DELAY AND DOPPLER PROCESSING FOR MULTI-TARGET DETECTION WITH IEEE 802.11 OFDM SIGNALING

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ABSTRACT
This paper investigates the processing of delay and Doppler information with IEEE 802.11p OFDM signaling for multi-target detection. We study the feasibility of extending IEEE 802.11p short-range communication (DSRC) in vehicles to automotive radio detection and ranging (radar) functionality. By exploiting the unique structure of 802.11p OFDM packets over multiple subcarriers and multiple time-slots, we apply the estimation of signal parameters via rotational invariance technique (ESPRIT) for concurrent multi-target detection and range/velocity estimation. Numerical results show sub-0.2 m accuracy in range estimation and sub-0.02 m/s accuracy in velocity estimation with high probability.

Index Terms— Radar, OFDM, 802.11, multi-target detection

1. INTRODUCTION
A recent report from the National Transportation Safety Board (NTSB) mandates the implementation of frontal collision detection and avoidance systems on all new vehicles for transportation safety [1]. This requirement can be addressed primarily through forward-facing radar [1] or dedicated short-range communication (DSRC) [2], i.e., IEEE 802.11p. Current automotive radars, such as DRW automatic cruise control (ACC) or Delphi ACC, require hundreds of MHz in bandwidth. Thus, they have to operate at millimeter wave (mmWave) spectrum (∼ 77 GHz) where large open bandwidth is available [3,4]. Unfortunately, complex antenna and hardware design with high cost may limit the market penetration of these mmWave radars.

Successful integration of DSRC with radar functionality can provide substantial opportunities for automotive radar at low cost. There has been significant research on passive radar with 802.11 waveform [5–8]. Active radar technology with 802.11 packets have been investigated as well but received less attention in [9,10]. IEEE 802.11p waveforms, which only operate with 10 MHz spectrum at 5.89 GHz, might not provide sufficient resolution for automotive applications. In [11], an exhaustive search method was proposed to obtain sub-1 m ranging accuracy for a single target using 802.11p. This result was already an improvement to the OFDM radar studied in [12–14], where the range resolution of 1.61 m and the velocity resolution of 1.97 m/s were achieved with 91.1 MHz of signal bandwidth at radar carrier frequency of 24 GHz [14].

In this work, we examine algorithmic signal processing techniques for delay and Doppler information with IEEE 802.11p OFDM signaling. We first show that periodogram-based methods used in [14] does not provide sufficient range/velocity resolution with 802.11p. We then study the application of the estimation of signal parameters via rotational invariance technique (ESPRIT) for processing 802.11p packets. Numerical results show sub-0.2 m accuracy in range estimation and sub-0.02 m/s accuracy in velocity estimation using only few measurements across multiple subcarriers and time-slots.

2. MODELING RADAR SIGNALS WITH DELAY AND DOPPLER EFFECTS
We consider a monostatic radar configuration with one transmit and one receive antenna. The wireless channel between the transmitter and receiver can be modeled as a multipath channel, where each path represents the round-trip reflection from a target. The multipath baseband equivalent of a radar channel then can be written as [15]

\[ h_b(\tau, t) = \sum_i a_i(t)e^{-j2\pi f_c \tau_i(t)} \delta(\tau - \tau_i(t)) \] (1)

where \( \tau_i(t) \) is the reflected signal delay from target \( i \), \( a_i(t)e^{-j2\pi f_c \tau_i(t)} \) is the baseband time-varying gain of the reflected signal from target \( i \), and \( f_c \) is the carrier frequency. We assume that the magnitude of channel gain \( a_i(t) \) follows the radar equation [16] and thus depends on the instantaneous target \( i \)’s range. Similarly, the time-varying delay \( \tau_i(t) \) is modeled accordingly to the range of target \( i \) such that

\[ \tau_i(t) = \tau_i^0 + 2vt_i/c \] (2)

where \( \tau_i^0 = 2d_i^0/c \) is the initial delay, \( d_i^0 \) is the initial range and \( v_i \) is the relative velocity between the target \( i \) and the radar. Herein, \( v_i > 0 \) (or \( v_i < 0 \)) indicates that the target is moving away (or towards) the radar.

