



High Resolution Automotive Radar Imaging Integrating Doppler Division Multiple Access and Frequency Agility

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Abstract: Millimeter-wave (MMW) radar is one of the indispensable sensors for advanced driver assistance systems (ADASs) and autonomous driving. This paper adopts inter-pulse frequency agility (FA) to synthesize a broadband signal, achieving broadband imaging without increasing the sampling rate. Meanwhile, given that Doppler division multiple access (DDMA) has higher transmission efficiency than time division multiple access (TDMA) and is widely used in automotive radar, this paper proposes a signal transmission framework that integrates DDMA and FA. Simulation results demonstrate that the proposed DDMA FA multiple-input multiple-output (DDMA-FA-MIMO) processing architecture can yield 30 to 40 times more detection points for a single target compared to traditional methods, preliminarily achieving target contour imaging.

Keywords: automotive radar; radar imaging; stepped frequency radar; 4D radar; DDMA-FA-MIMO

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1 Introduction

Advanced driver assistance systems (ADAS) have emerged as standard equipment in most automobiles, while autonomous driving technology is advancing rapidly^[1]. These systems rely on diverse sensors, including radar, light detection and ranging (LiDAR), and cameras, to assist the driver during driving and parking. Notably, millimeter-wave (MMW) radar exhibits significant advantages in environmental adaptability compared to other sensors like LiDAR and cameras, thus playing a pivotal role in automotive autonomous and assisted driving applications^[2]. Furthermore, in addition to measuring crucial parameters like the distance, speed, azimuth and elevation of targets, current 4D radars also can provide the capability of high-resolution point cloud imaging^[3].

However, the density of point clouds in 4D radar images is often sparse, and the superior angular resolution of LiDAR poses significant challenges to MMW radar. Given the limited aperture size, it is commonly perceived that inferior angular

resolution may hinder MMW radar's potential to play a significant role in future autonomous driving applications. Notably, Tesla's assisted driving solutions once excluded MMW radar from their systems, though it has been reintroduced in its latest iterations^[4].

The indispensability of MMW radar in automotive driving solutions hinges significantly on its ability to notably enhance resolution. This has emerged as a prominent research focus in the field. To enhance the resolution of MMW radar, extensive research focused on high-resolution and high-precision direction-of-arrival (DOA) has been conducted. Early approaches to DOA estimation relied on digital beamforming (DBF), whose angular resolution is limited by the physical array aperture, i.e., the Rayleigh limit^[5]. Increasing the antenna aperture is undoubtedly one of the most direct ways. However, considering the costs and installation difficulties, increasing the physical array aperture may not always be the most feasible solution. Fortunately, multiple-input multiple-output (MIMO) technology^[6–8] offers the potential to create a large virtual aperture without increasing costs dramatically. Specifically, time-division multiplexing (TDM) MIMO activates only one transmit antenna element at a time and establishes signal orthogonality in the time domain. Consequently, most commer-

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cial automotive MIMO radar devices on the market, such as Texas Instruments' (TI) AWR1843 and AWR2243^[9], and NXP Semiconductors' TEF82xx and SAF85XX^[10], are based on TDM MIMO. These devices further support multi-chip cascading to enhance the aperture. For instance, the Continental ARS540 achieves an angular resolution of 1.2° through cascading four chips^[11]. However, it is noteworthy that a resolution of around 1° is still insufficient, resulting in sparse point clouds that are unable to form a distinct target contour to effectively support autonomous driving. The sparsity of these point clouds poses challenges in accurately perceiving and identifying objects in the environment, which is crucial for safe and reliable autonomous driving. Similarly, synthetic aperture radar (SAR)^[12-14] technologies can also increase the aperture by integrating received signals from diverse positions. However, motion errors introduce significant phase errors into received signals, leading to severe degradation in system performance. Additionally, the efficacy of SAR in forward-looking applications is considerably constrained.

Resolution can also be improved by employing super-resolution algorithms, such as subspace projection methods, including multiple signal classification (MUSIC)^[15], estimation of signal parameters via rotational invariance techniques (ESPRIT)^[16] and their extensions^[17-18], maximum likelihood methods (MLMs)^[19], iterative adaptive approach for amplitude and phase estimation (IAA-APES)^[20], compressed sensing (CS)^[21-24], and neural network-based approaches^[25]. Overall, these algorithms suffer from high computational complexity to varying degrees. Operations such as matrix inversion and eigenvalue decomposition pose significant challenges for real-time processing in automotive radar systems, especially under cost constraints. Furthermore, the slow convergence speed of IAA-APES and the parameter sensitivity of CS-based algorithms may influence their effectiveness.

For rapidly moving vehicles, the constantly evolving surrounding environment poses challenges to traditional super-resolution algorithms. Moreover, in automotive radar scenarios, targets are often extended targets rather than the ideal point ones encountered in typical DOA estimation scenarios^[26]. This implies that the number of targets is unknown and can change rapidly. Even worse, the equivalent discrete scattering points within an extended target can exhibit significant variations in reflection intensity at any given moment. These characteristics introduce practical challenges for the aforementioned algorithms. For instance, subspace projection methods often require model selection procedures to estimate the number of targets, while the sparsity assumption of CS-based approaches may not hold true in such scenarios^[27].

Therefore, it is imperative to examine how we can empower MMW radar to genuinely cater to the demands of autonomous driving. LiDAR excels in spatial scanning with narrow beams and low sidelobes, enabling precise imaging of target edges and contours. In contrast, MMW radar faces difficulties in

achieving distinct imaging of target contours due to its wider beams and relatively higher sidelobe levels. For example, Fig.1 illustrates a common scenario in real-world automotive radar applications. The radar is mounted on Car 1, and two vehicles travel alongside each other with a specific distance in front. Assuming each vehicle's width is 1.5 m and the gap between them is set to 1 m, the angular values formed by the reflections from the edges of each vehicle are denoted by θ_1 , θ_2 , θ_3 , and θ_4 , respectively.

Given the distance R is 20 m, the angular values are: $\theta_1 = -2.15^\circ$, $\theta_2 = 2.15^\circ$, $\theta_3 = 3.58^\circ$, and $\theta_4 = 7.83^\circ$. Although θ_3 is over 1° larger than θ_2 , distinguishing between a single target and two separate targets using a high-resolution 4D radar with 0.5° angular resolution can be extremely challenging. This is because even a 4D radar of such high resolution is incapable of capturing all scattering points from the target vehicles, only detecting a limited number of points, which subsequently hinders accurate scene comprehension. This also demonstrates the significance of MMW radar in target contour imaging.

