

# Automatic Tracking of Bandpass Sampling OFDM Signals in Mobile Environment for Enhanced Security Communication

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## **Abstract**

*Countless security and defense systems require an accurate Angle of Arrival (AOA) detection for incoming signals. Angle of Arrival (AOA) estimation is basic procedure required by several monitoring and tracking systems for security applications and other defense purposes. In this paper, a combined OFDM system and bandpass sampling method using Multiple Signal Classification (MUSIC) algorithm for automatic (angle of arrival) AOA tracking is discussed. And we propose a new method that adding (time division multiplexing) TDM with bandpass sampling in the same time to avoid interference due to RF filter characteristics. Also, we consider Doppler effect for the targets' movement in mobile environment and after compensating the Doppler effect with a valid range, the system performances well. Simulation results show that the performances of MUSIC spectrum for AOA due to various conditions and demonstrates the accuracy of AOA estimations.*

**Keywords:** *MUSIC, AOA, Bandpass sampling, OFDM, Doppler effect*

## **1. Introduction**

Countless security and defense systems require an accurate Angle of Arrival (AOA) detection for incoming signals. Smart antenna is one of the possible solutions to increase the channel capacity due to an increase in the number of mobile units and the need for high-speed digital communication in mobile communication. Smart antenna utilizes the beamforming technique to spatially direct the electromagnetic power to an intended mobile unit while spatially null the signal power along other mobile units. The system needs the process of angle of arrival estimation to locate the mobile units before beamforming can be performed. Angle of arrival estimation technology play an important role in enhancing the performance of adaptive arrays for mobile wireless communications [1]. A number of angle of arrival estimation algorithms have been developed. For the most recent ones being MUSCI [2] and ESPRIT [3] algorithms, who both utilizing subspace-based on exploiting the eigen structure of the input covariance matrix and thus requires a higher computation effort. Although ESPRIT needs less computation, the MUSIC algorithm is found to be more stable and accurate [4]. In this paper, we use the MUSIC algorithm combine the OFDM bandpass sampling signal model to perform the antennas sensing to allow accurate azimuth. The accuracy of the estimation in azimuth increases proportional to the number of antenna elements utilized.

Bandpass sampling can be used for direct down conversion without analog mixers. In practice, the required sampling rate for ADC can be too high to be achieved if the Nyquist sampling theorem is to be satisfied [5]. So we use bandpass sampling which is a technique that samples high data rate signals with smaller sampling rate than Nyquist sampling rate to

relax the demand for ADCs. After down-sampling about over two band signals using bandpass sampling, the signals are digitized and then two band signals can be received [7-10].

In this paper, we propose a bandpass sampling technique with time division multiplexing (TDM). In previous system, although over two signals can be down-sampling without interference between signals, it is possible to generate interference due to RF filter characteristics. RF filter cannot cut adjacent band signals so the remaining adjacent band signals (undesired signals) can affect desired signals. So we propose bandpass sampling with TDM that can avoid previous problems to separate over two signals timely.