By taking the Fourier transform of (1), we obtain the baseband frequency response for frequency \( f \) at a given time \( t \):

\[ H_b(f, t) = \int_{-\infty}^{\infty} h_b(\tau, t)e^{-j2\pi ft}d\tau \]
\[ = \sum_i a_i(t)e^{-j2\pi (f+f_c)\tau_i(t)} \]
\[ = \sum_i a_i(t)e^{-j2\pi f_c\tau_i^0}e^{-j2\pi f c \tau_i^0}e^{-j2\pi (f+f_c)t v_i/c}. \] (3)

With OFDM in current IEEE 802.11 standards, frequency-domain channel estimations enabled by discrete Fourier transform (DFT) are typically provided. Assuming perfect synchronization, perfect band-limited signals, perfect estimation algorithms, we take the samples of the baseband frequency response at frequency spacing (subcarrier
bandwidth) $\Delta f$ and time spacing (sampling interval) $\Delta t$. With $K$ targets, the baseband frequency response at subcarrier $m$ and time-slot $n$ is given by

$$H_b[m, n] = a_0 e^{-j\theta_0} + \sum_{i=1}^{K} a_i[n] e^{-j2\pi f_i \bar{\kappa}_i} e^{-j2\pi \bar{\kappa}_i m} e^{-j\nu_i \frac{\pi}{c} n m} \;}$$

where $a_0$ is the time-invariant and frequency-invariant response due to the direct path between the radar’s transmit and receive antennas, $\theta_0$ is the constant phase shift of the direct path. For ease of presentation, we replace $\kappa_i = 2\pi \bar{\kappa}_i \Delta t_i = (4\pi d_i^2 / c) \Delta t_i$, $\nu_i = (4\pi v_i / c) \Delta t_i$ and $\bar{\kappa}_i = 2\pi f_i \bar{\kappa}_i$. Note that $\bar{\kappa}_i \ll \kappa_i$ and $\bar{\kappa}_i \ll \nu_i$.

### 3. SYSTEM PARAMETERIZATION

![symbol time 8μs an OFDM symbol (80 samples) CP training + data CP training + data](symbol.png)

$\Delta t = 0.4\text{ms}$

**Fig. 1.** Parameterizing 802.11 OFDM signaling for radar operation.

In Fig. 1, we illustrate the structure of IEEE 802.11 OFDM signaling that is used for radar operations. The 802.11 OFDM signaling is assumed to have $B = 10\text{ MHz}$ bandwidth with 64 subcarriers. The sampling time is $T = 1/B = 0.1\mu s$. Each OFDM symbol contains 16 samples for cyclic prefix (CP) and 64 samples for training and data. Of 64 subcarriers in 802.11, 48 are used for data transmission, 4 are used for training, and 12 are zeroed to reduce adjacent channel interference. Thus, we use the measurements obtained from 52 useful subcarriers in each OFDM symbol for radar functionality. We then take the measurements every 50 OFDM symbols, i.e., $\Delta t = 0.4\text{ ms}$, for Doppler processing. The setting of $\Delta t$, which has a profound effect on velocity estimation, will be explained shortly.