Advanced radar systems with higher resolution and denser point clouds are needed to meet the requirements of autonomous driving applications. The enhancement of range resolution has been greatly overlooked in previous studies, despite its potential to help form high-density point cloud output. As is well known, range resolution is inversely proportional to the radar signal bandwidth. To improve range resolution, one direct method is to use a large signal bandwidth. However, directly transmitting large bandwidth signals necessitates a higher sampling rate. Therefore, large bandwidth signals are only used in short-range radar applications, while in medium- or long-range detection radar, the signal bandwidth is typically 100 MHz or 150 MHz, and the corresponding range resolution is 1.5 m or 1 m. If a signal with a 4 GHz bandwidth supported by 77 GHz commercial devices is transmitted, the range resolution can be improved by tens of times, and the number of point clouds formed will also increase by tens of times. Instead of directly transmitting a large bandwidth signal, using frequency agility transmission provides an alternative solution^[28-31].

In addition, TDMA suffers from transmission energy loss due to the fact that only one antenna is active at any moment, and therefore the Doppler division multiple access (DDMA) approach is proposed^[32-33]. Instead of achieving waveform di-

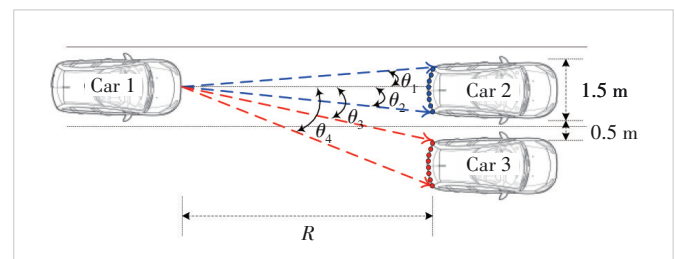


Figure 1. Automotive radar target detection

versity in the time domain, DDMA enables the design of orthogonal waveforms in the Doppler domain and the transmitted waveforms are modulated by multiple sets of linear phase codes. Thus, the transmitted waveforms encoded by DDMA are completely separated in the Doppler domain with various Doppler offsets. An integrated TDMA-DDMA-MIMO framework is proposed in Ref. [34], which combines the advantages of both.

In this paper, a high-resolution automotive radar imaging algorithm that integrates DDMA and frequency agility (FA) is proposed. This approach achieves high resolution in the range dimension, while utilizing DDMA to enhance emission efficiency, enabling high-density point cloud output. Moreover, most of the calculations involved can be completed using FFT, making it highly practical for engineering implementation. The main contributions of this work are summarized as follows:

- 1) An integrated DDMA-FA-MIMO framework is designed with a simple implementation of orthogonal waveform transmission, which achieves high range resolution with low sampling requirements and high emission efficiency.
- 2) A detailed analysis is provided regarding the coupling relationship between range and velocity, and a method for simultaneously measuring range and velocity parameters is given, which also takes into account DDMA- and FA-related target detection issues.
- 3) Based on the frequency-agile and DDMA transmission framework, a detailed analysis is provided on the performance characteristics of clutter echoes in the range-Doppler matrix, as well as the principles of achieving high range resolution.
- 4) Extensive practical experiments with Texas Instruments (TI) devices are carried out to further confirm and evaluate the effectiveness of the proposed approach.

2 Signal Model

This paper presents a high-resolution approach using a 77 GHz automotive radar system applying a frequency-modulated continuous wave (FMCW) form. We assume a MIMO antenna system composed of N transmitting antennas and L identical receiving antennas, and M pulses will be transmitted in one coherent processing interval (CPI).

In the proposed DDMA-FA-MIMO FMCW radar transmission scheme, an FMCW signal with bandwidth B and duration t_{sub} is transmitted for each pulse, but with a different modulated phase. For the n -th transmitter and m -th pulse, the transmitted signal can be represented as:

$$s_T^n(t, m) = G_n \cdot p_n(m) \cdot \text{rect}\left(\frac{t - t_{\text{sub}}/2}{t_{\text{sub}}}\right) \exp(j2\pi(f_0 + m \cdot \Delta f)t) \cdot \exp(j\pi\mu t^2) \quad (1),$$

where t denotes the time within each pulse, G_n is the gain of the n -th transmitter, f_0 denotes the starting frequency of the first pulse, Δf represents the frequency step between consecu-

tive emitted pulses, t_{sub} is the pulse width, B is the bandwidth of each pulse, m denotes the m -th pulse where $m \in [0, 1, \dots, M-1]$, n denotes the n -th transmitter where $n \in [1, 2, \dots, N]$, and μ represents the chirp rate of linear frequency modulation (LFM) signals. Especially, the vector $p_n(m)$ behaves as a phase modulator related to the pre-defined n -th transmitter to ensure MIMO waveform orthogonality. The signal transmission scheme is depicted in Fig.2.

As shown in Fig. 2, $\varphi_2, \dots, \varphi_N$ are the modulated phases for transmitters and $\varphi_1 = 0$. More generally, for MIMO systems with N transmitters, the phase modulation matrix in discrete form can be expressed as:

$$P = [p_1 \cdots p_n \cdots p_N] = \begin{bmatrix} 1 & 1 & 1 \\ 1 & e^{j\varphi_n} & e^{j\varphi_N} \\ \vdots & \vdots & \vdots \\ 1 & e^{j(N-1)\varphi_n} & e^{j(N-1)\varphi_N} \end{bmatrix} \quad (2).$$

