

Design of Multi-band Receiver with Pre-FFT Beamformer for Wireless Communications

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Abstract—Because of the limitations of the sampling speed and power consumption in the very high frequency several bands of the wireless communication systems, the current receiver cannot directly sample the incoming signals at the Nyquist rate just after a RF stage process. Instead, using the sub-sampling technique can convert multi-band signals from RF band to IF band without oscillators. But Due to the filter performance in RF bands, signals converted into low bands cause interference and degrade performances. In this paper, we like to propose a sub-sampling method and the pre-FFT beamformer for the multi-band receiver, and a new approach for the synchronization and compensation of the STO (sampling time offset) and CFO (carrier frequency offset) are proposed to avoid interference by TDM method. The purpose of designing the pre-FFT beamformer is to coherently combine the desired signals and suppress the undesired signals so that it is able to provide medium to high gains in a mobile environment. Simulation results show that after compensating the Doppler effects and timing offset the approach works effectively and we can get optimal performances in the multi-band receiver.

Keywords—Sub-Sampling, TDM, Multi-band receiver, Offset compensation, timing Synchronization ,Beamforming, LMS

I. INTRODUCTION

Recently, orthogonal frequency division multiple access (OFDMA) attract much attention as a multiple access technique because of its high spectral efficiency and tolerance to multipath channels. In uplink OFDMA systems, available subcarriers are divided into groups and assigned to multiple users for simultaneous transmissions. [1] The ideal universal receiver sampling the incoming signals just after a RF stage is very seducing but not realistic at this time due to limitations in sampling capabilities for high frequency systems. A multiple antenna array can be used at the receiver, not only for spectral efficiency or gain enhancement, but also for interference suppression .Software-defined radio (SDR) [2] principles have attracted more and more attention. But currently it is difficult to digitize RF signals using ADC. So digital IF method has been studied. That is to digitize signals at IF band after down-converting from RF band to IF band.

If a signal is digitized in IF band after down-converting with oscillator, it is possible to digitize a multi-mode signal. But it can be difficult that one receiver is to get non-

contiguous signals like one contiguous signal. To meet requirement of IMT-Advanced, carrier aggregation (CA) is used in LTE-Advanced [3]. CA is used to get up to 100MHz bandwidth. That is one method that combines carrier component (CC). So in this paper to get some CC, sub-sampling technique is used.

Sub-Sampling is a technique that samples high data rate signals with smaller sampling rate than Nyquist sampling rate. There have been studied about Sub-Sampling [4]. After down-sampling about over 2 band signals using Sub-Sampling, the signals are digitized, and then over 2 band signals can be received [5].

Based on pilot-based transmission ,the information of pilot is used to estimate optimum weights for the spatial filtering operated by smart antenna. In order to work well the OFDM symbol should be correctly synchronized and Doppler shift should be correctly estimated and compensated. Combined with beamforming the system can achieve optimal performance. Beamforming process is performed using a pre-FFT method[6] based on VSS-LMS(Variable Step Size)[7,8].

In this paper, we propose a Sub-Sampling technique with time division multiplexing (TDM). In previous system, although over 2 signals can be down-sampling without interference between signals, it is possible to generate interference due to RF filter characteristic. RF filter cannot cut adjacent band signals so the remaining adjacent band signals (undesired signals) can affect desired signals. So we propose sub-sampling with TDM that can avoid previous problems to separate over 2 signals in time.

II. SYSTEM MODEL

In this paper we consider 2 signals that have different center frequency. Transmitted signals are based on OFDM. Eq. (1) is the signals in time domain. In this paper, we 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.