## 2. System Model

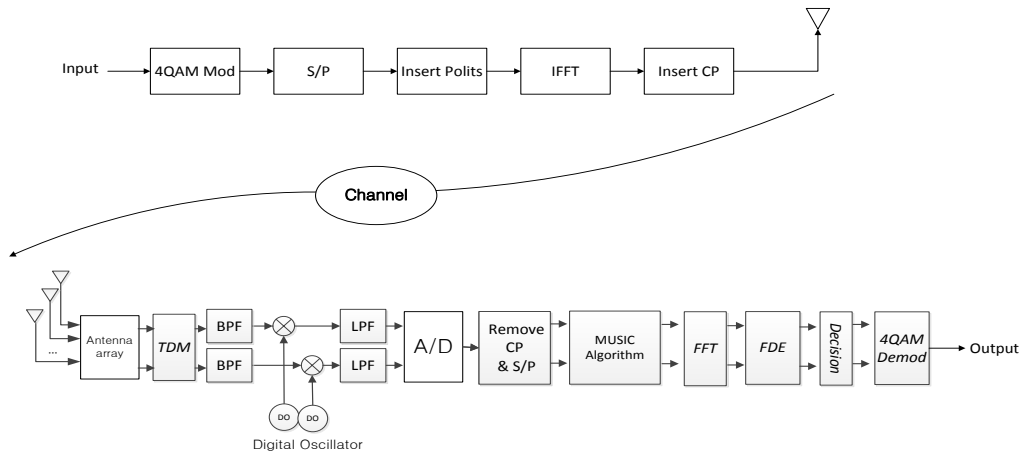


Figure 1. System Model

### 2.1. OFDM Signal and Multi-Antenna Receiver

In this paper we consider two signals that have different center frequency. Transmitted signals are based on OFDM. Eq. (1) is the signals in time domain. Assume that there are two received bands.  $X_{k,m}^A$  and  $X_{k,m}^B$  are transmitted signals respectively. As Eq. (1), the signals is represented after IFFT in time domain.

$$x(t) = \begin{cases} \frac{1}{\sqrt{N}} \sum_{k=0}^{N-1} X_{k,m}^A e^{j(\frac{2\pi k}{N} + f_A)t}, & x_A(t) \\ \frac{1}{\sqrt{N}} \sum_{k=0}^{N-1} X_{k,m}^B e^{j(\frac{2\pi k}{N} + f_B)t}, & x_B(t) \end{cases} \quad (1)$$

The receiver is equipped with an uniform linear array(ULA) of  $M$  sensors located along a straight line. The system can be viewed as the multiplication of each received ray by a steering vector considering the direction of arrival of each multipath. Assume that there are  $P$  ( $P < M$ ) uncorrelated narrowband signals  $x_p(t)$  received by ULA from different direction  $\theta_p$ , corrupted by AWGN, where  $p=1,2,\dots, P$ . The observation is given as

$$X(t) = \sum_{p=1}^P a(\theta_p) * x_p(t) + n(t) \quad (2)$$

where  $a(\theta)$  is the array steering vector given by

$$a(\theta) = [1 \ e^{-j2\pi d \sin \theta / \lambda} \ \dots \ e^{-j2\pi d \sin \theta (M-1) / \lambda}]^T \quad (3)$$

where  $d$  is the inter element spacing,  $\lambda$  is the signal wavelength. When we take snapshot at time  $k=1,2,\dots,K$ , we can get

$$X(k) = \sum_{p=1}^P a(\theta_p) * x_p(k) + n(k) \quad (4)$$

where noise  $n(k)$  is assumed to be both temporally and spatially white, and uncorrelated with signal  $s_p(k)$ .

## 2.2. Angle of Arrival Detection using MUSIC

MUSIC stands for MUltiple Signal Classification. The covariance matrix,  $R$ , is the collected data for each of the array receivers in the time domain. The correlation matrix is given as [6]

$$R = E[XX^H] = AR_sA^H + \sigma^2I \quad (5)$$

where  $R_s$  is the  $P \times P$  signal correlation matrix.  $\sigma^2$  is the white noise power. The noise subspace  $E_N$  used in MUSIC can be obtained from eigenvalue decomposition of  $R$ , and the spatial spectrum of MUSIC is given by

$$P(\theta) = \frac{1}{a(\theta)^H E_N E_N^H a(\theta)} \quad (6)$$

## 3. Doppler Effect and Compensation

The orthogonally among subcarriers is often destroyed by the CFOs due to oscillator mismatches. So Doppler effect was generated and degrades performance. Doppler effects cause shifting in frequency domain and phase rotation in time domain.

Signal  $x(t)$  is like (7) due to Doppler effect.

$$y_n = \sum_{k=0}^{N-1} H_k \cdot X_k \cdot e^{i2\pi \frac{k+\varepsilon}{N}} + z_n \quad (7)$$

Signal  $x(t)$  is like (8) due to Doppler effect in time domain.

Channel  $H$  is represented as product of  $X$ . Doppler effect is represented phase rotation in frequency domain.  $k, n, \varepsilon$  are sub-carrier, symbol, normalized Doppler frequency respectively in (7).

$$\begin{aligned} Y_p &= \sum_{m=0}^{N-1} \sum_{k=0}^{N-1} H_{k,m} \cdot X_{k,m} \cdot e^{i2\pi \frac{(k+\varepsilon)}{N}} \cdot e^{-i2\pi \frac{m}{N}} + Z_p \\ &= H_p \cdot X_p \cdot e^{i2\pi \varepsilon p} + \sum_{\substack{m=0 \\ m \neq k}}^{N-1} \sum_{k=0}^{N-1} H_{k,m} \cdot X_{k,m} \cdot e^{i2\pi \frac{(k-m)}{N}} \cdot e^{i2\pi \frac{\varepsilon}{N}} + Z_p \end{aligned} \quad (8)$$

In (8), first stage is phase rotation and second stage is ICI. Where  $p$  is symbol in frequency domain and  $k, m$  are sub-carrier before IFFT in transmitter and sample before FFT in receiver. Phase rotation of Doppler is different per symbol and

ICI is generated when one sub-carrier affects other sub-carriers.