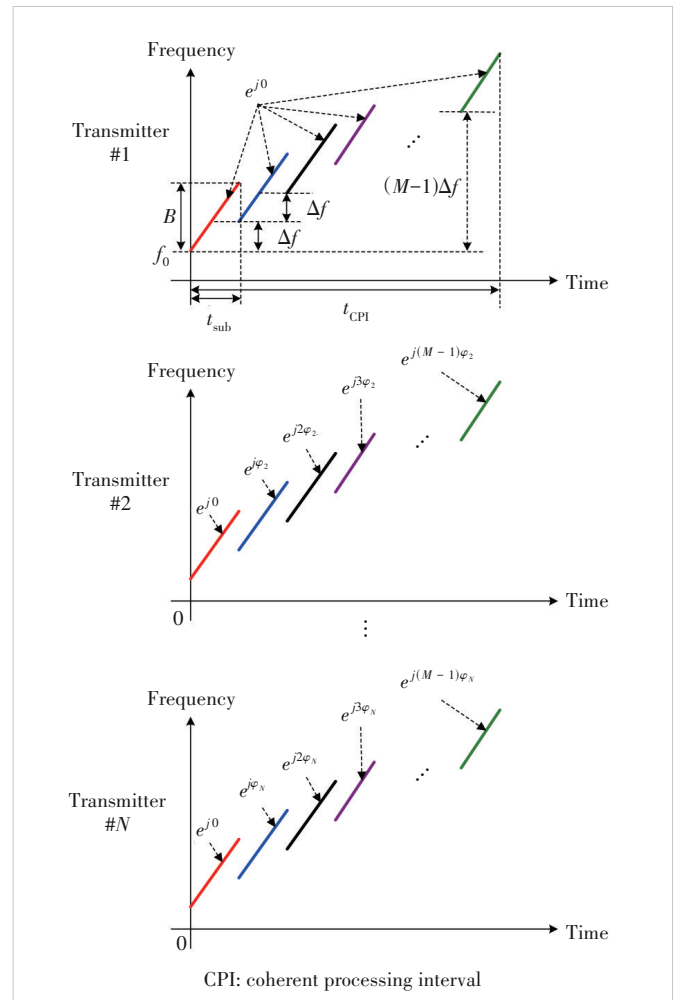


Figure 2. Signal transmission scheme integrating DDMA and FA

For the m -th pulse, the modulated phase vector is the m -th row in matrix $\mathbf{P}$, and the direction of the beam is determined by the currently modulated phase vector. DDMA can be regarded as a time-division phased array radar, and beams are directed towards different directions for different pulses.

In the following, it is assumed that the consecutive sequences do not overlap and that there is no gap between their transmissions. The signal received by the l -th receiver reflected from a moving point-like target located at θ can be written as:

$$s_R^{nl}(t, m, \theta) = \alpha \cdot G_l \cdot s_T^n(t - \tau_{nl}(t), m) \quad (3),$$

where $\tau_{nl}(t)$ denotes the time delay from the n -th transmitting antenna to the target and then to the l -th receiving antenna. The complex factor α includes free space attenuation, radar cross-section (RCS), and the reflection phase of a target. Under the narrowband assumption, when electromagnetic waves impinge on K targets from directions $\theta = [\theta_1, \theta_2, \dots, \theta_K]^T$, the echo can be more concisely expressed as

$$s_R^{nl}(t, m) = G_l \cdot \sum_{k=1}^K \alpha_k \cdot e^{j\varphi_k^{nl}} s_T^n(t - \tau_{0,k}(t), m) \quad (4),$$

where G_l is the gain of l -th receiver, α_k corresponds to the k -th target (also including attenuation, RCS, and reflection phase), and φ_k^{nl} is the delay phase difference caused by the antenna element position. This can be expressed as follows under the far-field assumption:

$$\varphi_k^{nl} = \frac{2\pi}{\lambda} (d_n + d_l) \sin \theta_k \quad (5),$$

where $\lambda = c/(f_0 + m \cdot \Delta f)$ is the wavelength, d_n are the baselines of the n -th transmitter and the reference element, and d_l are the baselines of the l -th receiver and the reference element. For simplicity, a linear array is assumed and only azimuth θ is considered, but the approach is also valid for 3D space when using a 2D array in a 4D radar system.

The round-trip delay $\tau_{0,k}(t)$ for a target with constant velocity can be represented as:

$$\tau_{0,k}(t) = \frac{2(r_{0,k} + v_{0,k}(m \cdot t_{\text{sub}} + t))}{c} = \tau_{0,k} + \frac{2v_{0,k}(m \cdot t_{\text{sub}} + t)}{c} \quad (6),$$

where $r_{0,k}$ denotes the range from the reference transmit antenna (the first transmitter of the radar system) to the k -th target at the beginning of the CPI, $v_{0,k}$ is the velocity, and c is the speed of light.

Generally, MMW radars adopt the dechirp mode in the radio frequency module. By mixing the transmitted and received

signals, the de-chirped intermediate frequency (IF) signal, which will be sampled by an analog-to-digital converter (ADC), can be expressed as:

$$s_{\text{IF}}^{nl}(t, m) = G_l \cdot G_n \cdot p_n(m) \sum_{k=1}^K \alpha_k \cdot e^{j\varphi_k^{nl}} \cdot \exp(-j2\pi(f_0 + m \cdot \Delta f)\tau_{0,k}(t)) \cdot \exp(-j2\pi\mu\tau_{0,k}(t)) \cdot \exp(j\pi\mu\tau_{0,k}^2(t)) \quad (7).$$

Unlike the signal models described in most literature, the frequency variation of Δf between adjacent pulses yields a new phase term $\exp(-j2\pi \cdot m \cdot \Delta f \tau_{0,k}(t))$, which plays a crucial role in achieving high range resolution.

Neglecting the last term and substituting Eq. (6) into Eq. (7), we can get:

$$s_{\text{IF}}^{nl}(t, m) = G_l \cdot G_n \cdot p_n(m) \sum_{k=1}^K \alpha_k \cdot e^{j\varphi_k^{nl}} \cdot \exp(-j2\pi(f_0 + m \cdot \Delta f)\frac{2r_{0,k}}{c}) \cdot \exp(-j2\pi(f_0 + m \cdot \Delta f)\frac{2v_{0,k}(m \cdot t_{\text{sub}} + t)}{c}) \cdot \exp(-j2\pi\mu\frac{2r_{0,k}}{c}) \cdot \exp(-j2\pi\mu\frac{2v_{0,k}(m \cdot t_{\text{sub}} + t)}{c}) \quad (8).$$

Since the Doppler frequency variation caused by different carrier frequencies is small, the Doppler frequency $f_{D,k}$ corresponding to the k -th target is:

$$f_{D,k} = f_0 \frac{2v_{0,k}}{c} = \frac{2v_{0,k}}{\lambda_0} \approx (f_0 + m \cdot \Delta f) \frac{2v_{0,k}}{c} \quad (9).$$

Let $f_{R,k}$ denote the range-dependent frequency component corresponding to the k -th target, which is represented by:

$$f_{R,k} = -\mu \frac{2r_{0,k}}{c} \quad (10).$$

The beat frequency signal denoted by $f_{B,k}$ is

$$f_{B,k} = f_{R,k} + f_{D,k} \quad (11).$$

In automotive radars, the Doppler frequency of the target is typically negligible compared with the frequency resolution of the beat signal, i.e., $f_d = f_s/Q$, where f_s is the sampling frequency of the beat signal and Q is the number of fast-time

samples. Thus, we can approximate $f_{B,k} \approx f_{R,k}$ without loss of generality.