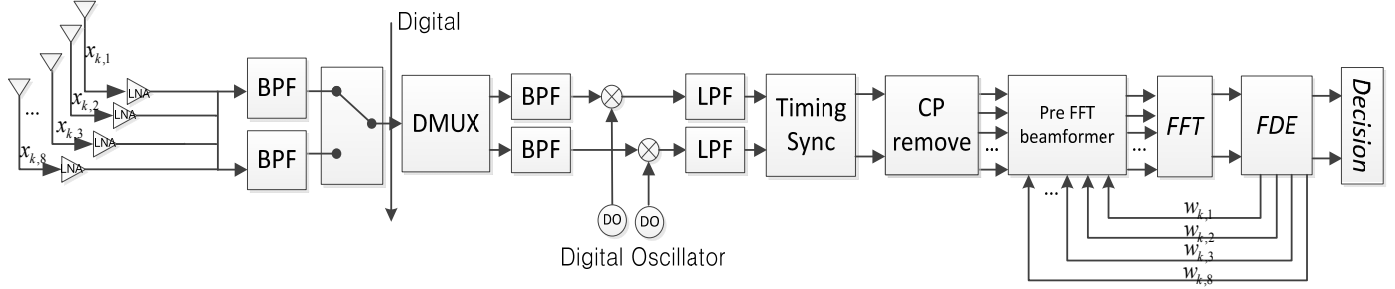


Figure 1. Multi-band receiver structure with Pre-FFT beamformer using sub-sampling and TDM.

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

Each band is represented as $x_A(t)$ and $x_B(t)$ in time domain and has center frequencies as f_A and f_B . The channel is considered affected by AWGN, multipath and Doppler shift. In our case we model the effect of multipath using a method of a linear system which can be written as [12]:

$$h(t) = \beta_0(t)\delta(t) + \sum_{m=1}^M \beta_i(t) \cdot \delta(t - \tau_i) \quad (2)$$

where M is the number of multipaths and τ_i is the delay of arrival of the m -th path. $\beta_0(t)$ and $\beta_i(t)$ are complex values that take into account attenuation and Doppler shift.

The receiver is equipped with an uniform linear array (ULA) of L 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 (DoA) of each multipath. If a ray arrives with an angle of θ , then the signal carried by the ray will be multiplied by $e^{j\pi h \sin \theta}$ which $h = 0, \dots, L-1$. The signal received by the ULA can be written by a $L \times N_T$ matrix

$$X = a(\theta_d) \cdot s^T + \sum_{m=1}^M a(\theta_m) \cdot r_m^T + M \quad (3)$$

where s and r_m are the column vectors of the direct and reflected rays attenuated by the Doppler shift. θ_d and θ_m are the respective angles of arrival and $a(\cdot)$ is the column steering vector

$$a(\theta) = \begin{bmatrix} 1 & e^{j\pi \sin \theta} & \dots & e^{j\pi(L-1)\sin \theta} \end{bmatrix}^T. \quad (4)$$

The AWGN effect is represented by the matrix M . And since it can be considered omnidirectional, it affects every antenna and M is generated independently.

Two signals are received through one antenna. After amplification of received signals in LNA, each signal passes through RF filter. After BPF, each signal is sampled in time. The sampling frequency is smaller than Nyquist rate but larger than signal bandwidth of twice like figure 1. At this point, TDM and Sub-Sampling is processed at the same time. After ADC, each band sampling frequency is 110MHz. And using digital oscillator the center frequency of converted signal is moved to DC. The Digital oscillator is processed in digital part, so there are no problems that are generated by analog oscillator. And then the signals are synchronized using synchronization signal.

III. THE BEAMFORMING ALGORITHM

A. LMS algorithm

The LMS algorithm is a method to filter a sequence of noisy data. It is based on the availability of a reference signal used to design the denoising filter due to the estimation of an error signal. The filter design is based on the minimization of the mean squared value of the error signal. The obtained expression is a quadratic function of w_k . The basic idea of the LMS algorithm is to evaluate the vector w_k with an iterative approach. Filter coefficients are updated to approach the minimum squared error (MSE) value with null gradient. The updating is performed by

$$w_{k+1} = w_k - \mu \cdot g(w_k) \quad (5)$$

where $g(\cdot)$ is the stochastic gradient of the estimated MSE calculated at the k -th due to w_k :

$$g(w_k) = \nabla_{w_k} (M \hat{S} E_k). \quad (6)$$

The convergence characteristics of the algorithm depend on the positive scalar μ . Here we have chosen to use

VSSLMS(variable step size LMS) to adjust the values of μ according to the working condition[9].

