\)th OFDM symbol. Let \(N_c\) be the number of subcarriers and \(\Delta f\) be the subcarriers spacing, then the joint transmit waveform in baseband neglecting the CP is

\[
x_i(t) = \sum_{m=0}^{N_c-1} \sum_{n=0}^{N_s-1} a_{nm} e^{j2\pi fn fT_n} e^{j\beta \sin(\beta)(t - i\tau - nT_{sym})/T_s} (t - nT_{sym}),
\]

where \(s(t)\) is a rectangular pulse of the width \(T_{sym}\), \(i\in[1, N_c]\), \(n\) and \(m\) are frequency and time indices, respectively, and \(f_n = n\Delta f = nT_{sym}\) [7]. The received signal at the \(p\)th Rx over a CPI is seen in (8)

\[
\tilde{y}_p(t) = \sum_{m=0}^{N_c-1} \sum_{n=0}^{N_s-1} d_{qp} a_{nm} e^{j2\pi fn fT_n} e^{j\beta \sin(\beta)(t - i\tau - nT_{sym})/T_s} (t - nT_{sym} - \tau_q) + \tilde{N}_p(t).
\]

In (8), \(\tilde{N}_p(t)\) is the additive noise on antenna \(p\). Similar to PMCW-JRC, \(d_{qp}\) denotes the PL, the phase shift caused by carrier frequency, and the RCS of the target; \(d_{qp}\) is independent of the subcarrier index due to a narrow-band assumption. Similarly, the Doppler is assumed to be identical for all of the subcarriers given a small intercarrier spacing. For notational convenience, we omit the noise in the following equation. We sample (8) at intervals \(t_s = 1/N_c \Delta f\) as

\[
\tilde{y}_p[t_s] = \sum_{m=0}^{N_c-1} \sum_{n=0}^{N_s-1} d_{qp} a_{nm} e^{j2\pi fn fT_n} s_l[(t_s - nT_{sym} - \tau_q)],
\]

where \(l\in[1, L]\), \(n\in[1, N_s]\), \(L \leq N_s\), \(d_{qp} = \Sigma_{q=1}^{N_c} d_{qp}\) as before, and \(\hat{s}_{nm} = a_{nm} e^{j2\pi fn fT_n} e^{j\beta \sin(\beta)(t - i\tau - nT_{sym})/T_s}\) contains information about range, Doppler, angle of arrival, and communications. We assume that the number of inverse FFT (IFFT) points \(N_t\) is equal to the number of fast-time samples \(L\) in each OFDM symbol. The received signal samples can be viewed as a radar data cube in spatial, spectral, and temporal domains with \(N_t\) antennas, \(N_c\) subcarriers, and \(N_s\) OFDM symbols. Let us stack the entire DPSK symbols into a matrix \(A\in\mathbb{C}^{N_c\times N_s}\) and let \(a_m = [A]_{m}\) be the communications symbols over all of the subcarriers at \(m\)th OFDM symbol time. For a given OFDM symbol, i.e., \(m\), collecting signals from all of the subcarriers across different antennas leads to the following slow-time slice of the data cube:

\[
Y_m^{OFDM-JRC} = F_N \text{Diag}(d) \Xi (-\Delta f / c) \text{Diag}(d) C \in \mathbb{C}^{N_c\times N_s},
\]

where \(m\in[1, N_c]\), \(\Xi = \Xi((-\Delta f / c)l) = [e^{-j2\pi fn fT_n} l = 1, Q]_{l=1}^{N_c \times Q} = e^{j\beta \sin(\beta)(t - i\tau - nT_{sym})/T_s} \in \mathbb{C}^{N_c \times Q}, C = e^{j\beta \sin(\beta)(t - i\tau - nT_{sym})/T_s} \in \mathbb{C}^{Q \times N_s}\), and \(d = [d_1 \ldots d_Q]\). Additionally, \(F_N = [e^{j2\pi n \phi} l = 1, N_c \times N_s]_{l=0}^{N_c - 1} = e^{j2\pi n \phi}\) denotes the \(N_c\)-point IFFT matrix. To estimate Doppler shifts, we consider a subcarrier slice of data cube (9)
where \( a_i = [A_j] \in \mathbb{C}^N\) are the DPSK symbols over slow time, i.e., \( \Xi(f_{b_i}, T_{sym}) = [e^{j2\pi f_{b_i} m T_{sym}}]_{m=1}^{Q} \).

As in PMCW-JRC, the receive processing of OFDMA-JRC is affected by the coupling of communications symbols with a radar parameter (i.e., the range in case of OFDMA-JRC). To ensure that range estimation does not suffer from using all of the subcarriers, FDM is employed (2) such that \( \mu\%\) of the OFDMA subcarriers are allocated to radar (with known \( \alpha_{n,m} \) on these subcarriers) and the rest to JRC. The rest of the OFDMA-JRC receive processing is similar to PMCW-JRC (Figure 2) [7].