Constant phase terms and some less significant terms can be omitted, leading to:

$$\begin{aligned}
 s_{\text{IF}}^{nl}(t, m) &\approx G_l \cdot G_n \cdot p_n(m) \sum_{k=1}^K \alpha_k \cdot e^{j\varphi_k^{nl}} \cdot \\
 &\exp\left(-j2\pi(m \cdot \Delta f) \frac{2r_{0,k}}{c}\right) \cdot \\
 &\exp\left(-j2\pi \left(\underbrace{f_{D,k} + f_{R,k}}_{f_{B,k}}\right) t\right) \cdot \\
 &\exp\left(-j2\pi f_{D,k} \cdot (m \cdot t_{\text{sub}})\right) = \\
 &G_l \cdot G_n \cdot p_n(m) \sum_{k=1}^K \alpha_k \cdot e^{j\varphi_k^{nl}} \cdot \\
 &\exp\left(-j2\pi f_{R,k} t\right) \cdot \\
 &\exp\left(-j2\pi \left(f_{D,k} t_{\text{sub}} + \frac{2r_{0,k}}{c} \Delta f\right) \cdot m\right)
 \end{aligned} \quad (12).$$

Compared with the traditional signal models in Refs. [7] and [28], the proposed model introduces a new term $\exp\left(-j2\pi(m \cdot \Delta f) \frac{2r_{0,k}}{c}\right)$ related to the frequency agility Δf and the range $r_{0,k}$. Due to this additional term, the Doppler frequency is coupled with range across pulses.

The received signal $s_{\text{IF}}^{nl}(t, m)$ is sampled, and then a fast Fourier transform (FFT) is applied to each individual chirp signal. This splits the radar echo signal into different range gates, and the range resolution is determined by the bandwidth of each pulse B . The beat frequency $f_{B,k}$ can be measured for each target, and the range information of potential targets can be acquired.

For total M pulses, the above procedure is repeated. Then, a second FFT inside each single range gate is applied to the resulting signal spectra. However, it is worth noting that the frequency obtained by the second FFT across pulses is not the Doppler frequency f_D , as it is coupled with the target distance. After the range-dimension FFT (column-wise) and the Doppler-dimension FFT (row-wise), the range-Doppler matrix (RDM) is obtained.

3 Proposed Approach

3.1 Analysis of Range-Doppler Spectra

In traditional waveform diversity MIMO radar systems, the transmitted signals from different transmitters are separated by matched filters, and we can obtain NL RDMs directly. However, for traditional DDMA, the signals are orthogonal in the Doppler domain, so only L RDMs can be obtained, and each RDM will be divided into N matrices. In other words, one tar-

get corresponds to N peaks in one RDM.

For example, in a typical case where two TI AWR2243 chips are cascaded, $N=6$ and $L=8$.

False peaks pose a significant issue in DDMA transmission. The method of transmitting a segment of blank spectrum, as shown in Fig. 3b, is a relatively simple and effective approach. Ref. [32] provides a detailed elaboration on this issue. It should be noted that the frequency $f_{p,n}$ ($n = 1, 2, \dots, N$) obtained by the second FFT across pulses is not the real Doppler frequency f_D of the target, as it also includes the influence of range. Therefore, for targets with the same radial velocity but different ranges, the positions will be different.

3.2 Principles for Obtaining High-Resolution Range-Doppler Spectra

The adopted signal processing flow is illustrated in Fig. 4. For simplicity, a single receiving antenna is taken as an example. After each pulse is received, the same signal is applied to the dechirp process for all receivers, since all transmitted signals are identical only with different modulated phases for DDMA. As analyzed above, for each pulse, DDMA can be viewed as a time-shared phased array radar. Therefore, the received signal is essentially the same, with only differences in amplitude due to the varying gain at the target location, which is caused by the beam pattern formed by the modulated phases between pulses. As a result, reference transmitted signals from one of the transmitting antennas can be used for the entire dechirp process.

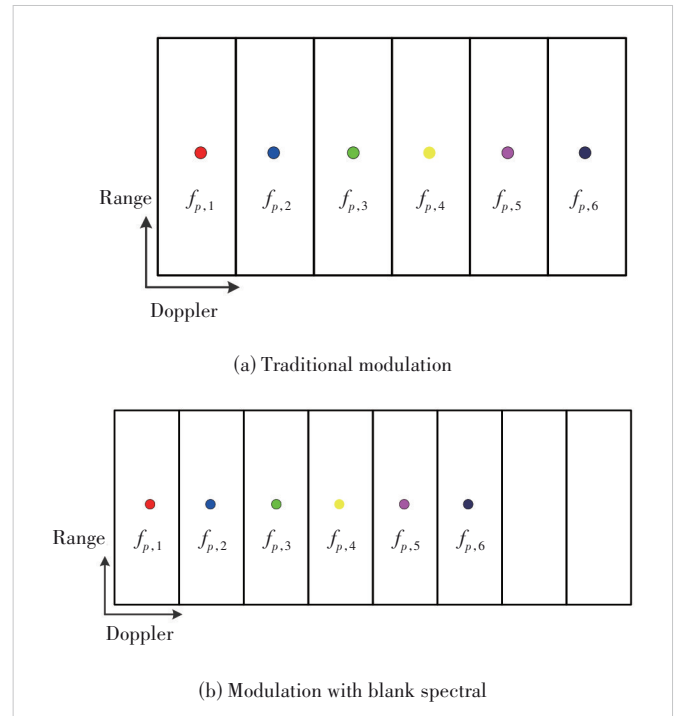


Figure 3. Range-Doppler matrix of a slow-moving target with the traditional DDMA modulation ($N=6$)