### 3.1. Ambiguity

#### 3.1.1. Range Ambiguity

The baseband frequency response (1) over $N$ subcarriers at a given time-slot $\bar{n}$ can be organized into

$$H_b[m, \bar{n}] = a_0 e^{-j\theta_0} + \sum_{i=1}^{K} a_i[\bar{n}] e^{-j2\pi f_i \bar{\kappa}_i} e^{-j\nu_i \frac{\pi}{c} \bar{n}} (\kappa_i + \bar{\kappa}_i \bar{n}) m \text{ },$$

where $a_0$ is the time-invariant and frequency-invariant response due to the direct path between the radar’s transmit and receive antennas, $\theta_0$ is the constant phase shift of the direct path. For ease of presentation, we replace $\kappa_i = 2\pi \bar{\kappa}_i \Delta t_i = (4\pi d_i^2 / c) \Delta t_i$, $\nu_i = (4\pi v_i / c) \Delta t_i$ and $\bar{\kappa}_i = 2\pi f_i \bar{\kappa}_i$. Note that $\bar{\kappa}_i \ll \kappa_i$ and $\bar{\kappa}_i \ll \nu_i$.

From (5), $H_b[m, \bar{n}]$ is the summation of complex-valued sinusoidal signals at angular frequencies $-\kappa_i + \bar{\kappa}_i \bar{n}_i$. Since $\kappa_i + \bar{\kappa}_i \bar{n}_i = (4\pi \Delta t / c) (d_i^2 + \Delta \bar{n} v_i)$, the range of target $i$ at time-slot $\bar{n}$ can be deduced from the estimation of $\kappa_i$ and $\bar{\kappa}_i$.

To detect an arbitrary target $i$ at range $d_i[\bar{n}] \triangleq d_i^2 + \Delta \bar{n} v_i$ without ambiguity, $\kappa_i + \bar{\kappa}_i \bar{n}_i$ must comply

$$0 \leq \kappa_i + \bar{\kappa}_i \bar{n}_i \leq 2\pi \quad \Rightarrow \quad d_i[\bar{n}] \leq \frac{c}{2\Delta t} = \frac{cN}{2B} \text{ } (6)$$

where $B$ is the bandwidth. Since the signal delay must be nonnegative, the unambiguous range interval is defined as $[0, 2\pi]$.

Effectively, the detection range without ambiguity depends on the subcarrier spacing $\Delta t$. For a IEEE 802.11p signal with 10 MHz of bandwidth divided into 64 subcarriers, the detection range is 960 m, which is well above existing automotive radars on the market, such as TRW’s 77 GHz ACC system and Delphi’s 76 GHz ACC system.

#### 3.1.2. Velocity Ambiguity

To detect a target using its Doppler shift, the baseband frequency response (1) is observed over $T > 1$ time-slots

$$H_b[m, n] = a_0 e^{-j\theta_0} + \sum_{i=1}^{K} a_i[n] e^{-j2\pi f_i \bar{\kappa}_i} e^{-j\nu_i \frac{\pi}{c} n \bar{n}} \text{ },$$

where $a_0$ is the time-invariant and frequency-invariant response due to the direct path between the radar’s transmit and receive antennas, $\theta_0$ is the constant phase shift of the direct path. For ease of presentation, we replace $\kappa_i = 2\pi \bar{\kappa}_i \Delta t_i = (4\pi d_i^2 / c) \Delta t_i$, $\nu_i = (4\pi v_i / c) \Delta t_i$ and $\bar{\kappa}_i = 2\pi f_i \bar{\kappa}_i$. Note that $\bar{\kappa}_i \ll \kappa_i$ and $\bar{\kappa}_i \ll \nu_i$.

At a given subcarrier $\bar{m}$, $H_b[\bar{m}, n]$ is a summation of complex-valued sinusoid signals at angular frequencies $-(\nu_i + \bar{\kappa}_i \bar{n}_i)$. Since $\nu_i + \bar{\kappa}_i \bar{n}_i = (4\pi \Delta t / c) (f_i + \bar{m} \bar{n}_i)$, the estimation of the frequency $\nu_i + \bar{\kappa}_i \bar{n}_i$ also provides the estimation of target $i$’s velocity.