Normalized offset value is divided by carrier spacing. We consider direction of receiver is heading toward to transmitter.

$$\varepsilon = \frac{f_d}{\text{carrier spacing}}, f_d = \frac{v \cdot f_c}{c} \quad (9)$$

In this system, we compensate those problems with synchronization signal and block type pilot and assume that the receiver speed is constant.  $f_d$ ,  $c$ ,  $v$  are Doppler frequency, the velocity of light, the speed of receiver respectively.

The problem of Doppler effect is compensated with block type pilot. The phase rotation is estimated by doing interpolation between pilots, because the receiver speed is not dramatically changed.

$$Y_p = H_p \cdot X_p e^{i2\pi\varepsilon p} + Z_p \quad (10)$$

Phase rotation is estimated using received pilot signals.

$$P(i) = \sum_{n=1}^N \text{mean} \left\{ \sum_{n=1}^{64} \text{Block\_Pilot}(i+n-1) \right\} \quad (11)$$

$$\frac{\text{angle}\{P(i)\} - \text{angle}\{P(i+1)\}}{\text{pilot\_interval}} \cdot ([1: \text{pilot\_interval} - 1])$$

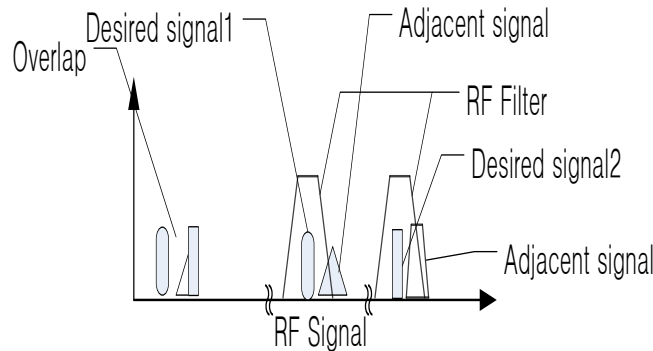
$i=1, 2, \dots$  (12)

$P$  is average of block type pilot. Eq. (12) represents linear interpolation using  $P$ , so the symbols that have not pilots is estimated. But if there is too long pilot interval or Doppler frequency, it is difficult to compensate the effect.

## 4. Proposed Bandpass Sampling

### 4.1. Existing Method

Existing multi-band system with bandpass sampling finds sampling frequency that doesn't overlap signals between multi-band signals according to (7). But to select multi-band signals respectively, RF filter is used. Although RF filter has good Q value, the RF filter can't remove all adjacent signals. So the remaining adjacent signal is able to be overlap when multi-band signals are converted at low frequency band.



**Figure 2. System Model Problem when Signals are Sub-Sampled from RF Band**

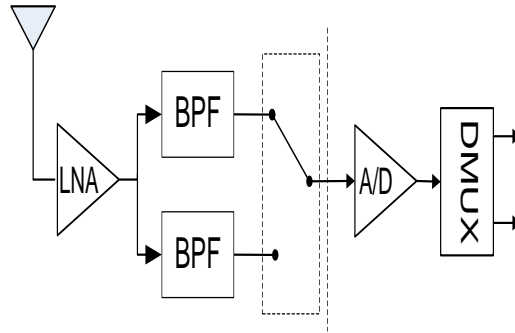
Bandpass sampling about multi-band of over 2 bands meet condition like (13) [7]. To convert the two signals in low frequency band without interference between signals,  $F_{IF,A}$  and  $F_{IF,B}$  have to meet (13).