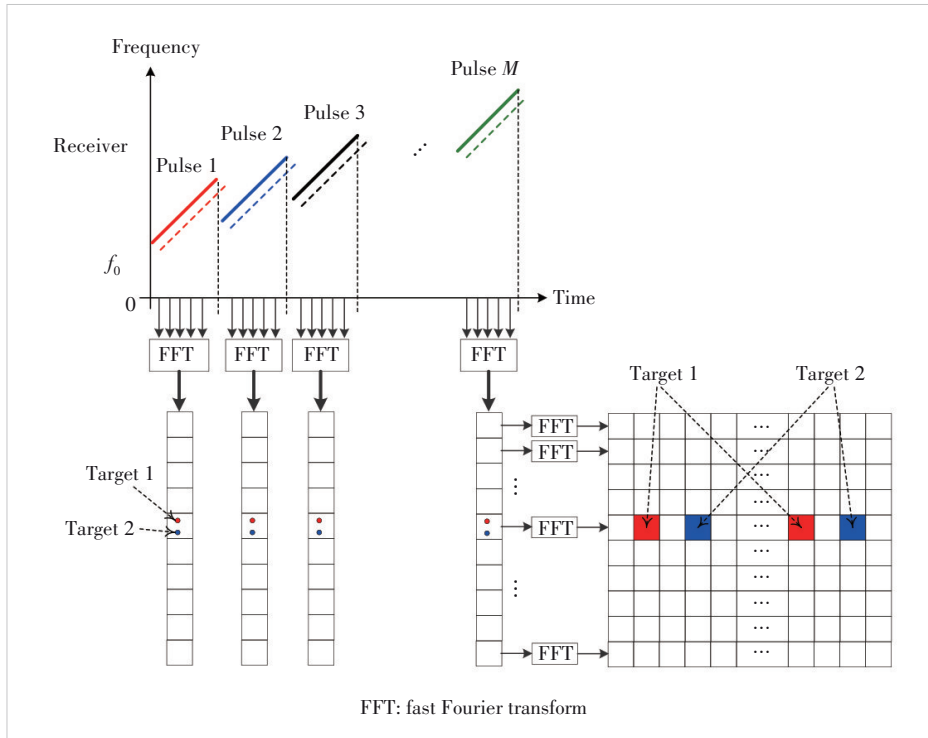


Figure 4. Diagram of signal processing

This paper focuses on improving the range resolution through frequency agility. Two point targets located in the same range cell are configured, as illustrated in Fig. 4. Although two point targets are depicted within a single range cell in the figure, they are indistinguishable in the range-dimension spectrum of each pulse after the dechirping process.

The above derivation also shows that due to the frequency agility across pulses and the distance between the two targets, an additional frequency value can be formed during the FFT processing along the pulse dimension, which is superimposed on the Doppler frequency of the targets. Therefore, if the frequency value resulting from the combined effect of the distance difference and the frequency hopping value is greater than the Doppler resolution, the two point targets will be located within different Doppler cells after performing two-dimensional FFT processing, represented by the red box (target 1) and the blue box (target 2) in Fig. 4. Additionally, due to DDMA

modulation, a target corresponds to multiple peaks, where all the red boxes correspond to target 1.

To illustrate the differences more clearly, the following analysis covers three typical scenarios, including an ideal point target within the range cell ranging from r_1 to $r_1 + \Delta R$, two ideal point targets with slightly different distances, and continuous ground clutter within this range cell. For these three scenarios, we compare the range-Doppler (R-D) spectra obtained using the traditional method of transmitting a chirp pulse train with those of the approach using inter-pulse frequency agility. When there is a difference in the frequency dimension of each transmitted pulse—for instance, when each pulse jumps according to a frequency Δf —the resulting R-D spectrum varies for different radar detection scenarios, as illustrated in Fig. 5.

As shown in the figure, for an ideal stationary point target in the R-D spectrum, the range cell re-

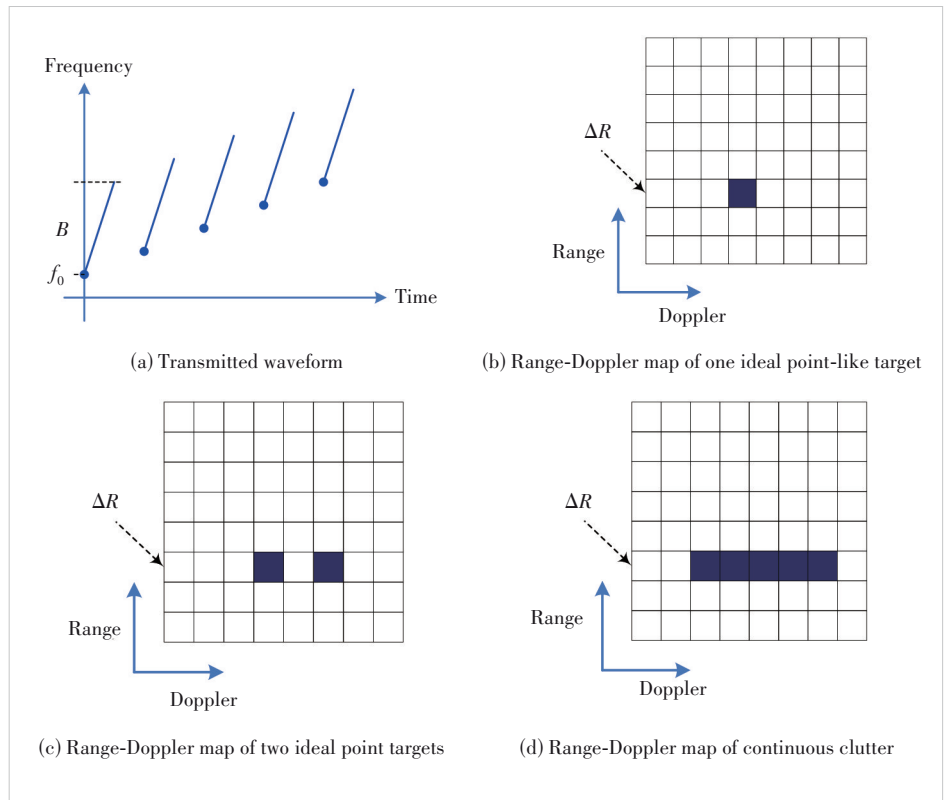


Figure 5. Range-Doppler spectra of typical scenarios for FA FMCW signals

mains consistent with that of a conventional chirp signal, but the Doppler frequency is no longer zero due to DDMA modulation. Similarly, for two ideal point targets, two peaks appear in the R-D spectrum, both exhibiting non-zero Doppler frequencies at different positions in the Doppler dimension. A continuous clutter target corresponds to a continuous velocity interval in Doppler cells.