To estimate $\nu_i$ without ambiguity, $\nu_i + \bar{\kappa}_i \bar{n}_i$ must comply

$$-\pi \leq \nu_i + \bar{\kappa}_i \bar{n}_i \leq \pi \quad \Rightarrow \quad -\frac{c}{4f_i \Delta t} \leq \nu_i \leq \frac{c}{4f_i \Delta t} \text{ } (7)$$

Effectively, the maximum detection velocity depends on the carrier frequency and the sampling interval. With the carrier frequency of IEEE 802.11p at $f_c = 5.89\text{ GHz}$ and a sampling interval $\Delta t = 0.4\text{ ms}$, it is possible to detect and estimate a target’s velocity within $[-32, +32] \text{ m/s}$ or $[-72, +72] \text{ mph}$.

### 3.2. Resolution with Periodogram-based Methods

A straightforward approach to estimate the frequency of a sinusoid is to examine the power spectral density of a sample vector. However, classical periodogram-based methods, such as the DFT, are only capable of resolving spectral lines separated by more than $1/N$ cycles per sampling interval [17]. Therefore, $2\pi / N$ is the spectral resolution limit of the DFT method. In [14], the authors utilized the DFT/IDFT of the OFDM-based radar signal to compute the periodogram and estimate the target range and velocity. Under the consideration of IEEE 802.11p signaling in this paper, the same approach would resolve as follows:

- **Range resolution:** $\Delta d = \inf_{i,j} \| d_i - d_j \| \geq \frac{2\pi}{4\pi N \Delta f / c} = \frac{c}{2B}$, which is dependent on the spectrum bandwidth $B$. With $B = 10\text{ MHz}$, the range resolution is 15 m. The closest detectable target is also at 15 m, which is clearly insufficient for automotive applications.

- **Velocity resolution:** $\Delta v = \inf_{i,j} \| \nu_i - \nu_j \| \geq \frac{c}{2f_c T \Delta t}$, which is dependent on the carrier frequency and the observation time $T \Delta t$. If the automotive radar requires information update every 50 ms, the velocity resolution will be $\approx 0.25 \text{ m/s}$. Note that a smaller $\Delta t$ would increase the maximum detectable velocity. However, $T$ must be increased accordingly to maintain the same resolution.

### 4. ESPRIT FOR TARGET DETECTION AND ESTIMATION WITH 802.11P

#### 4.1. ESPRIT

In this section, we briefly revisit the ESPRIT method for estimation of line spectra. Line spectral estimation deals with the signals from
a sample vector of length \(N\) containing \(K\) sinusoids:
\[
y[t] = x[t] + e[t]; \quad x[t] = \sum_{i=1}^{K} \alpha_i e^{j(\omega_i t + \theta_i)} \quad (9)
\]
where \(e[t]\) is the zero-mean AWGN with variance \(\sigma^2\).

Let \(\mathbf{y}[t] = [y[t], y[t-1], \ldots, y[t-L+1]]^T\) as a sample vector of length-\(L\) for some \(L > K\). The covariance matrix model of the data can be readily derived as \(\mathbf{R} \triangleq \mathbb{E}[\mathbf{y}[t] \mathbf{y}[t]^*]\), whose eigensstructure contains complete information on the frequencies \(\omega_i\)'s [17]. Perform eigen-decomposition of \(\mathbf{R}\) and let \(\mathbf{S} = [\mathbf{s}_1, \ldots, \mathbf{s}_K]\) be the matrix containing \(K\) orthonormal eigenvectors associated with the \(K\) largest eigenvalues of \(\mathbf{R}\). Let \(\mathbf{S}_1 = [\mathbf{I}_{L-1} 0]\) and \(\mathbf{S}_2 = [0 \mathbf{I}_{L-1}]\). The ESPRIT method estimate the frequencies \(\omega_i\) as 
\[
\hat{\omega}_i = \arg\max \left| \mathbf{S}_1^* \mathbf{S}_1 \right| = \frac{1}{N} \sum_{j=0}^{N-1} y[j] \mathbf{S}_1^* \mathbf{S}_1 e^{-j2\pi j N}, \quad i = 1, \ldots, K
\]