$$\begin{aligned}
 &0 < F_{IF,A} - BW_A / 2, \quad F_S > F_{IF,A} - BW_A / 2 \\
 &0 < F_{IF,B} - BW_B / 2, \quad F_S > F_{IF,B} - BW_B / 2 \\
 &\text{if } F_{IF,B} > F_{IF,A} \\
 &\quad F_{IF,B} - BW_B / 2 > F_{IF,A} + BW_A / 2 \\
 &\text{if } F_{IF,A} > F_{IF,B} \\
 &\quad F_{IF,A} - BW_A / 2 > F_{IF,B} + BW_B / 2
 \end{aligned} \tag{13}$$

First, the signals that are converted in low frequency band are large than 0 and smaller than  $F_S/2$  respectively. Second, the low frequency part of  $F_{IF,A}$  is larger than the high frequency part of  $F_{IF,B}$  ( $F_{IF,B} < F_{IF,A}$ ) or, the low frequency part of  $F_{IF,B}$  ( $F_{IF,A} > F_{IF,B}$ ) is larger than the high frequency part of  $F_{IF,A}$  ( $F_{IF,B} > F_{IF,A}$ ).

#### 4.2. Proposed Structure

We propose a method that adds TDM method in bandpass sampling method.



**Figure 3. Multi-band Receiver Structure with Joint sub-sampling and TDM Method**

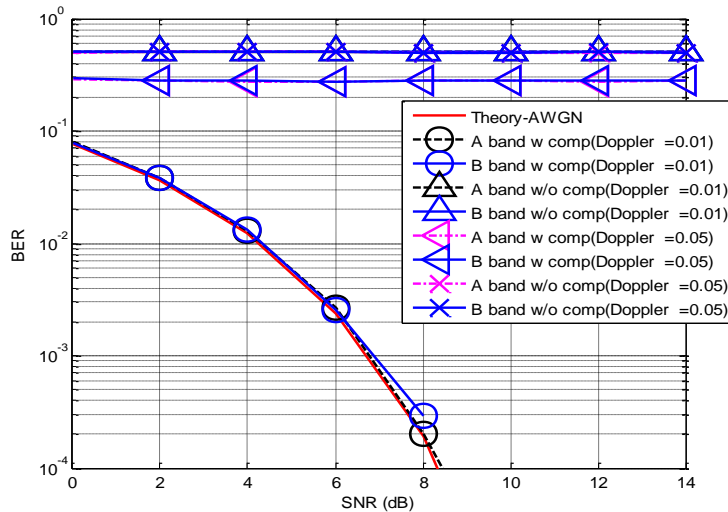
$$\begin{aligned}
 &0 < F_{IF,A} - BW_A / 2, \quad F_S > F_{IF,A} - BW_A / 2 \\
 &0 < F_{IF,B} - BW_B / 2, \quad F_S > F_{IF,B} - BW_B / 2
 \end{aligned} \tag{14}$$

The proposed structure is given as Figure 3. Multi-band signals are received with antennas and the signals pass through LNA. Afterward, the multi-band signals are divided into two signals by filter. And each signal is sampled as two times faster than the existing bandpass sampling frequency. At Sampling & Holder and ADC, TDM and bandpass sampling are performed at the same time. The signals that are received with TDM have no interference between receiving singles because the signals is divided in time. Therefore, the converted signals just satisfy (14) instead of (13). So it is possible to give an low sampling frequency.

## 5. Simulation and Discussion

**Table 1. Simulation Parameters**

OFDM system	
The number of Subcarriers	64
Bandwidth	20MHz
Symbol Period	4 $\mu s$
Subcarrier Spacing	312.5KHz
CP Length	0.8 $\mu s$
Modulation	4-QAM
Channel	AWGN
Target A azimuth	10°
Target B azimuth	50°



**Figure 4. BER performance with Doppler Effect**

Simulation environment is showed in Table 1. Figure 4 indicates BER performance when Doppler effect occurs. We can see the performance according to Doppler scale. The two bands have no difference according to Doppler scale. The two bands have no difference due to TDM. In the case of Doppler effect  $\varepsilon = 0.01$ , for that both A band and B band without compensating, we can't communicate because of the phase rotation. And after compensating phase rotation, there is small performance degradation comparing to the theory curve because of existing ICI. And when the Doppler Effect is give  $\varepsilon = 0.05$ , we can't communicate as we use block type pilot and do linear interpolation which is difficult to estimate fast phase rotation.

Figure 5 shows spatial spectrum of MUSIC with 4 receiver antenna arrays due to different SNRs. We can see that both target A and target B are tracked with accurate angle whose are 10° and 50°. And we can also see that more higher the SNR is, the shaper the spectrum pointing the angle performances.