These effects can be explained by the physical process of waveform propagation. When signals are transmitted at different frequencies, a regular phase change occurs between pulses due to the change in pulse frequency. Performing an FFT on the pulse dimension allows this continuous phase change to manifest as a specific velocity value. Assuming that the precise distance of point target a is r_a , the precise distance of point target b is r_b , and the starting distance of the continuous clutter is r_1 with a distance length of ΔR .

3.3 Proposed Approach for High Resolution Automotive Radar Imaging

This section describes how to obtain a high-resolution R-D spectrum. The process begins with transmitting a sequence of LFM pulses. Then, based on the detected target distance information, the boundaries of the current range bin are determined. Next, the corresponding Doppler channel boundaries are calculated, and a judgment is made on whether Doppler ambiguity exists. After that, a high-resolution range spectrum is obtained, accounting for the presence of Doppler ambiguity. Finally, angle estimation is performed to generate the target contour, and the calculation process can be described as follows:

1) Step 1: Transmit a series of M LFM pulses with a step size of Δf in the frequency hopping sequence. Perform a 2D FFT on the resulting pulse echo data matrix to obtain the R-D spectrum.

2) Step 2: Based on the detected target range bin i , determine the distance interval $[r_{i,\min}, r_{i,\max}]$ for that range segment. The calculation method is as follows:

$$r_{i,\min} = (i - 1) \cdot \Delta R \quad (13),$$

$$r_{i,\max} = i \cdot \Delta R \quad (14),$$

where $\Delta R = c/(2B)$ is the range resolution corresponding to the bandwidth B of a single pulse transmission signal.

3) Step 3: Given the boundaries of the range bin, calculate the corresponding boundaries of the velocity interval and determine the velocity channel number:

$$v_{i,\min} = r_{i,\min} \frac{\Delta f}{f_{\text{start}} \cdot T} \quad (15),$$

$$v_{i,\max} = r_{i,\max} \frac{\Delta f}{f_{\text{start}} \cdot T} \quad (16),$$

where f_{start} is the frequency of the first pulse within the current CPI processed for the velocity dimension FFT, Δf is the frequency step between pulses, and T is the pulse repetition interval.

4) Step 4: Determine the velocity channel number based on the calculated velocity interval boundaries.

Assuming the channel numbers are counted from 0 to $M - 1$, the channel numbers corresponding to the minimum and maximum values of the velocity interval are, respectively:

$$q_{i,\min} = \lfloor v_{i,\min} / \Delta v \rfloor \quad (17),$$

$$q_{i,\max} = \lfloor v_{i,\max} / \Delta v \rfloor \quad (18),$$

where $\lfloor \cdot \rfloor$ denotes the operation of rounding down to the nearest integer.

5) Step 5: Calculate the high-resolution range profile within the velocity interval.

Based on the obtained velocity interval numbers $q_{i,\min}$ and $q_{i,\max}$ corresponding to the current range bin, calculate the output of the high-resolution range profile within that velocity interval:

$$r_{ij} = \frac{(q_{ij} \cdot \Delta v) \cdot f_{\text{start}} \cdot T}{\Delta f} \quad (19),$$

where q_{ij} represents the velocity channel index between $q_{i,\min}$ and $q_{i,\max}$.

6) Step 6: Correct the high-resolution range profile output when resolving velocity ambiguity.

The above formula calculates the distance value r_{ij} , which is valid when there is no velocity ambiguity. If velocity ambiguity occurs, the actual distance value should be corrected to:

$$r_{ij} = \frac{((q_{ij} + (n_{\text{amb}} - 1) \cdot M) \cdot \Delta v) \cdot f_{\text{start}} \cdot T}{\Delta f} \quad (20),$$

where n_{amb} represents the ambiguity order of the velocity channel. When there is no ambiguity, this value is 1. For the current range bin, it can be determined whether the value of the target range exceeds the maximum unambiguous range by checking if:

$$n_{\text{amb}} = \text{floor}(r_{ij} / r_{v,\max}) \quad (21).$$

When the calculated lower and upper boundaries of a range bin differ, it is necessary to split that range bin, i.e.,

$$\text{floor}\left(r_{i,\min}/r_{v,\max}\right) \neq \text{floor}\left(r_{i,\max}/r_{v,\max}\right) \quad (22).$$

When this condition holds, first calculate:

$$n_{\text{amb}} = \text{floor}\left(r_{i,\min}/r_{v,\max}\right) \quad (23).$$

Search for the velocity channels, and when the velocity channel returns from $M - 1$ back to 0, increment the ambiguity order by 1, i.e.,

$$n_{\text{amb}} = n_{\text{amb}} + 1 \quad (24).$$

7) Step 7: Perform DOA estimation on the high-resolution range spectrum.

For each received data matrix, calculate the corresponding high-resolution R-D spectrum matrix. Perform DOA estimation on the first column of this matrix, which corresponds to the zero-velocity channel at different range dimensions. Methods such as DBF or MUSIC can be used for estimation. This allows for an accurate description of clutter distribution in 2D space. Since the presence of different velocities for the same target has a positive effect on target contour imaging, the Doppler spread issue of the target is not specifically addressed in the paper. Instead, it can be processed according to the procedure outlined previously.

4 Simulation and Experimental Results

First, it is verified that high resolution in the range dimension can be achieved through frequency agility between pulses. With a radar frequency f_0 of 77 GHz, a signal bandwidth B of 100 MHz within a single chirp, 1 024 pulses emitted, and a frequency interval Δf of 3 MHz between pulses, the range resolution calculated based on the bandwidth of a single pulse is 1.5 m. Since the actual system verification later in this paper uses a cascade of two TI AWR2243 commercial chips, to ensure consistency with the actual radar system, the simulation here also adopts a 6-transmitter 8-receiver MIMO array, which is equivalent to a uniform linear array containing 48 elements arranged at half-wavelength intervals.