Once the frequencies are estimated as \(\hat{\omega}_i\), the amplitudes \(\alpha_i\)'s and phases \(\theta_i\)'s (replaced by \(\hat{\beta}_i = \alpha_i e^{j\theta_i}\)) can be estimated by the least square method, whose closed-form solution is
\[
\hat{\beta} = (\mathbf{W}^* \mathbf{W})^{-1} \mathbf{W}^* \mathbf{y} \quad (10)
\]
where
\[
\mathbf{W} = \begin{bmatrix}
e^{j\hat{\omega}_1} & \cdots & e^{j\hat{\omega}_K} \\
\vdots & \ddots & \vdots \\
e^{j\hat{\omega}_1} & \cdots & e^{j\hat{\omega}_K} 
\end{bmatrix}.
\]

### 4.2. Range and Velocity Estimation using ESPRIT

#### 4.2.1. Range Estimation with Reflected Signal Delays (Method I-A)

The measured channels in (5) at a given time-slot \(n\) across the \(N\) subcarriers resemble a composition of multiple frequencies at \(\kappa_1, \ldots, \kappa_K\) and the direct path \(a_0 e^{-j\theta_0}\). At first, we assume the number of targets is known a priori. Hence, ESPRIT can be readily to estimate \(\kappa_1, \ldots, \kappa_K\) and the target ranges accordingly. We note that the covariance matrix \(\mathbf{R}\) has to be constructed from sample vectors of length-\(L\), where \(L\) is set to be \(K + 2\) to account for all targets and the direct path. The estimated frequency closest to 0, which corresponds to the direct path, is then discarded.

If the number of targets \(K\) is an unknown with a deterministic upper bound, the length of the sample vectors must be set higher than this upper bound. The number of eigenvalues of the resulting covariance matrix \(\mathbf{R}\) whose values exceed the noise floor then indicates the number of targets. The process of estimating the target ranges follows accordingly.

#### 4.2.2. Velocity Estimation with Doppler Frequencies (Method I-B)

The channel response at a particular subcarrier \(m\) can be measured over multiple time-slots. With \(T\) measurements, the channel response vector in (7) is a summation of multiple frequencies at \(\nu_i\), \(\ldots, \nu_K\) and the direct path \(a_0 e^{-j\theta_0}\). Therefore, ESPRIT can be used for estimating the frequencies \(\nu_i\), \(\ldots, \nu_K\) and hence the velocities of the \(K\) targets. Due to the time-varying ranges of the targets, the amplitudes of the sinusoids \(a_i[n]\) given in (7) may change during the observing interval \(T \Delta \nu_i\). However, it is arguably true that \(a_i[n]\) remains constant with a short observing time (\(\sim 50\) ms).

### 4.3. Target Detection and Estimation using Phase Shifts

In the previous section, the target identification and range-velocity estimation are performed with the measurements either at one time-slot across multiple subcarriers or at one subcarrier across multiple time-slots. In this section, we look at the processing of signal measurements across multiple time-slots and multiple subcarriers.

#### 4.3.1. Velocity Estimation using Phase Shifts (Method II-A)

The measurements at two consecutive time-slots across the \(N\) subcarriers are given by
\[
H_v[m, n] = a_0 e^{-j\theta_0} + \sum_{i=1}^{K} a_i[n] e^{-j\theta_i} e^{-j\nu_i (n+1)} e^{-j(\kappa_i + n \kappa_i)}
\]  
\[
H_v[m, n+1] = a_0 e^{-j\theta_0} + \sum_{i=1}^{K} a_i[n] e^{-j\theta_i} e^{-j\nu_i (n+1)}
\times e^{-j(\kappa_i + (n+1) \kappa_i)} m, \quad m = 1, \ldots, N (12)
\]

The phase shift between the two corresponding sinusoidal signals is of interest, since it reveals the velocity of the target. By estimating this phase shift, amounted as \(\nu_i\), it is possible to find the velocity of target \(i\) using only two time-slots.