Figure 6 show that MUSIC spectrum performance due to different array antenna numbers under the same SNR enviornment. And we can see that the more array antenna number is, the more accurately the system can separate the signals which performances as a sharper spectrum.

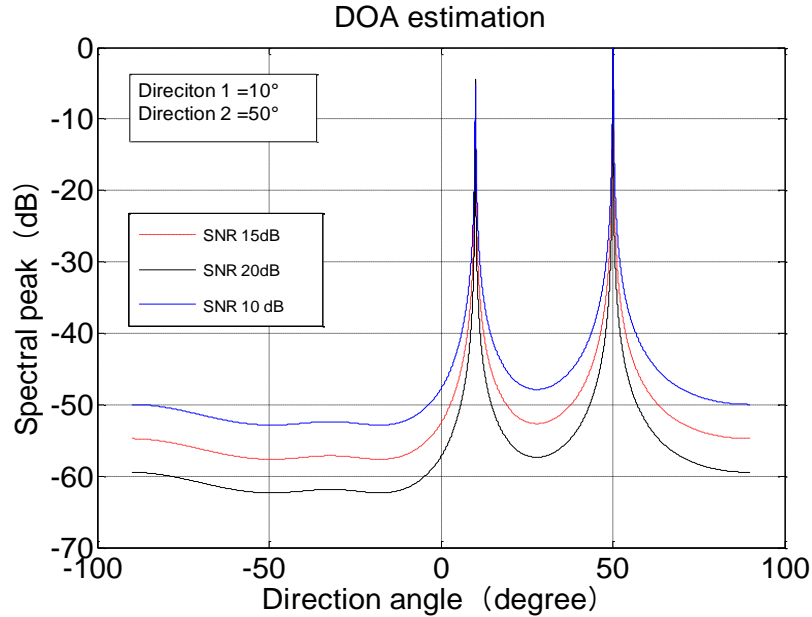


Figure 5. AOA Estimation Due to SNR Change

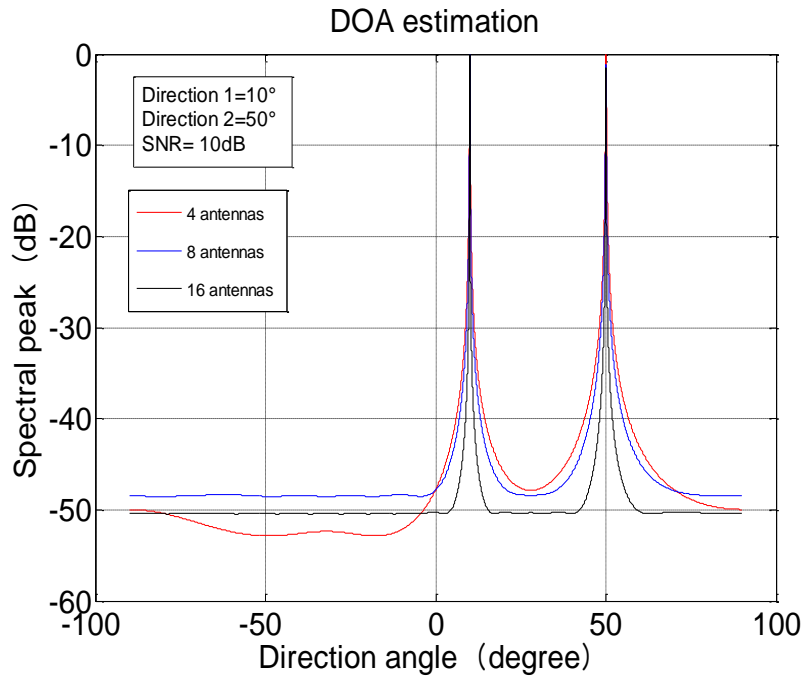


Figure 6. AOA Estimation due to Array Antenna Numbers with SNR=10dB AOA Estimation due to SNR Change

## 6. Conclusion

In this paper, we discussed and performance a automatic AOA tracking method using MUSIC algorithm by bandpass sampling method. And we also proposed an adding TDM with bandpass sampling method which can avoid interference. By considering the Doppler effect and compensating the effect, system using proposed method performances well. And simulation shows using MUSIC algorithm to estimate the AOA under different conditions.

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