Assume that there are three ideal point targets located at 150.2 m, 150.5 m, and 151 m, respectively. If processed using the traditional non-frequency agile pulse train mode, these three targets would all fall within the same range cell and could not be resolved. However, by adopting a frequency agile pulse emission mode, the range resolution after synthesizing a wideband signal is approximately 0.049 m, allowing for the complete differentiation of the three targets.

In the following simulations, to closely approximate real-world testing scenarios, the simulated objects are modeled as extended targets, such as vehicles 5 m in length and 1.5 m in width. These models contain a large number of randomly distributed scattering points, with random positions, reflection in-

tensities, and phases. The performance of different algorithms is compared. Traditional angle measurement methods using pulse trains with the same carrier frequency employ the MUSIC algorithm. However, for the DDMA-FA-MIMO-based algorithm, only the simplest DBF method is considered for angle measurement. In practical applications, if the computational complexity is acceptable, some super-resolution algorithms can also be integrated into this processing framework. The range-Doppler spectrum for a single vehicle is shown in Fig. 6.

In Fig. 7, the DDMA-FA-MIMO method generates denser and more concentrated radar point clouds than the traditional method, with the colorbar indicating signal reflection intensity. Based on the point cloud output results, the outlines of the vehicles can be accurately determined.

We conducted real-world verification using a radar system consisting of two cascaded TI AWR2243 commercial chips. The radar detection results for real-world road scenarios are

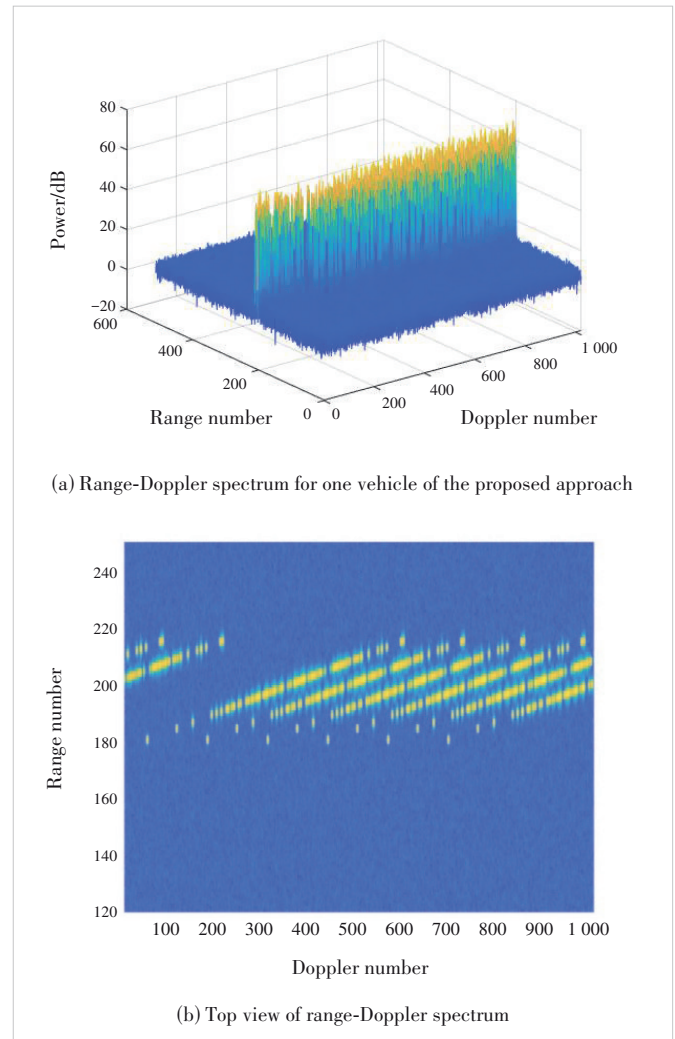


Figure 6. Range-Doppler spectrum and top-view imaging of the proposed MIMO radar approach