Comparison of PMCW- and OFDMA-JRC

Despite the fact that OFDMA encodes radar and communications simultaneously in the entire time and space, PMCW does so in the entire frequency and space; hence, their DoF and design spaces are in different domains. Although it turns out that the receive system models of both waveforms are mathematically identical after matched filtering and retrieve all of the JRC parameters using similar superresolution algorithms [7, 45], their individual performances mimic their respective communications and radar-centric properties. For example, the AF of the bistatic PMCW-JRC inherits the low sidelobes from its parent stand-alone PMCW radar waveform, as shown in a comparison with the AF of OFDMA-JRC in Figure 3, given the same bandwidth. On the other hand, PMCW-JRC is more sensitive to the number of users while the orthogonality of waveforms in OFDMA-JRC makes the latter robust to interchannel interference. Finally, in a networked vehicle scenario, it requires less-complex infrastructure and processing to apply PMCW with predefined or stored sequences rather than using OFDM to adaptively allocate bands to each user [7, 22]. A comparison of estimation errors in the coupled parameter—range for OFDMA-JRC and Doppler for PMCW-JRC—using JRC superresolution recovery [7] is shown in Figure 4 for \( \mu = 50\%\).

Joint coding

Recently, existing mm-wave communications protocols that are embedded with codes that exhibit favorable radar AFs are garnering much attention for JRC. In particular, the 60-GHz IEEE 802.11ad standard wireless protocol has been employed with time-division multiplexing of radar-only and radar-communications frames. In general, these designs have temporal DoF (for a monostatic radar case). The IEEE 802.11ad single-carrier physical layer (SCPHY) frame consists of a short training field (STF), a channel-estimation field (CEF), a header, and a data and beamforming training field. The STF and CEF form the SCPHY preamble. The CEF contains two 512-point sequences, i.e., \( G_{u_{512}} \) and \( G_{v_{512}} \), each containing a Golay complementary pair with a length of 256, \( \{G_{u_{256}}, G_{v_{256}}\} \) and \( \{G_{u_{256}}, G_{v_{256}}\} \), respectively. A Golay pair has two sequences, i.e., \( G_a \) and \( G_b \), each of the same length \( N \) with entries \( \pm 1 \), such that the sum of their aperiodic autocorrelation functions has a peak of \( 2N \) and zero sidelobes:

\[
G_a[n]\ast G_a[-n] + G_b[n]\ast G_b[-n] = 2N\delta[n],
\]

where \( \ast \) denotes linear convolution. This property is useful for channel estimation and target detection.

By exploiting the preamble of a single SCPHY frame for radar, the existing mm-wave 802.11ad waveform simultaneously achieves a cm-level range resolution and a gigabyte/s data rate [17]. The limited-velocity-estimation performance of this waveform can be improved by using multiple fixed-length frames in which preambles are reserved for radar [17].

**FIGURE 3.** The AFs of bistatic mm-wave JRC using (a) OFDMA and (b) PMCW signals with (c) Doppler and (d) delay cuts [7].
Although this increases the radar integration duration leading to a more-accurate velocity estimation, the total preamble duration is also prolonged, causing a significant degradation in the communications data rate [33]. A joint coding scheme based on the use of sparsity-based techniques in the time domain can minimize this tradeoff between communications and radar [32]. Here, the frame lengths are varied such that their preambles (exploited as radar pulses) are placed in non-uniformly. These nonuniformly pulsing in a CPI are then used to construct a virtual block of several pulses, which increase the radar pulse-integration time and enable an enhanced-velocity-estimation performance. If the channel is sparse, the same performance can be achieved in the frequency domain by using sub-Nyquist processing [11]. In [13], the wide bandwidth of the mm-wave is exploited using a Doppler-resilient 802.11ad link to obtain very-high-resolution profiles in range and Doppler with the ability to distinguish among various automotive targets. Figure 5 shows distinct, detailed movements of each wheel of a car and the body parts of a pedestrian as detected by an 802.11ad-based Doppler-resilient short-range radar.

Carrier exploitation

Selecting active subcarriers and controlling their power levels or PAPR in an adaptive manner is also useful for interference management. To achieve high range resolution, radar systems require wide transmit bandwidths. On the other hand, communications systems often allocate the resource blocks from a certain number of subcarriers to each user based on a channel-quality indicator that satisfies their rate and system QoS requirements. Through feedback from the Rxs, spectrum sensing, databases, or other sources, the Txs of both systems can have information about the occupancy of different subcarriers, instantaneous or desired SINR levels, channel gains, and power constraints imposed by other coexisting subsystems. This awareness can be exploited by adaptively optimizing the power allocation among different subcarriers. An example of optimizing subcarrier power \( P_k \) allocations and imposing minimum-desired rate constraints on wireless communications users and maximum power constraint \( P_T \) for the radar is as follows:

\[
\begin{align*}
\text{maximize} & \quad P_D \\
\text{subject to} & \quad P_{FA} \leq \alpha, \\
& \quad \sum_{k=0}^{N-1} P_k \leq P_T, \\
& \quad \log(1 + \text{SINR}_k) \geq t_k, \forall k,
\end{align*}
\]

where \( \eta \) is the detection threshold for a likelihood ratio test using the Neyman–Pearson detection strategy with false alarm constraint \( \alpha \). Two example power allocations from the radar perspective are depicted in Figure 6. A water-filling solution [Figure 6(a)] obtained by maximizing MI between the received data and target and channel responses allocates radar power to those parts of the spectrum where the signal experiences the least attenuation and interference level is low. The second approach [Figure 6(b)] takes into account the channel gains and required SINR values at the communications subsystems, while maximizing the radar performance in the Neyman–Pearson sense for target-detection tasks.

Cognition and learning in mm-wave JRC

More-recent enabling architectures and technologies for mm-wave JRC where the system can sense, learn, and adapt to changes in the channel are discussed in the following sections.

Cognitive systems

Cognitive radars and radios sense the spectrum and exchange information to build and learn their channel states. This typically involves channel estimation and feedback on

![Figure 4](image-url) **Figure 4.** The root-mean-square error (RMSE) of the estimated range of a single target using OFDMA-JRC with respect to (a) signal-to-noise ratio (SNR) and (b) BER using half, i.e., \( \mu = 50\% \), or all of the subcarriers (full \( N_c \)) with perfect and imperfect recovery of communications symbols. The RMSE in the Doppler estimate of a single target for PMCW-JRC using all and half-frames with respect to (c) SNR and (d) BER. In both cases, JRC superresolution algorithms [7] have been employed.
FIGURE 5. The radar signatures generated from the animation models of (a) a small car and (b) a pedestrian using Doppler-resilient 802.11ad standard waveforms [13], [46]. As the targets move radially in front of the radar on the marked trajectories, the movements of the front right (FRWs), front left (FLWs), rear right (RRWs), and rear left wheels (RLWs), respectively, of the car as well as the torso, arms, and legs of the pedestrian are individually observed in (c) and (d) the range-time and (e) and (f) Doppler-time domains.
channel quality. Spectrum cartography methods, which generate a map of spectrum access in different locations and frequencies at different time instances, have been developed in this context [48]. Based on the obtained awareness, the operational parameters of Tx and Rx in each subsystem are adjusted to optimize their performance [3]. Channel-coherence times should be long enough for JRC to apply cognitive actions. Because this duration occurs in nanoseconds for mm-wave environments, compressed sensing-based solutions aid in reducing required samples for cognitive processing [11], [49].

Fast waveforms
Algorithms that develop cognitive waveforms should have low computational complexity to redesign waveforms on the fly, typically within a single CPI. This is especially important for mm-wave systems where the fast-time radar waveform can easily have a length of tens of thousands samples. In [50], waveform designs in spectrally dense environments do not exceed a quadratic complexity. In [11] and [20], the mm-wave radar based on sub-Nyquist sampling adaptively transmits in disjoint subbands, and the vacant slots are used by vehicular communications.

Machine learning
To facilitate the fast configuration of mm-wave JRC links with low latency and high efficiency, machine learning is useful for acquiring situational awareness. This involves learning the spectrum state’s evolution over time (including classifying radar target responses or other waveforms occupying the spectrum), acquiring the channel responses, identifying an underutilized spectrum, and exploiting it in an opportunistic manner. The deep-learning methods are widely applied for tasks such as target classification, automatic waveform recognition, and determining optimal antennas and RF chains [51]. Optimal policies for coexisting systems may be learned using reinforcement learning approaches such as a partially observable Markov decision process and a restless multiarm bandit [52].

Game-theoretic solutions
The interaction between radar and communications systems sharing a spectrum can be analyzed from a game theory perspective [53]. The two systems (i.e., players) form an adversarial, noncooperative game because of conflicting interests in sharing the spectrum. The game is also dynamic due to continuously evolving spectral states over time. The utility function is designed to reflect possible strategies based on the respective players’ requirements. The solutions result in a Nash or Stackelberg equilibrium, which are the game states with the property that none or one of the players can do better, respectively. In comparison to sub-6 GHz, the solution space for mm-wave is several gigahertz wide with much lower maximum transmit power.

Summary
We outlined various aspects of implementing JRC systems at mm-wave. The sheer number of mm-wave antennas and huge bandwidth pose new challenges in waveform design and Rx processing that were not seen in other bands. The dynamic and highly variable environments of mm-wave applications require continuous cognition of the mm-wave channel by both radar and communications. While there are still many open problems in this area, mm-wave JRC is a precursor to an emerging frontier of sub-mm-wave or terahertz JRC where terahertz communications would coexist with the promising
technology of low-terahertz (0.1–1-THz) automotive and imaging radars.

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