We note that Method II-A is deployed in conjunction with Method I-A. Denote \(\kappa_i\)'s as the estimated frequencies using Method I-A. The amplitudes and phases of the sinusoids can be estimated using the least square solution (10). Denote \(\beta[n] = [a_0 e^{-j\theta_0}, a_1 e^{-j\theta_1} e^{-j\nu_1 \kappa_1}, \ldots, a_K e^{-j\theta_K} e^{-j\nu_K \kappa_K}]^T\), \(\mathbf{h}_n = [H_v[0, n], \ldots, H_v[N, n]]^T\), and \(\mathbf{W}_n = [\mathbf{w}_0, \ldots, \mathbf{w}_K]\) where \(\mathbf{w}_i = [e^{-j\kappa_i}, \ldots, e^{-j\kappa_i N}]^T\). Then, we have the estimate of \(\beta[n]\) as
\[
\hat{\beta}[n] = (\mathbf{W}_n^* \mathbf{W}_n)^{-1} \mathbf{W}_n^* \mathbf{h}_n. \quad (13)
\]

The phase shift of sinusoidal signal \(i\) can be estimated as
\[
\hat{\nu}_i = \angle \left( \frac{\hat{\beta}[n+1]}{\hat{\beta}[n]} \right), \quad i = 1, \ldots, K (14)
\]

which would provide the estimated target velocity. Note that Method II-A requires the targets be resolved with distinctive ranges using Method I-A first. Method II-A then enables the matching (range, velocity) for each target.

#### 4.3.2. Range Estimation using Phase Shifts (Method II-B)

We can take a similar approach to identify and estimate the targets’ ranges by calculating the phase shift between the sinusoids using the two sinusoidal waveforms corresponding to target \(i\) can be used to estimate the target ranges. The angular frequency \(\hat{\nu}_i\) can be first estimated by the ESPRIT algorithm as in Method I-B, while the amplitudes and phases of the ranges are estimated by the least square method. The phase shift \(\hat{\kappa}_i\) can be estimated in a similar fashion as in (14).

It is noted that the range estimation using the estimated phase shift requires the observation of the sinusoidal signals over multiple time-slots. For instance, with \(T = 128\), the observing time is approximately 50 ms. During this time interval, the range of target \(i\) has been perturbed by an amount of \(T \Delta \nu_i\). However, the phase shift estimation using the two sinusoidal signals does not take into account this perturbation and only provides the mean of target \(i\)'s range during this interval. To yield a more accurate range estimation at the end of \(T\) time-slots, the estimated range in Method II-B should be compensated by an amount of \(T \Delta \nu_i / 2\).

### 5. SIMULATION RESULTS

This section presents numerical results to demonstrate the detection and estimation capability of automotive radar with IEEE 802.11p packets. Fig. 2 illustrates the root mean square (RMS) error in range estimation of a one target with variable radar cross section (RCS) values. The target is assumed to travel at velocity 10 m/s relatively to the radar. For stronger targets (\(\sigma \geq 0.1 \text{ m}^2\)), both methods
This paper exploits the use of 802.11p OFDM packets for radar application. By observing the OFDM packets over multiple subcarriers and frequencies, delay and Doppler information can be extracted for ranging and velocity estimation. Numerical results show that ESPRIT can enable sub-0.2 m accuracy in range estimation and sub-0.02 m/s accuracy in velocity estimation with only 10 MHz bandwidth at 5.89 GHz. In addition, the resolvability of the radar can be improved at least threefold.

6. CONCLUSION

This paper exploits the use of 802.11p OFDM packets for radar application. By observing the OFDM packets over multiple subcarriers and frequencies, delay and Doppler information can be extracted for ranging and velocity estimation. Numerical results show that ESPRIT can enable sub-0.2 m accuracy in range estimation and sub-0.02 m/s accuracy in velocity estimation with only 10 MHz bandwidth at 5.89 GHz. In addition, the resolvability of the radar can be improved at least threefold.
7. REFERENCES