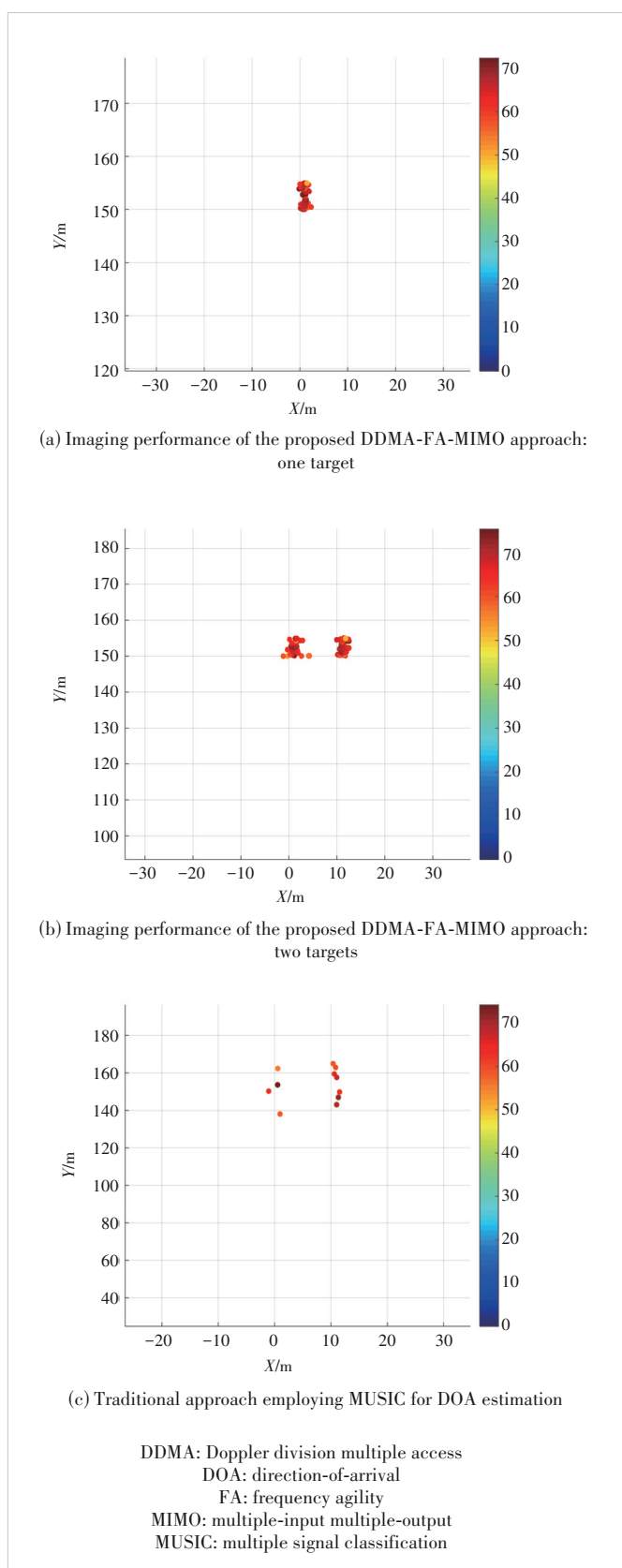


Figure 7. Comparison of target imaging performance between traditional method and DDMA-FA-MIMO approach

presented in Fig. 8, where stationary pedestrians and vehicles are clearly identified, and the imaging of various vehicles and pedestrians at near and far distances is all very clear.

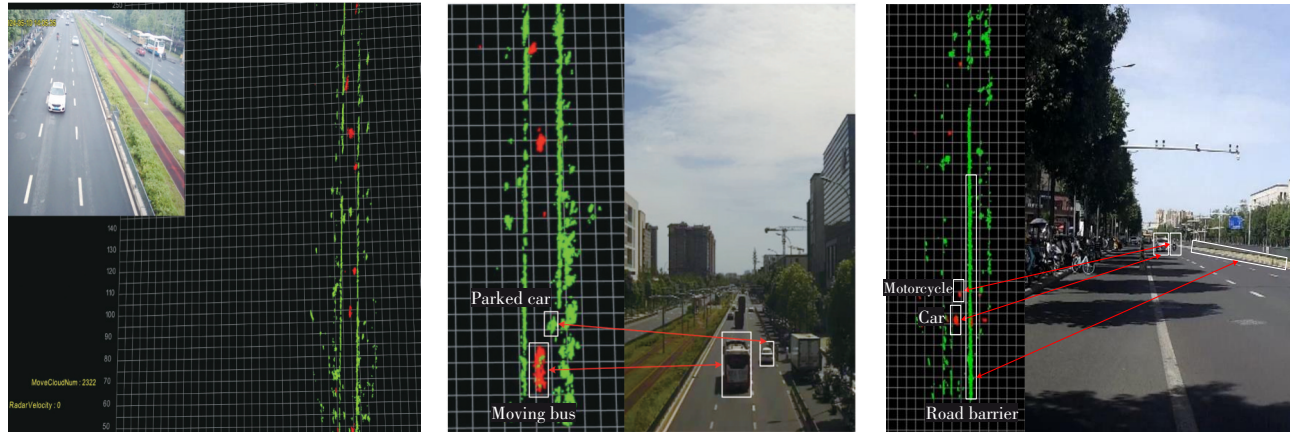
To avoid vehicle obstructions encountered during ground-level installation, the radar was deployed on an overhead bridge. Separate tests were performed for oncoming and receding vehicle targets, as shown in Figs. 8a and 8b. The results reveal highly dense point-cloud outputs with well-defined target outlines. Although road surface reflections are not ideal specular reflections, they are relatively continuous. CFAR processing was applied to detect target contour output^[35–36], which effectively removed sporadic noise points. Therefore, grass or road edges were clearly detected, while there was no detection output for the road surface. Both moving buses and stationary roadside vehicles have distinct contours. Fig. 8c depicts the ground-based radar test scenario. The size of the point cloud envelope is related to the target's height, and vehicle contours are significantly larger than those of motorcycles. These contours enable clear identification of various road targets.

5 Conclusions

The high angular resolution of LiDAR enables it to play a crucial role in ADAS and even autonomous driving. To enhance the competitive advantage of MMW radar over LiDAR, extensive research has been conducted on improving angular resolution. However, it is worth noting that LiDAR achieves its high angular resolution through a very narrow emission beam. If MMW radar could generate a very narrow emission beam while maintaining low sidelobes, it could achieve similar effects as LiDAR does. Nevertheless, achieving such a narrow emission beam comes with significant costs. Currently, MMW radar relies primarily on backend angle super-resolution, compressed sensing, and intelligent algorithms to achieve high angular resolution.

In practical automotive radar applications, targets are typically extended targets rather than isolated ideal point targets. This results in many equivalent scattering points, significant differences in the strengths of these scattering points, and rapid dynamic changes in equivalent scattering points. These factors limit the performance of the aforementioned algorithms in practical applications.

Based on the analysis of practical automotive radar application scenarios, this paper highlights the significant value of range resolution, suggesting that the traditional challenge of angular discrimination can be transformed into a range discrimination problem. Since range resolution is directly related to the bandwidth of the transmitted signal, and transmitting a large signal bandwidth for medium- to long-range MMW radars significantly enhances the sampling rate, which increases costs and is unachievable with current commercial devices. Therefore, this paper proposes a method of pulse-dimension frequency agility to synthesize wideband signals.



(a) Scenario 1: approaching vehicle testing

(b) Scenario 2: departing vehicle testing

(c) Scenario 3: ground testing

Figure 8. Field test results of practical radar system

At the same time, to increase signal transmission efficiency, this paper combines the frequency-agile transmission method with the DDMA orthogonal MIMO transmission method, proposing the DDMA-FA-MIMO transmission framework. It conducts a detailed analysis of scenarios involving ideal single-point and multi-point targets, as well as continuous clutter targets, under this framework. The paper elucidates the physical mechanism of high-resolution imaging and provides the implementation process of the algorithm.

Through theoretical simulations and practical system tests using the dual-cascaded TI AWR2243 commercial chip, the proposed approach has been verified to generate high-density point clouds for targets and realize preliminary imaging of target contours. These capabilities provide a solid foundation for target recognition in ADAS and autonomous driving systems.

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