B. Pre FFT Beamforming

We have opted for the pre-FFT method because of its much lower complex for needing only one FFT per OFDM symbol is required while the post-FFT method needs L FFT[6]. Furthermore, the output of pre-FFT beamforming is a time domain signal which can be processed by a synchronization algorithm. This cannot be possible with the post-FFT scheme for its output is in the frequency domain. We use the symbol “ $\sim$ ” to indicate a FFT operation. And we define that

- $\bar{s}$, the vector of dimensions $J \times 1$, with J = the number of pilot tones containing the reference signal.
- w_k , $L \times 1$ vector containing the complex gains.
- $\tilde{X}_k$, the $L \times N$ matrix whose every row is the FFT of corresponding row of X_k where is the k-frame received data matrix without CP.
- D , the $J \times N$ matrix indicating the pilots position.

The instantaneous estimation of MSE can be written as:

$$M \hat{S} E_k = \left\| \bar{s}^T - w_k^H \tilde{X}_k D^T \right\|^2. \quad (7)$$

Using the generalized derivative rules, the gradient can be rewritten as:

$$g(w_k) = 2 \tilde{X}_k D^T (D \tilde{X}_k^H w_k - \bar{s}^*) = 2 \tilde{X}_k \tilde{e}_k \quad (8)$$

where

$$\tilde{e}_k = D^T (D \tilde{X}_k^H w_k - \bar{s}^*).$$

Using the generalized Parseval equality to rewrite (8) as:

$$\begin{aligned} e_k &= \text{IFFT} \left\{ D^T (D \tilde{X}_k^H w_k - \bar{s}^*) \right\} \\ &= \text{IFFT} \left\{ D^T (\bar{y}_k - \bar{s})^* \right\}. \end{aligned} \quad (9)$$

IV. TIMING AND FREQUENCY SYNCHRONIZATION

Signal $x(t)$ is like (10) due to Doppler effect.

$$y_n = \sum_{k=0}^{N-1} H_k \cdot X_k \cdot e^{i2\pi \frac{k+\varepsilon}{N}} + z_n \quad (10)$$

Signal $x(t)$ is like (11) 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 , ε are sub-carrier, symbol, normalized Doppler frequency respectively in (10).

$$\begin{aligned} Y_p &= \sum_{m=0}^{N-1} \sum_{k=0}^{N-1} H_{k,m} \cdot X_{k,m} \cdot e^{i2\pi \frac{(k+\varepsilon)}{N}} \cdot e^{-i2\pi \frac{m}{N}} + Z_p \\ &= H_p \cdot X_p \cdot e^{i2\pi p} + \sum_{\substack{m=0 \\ m \neq k}}^{N-1} \sum_{k=0}^{N-1} H_{k,m} \cdot X_{k,m} \cdot e^{i2\pi \frac{(k-m)}{N}} \cdot e^{i2\pi \frac{\varepsilon}{N}} + Z_p \end{aligned} \quad (11)$$

In (11), 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

In this system, we compensate those problems with synchronization signal and block type pilot and assume that the receiver speed is constant. 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.

$$Y_p = H_p \cdot X_p \cdot e^{i2\pi p} + Z_p. \quad (12)$$

Phase rotation is estimated using received pilot signals.

$$P(i) = \sum_{i=1}^N \text{mean} \left\{ \sum_{n=1}^{64} \text{Block_Pilot}(i+n-1) \right\} \quad (13)$$

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

$$i=1, 2, \dots \quad (14)$$

P is average of block type pilot. Eq. (14) represents linear interpolation using P , so the symbols that have not pilots is estimated.

A signal that has N samples is a symbol. If the symbol is delayed, the symbol affects next a symbol. So ISI occurs. But ISI is removed by cyclic prefix (CP). Eq. (15) represents phase error and ISI.

$$\begin{aligned} y_n &= h_n \otimes x_{n+\delta} + z_n \\ Y_p &= \sum_{m=0}^{N-1} \sum_{k=0}^{N-1} H_{k,m} \cdot X_{k,m} \cdot e^{i2\pi \frac{k-m}{N}} \cdot e^{i2\pi \frac{\delta_m}{N}} + Z_p \\ &= H_p \cdot X_p \sum_{m=0}^{N-1} e^{i2\pi \frac{\delta_m}{N}} + \sum_{\substack{m=0 \\ m \neq k}}^{N-1} \sum_{k=0}^{N-1} H_{k,m} \cdot X_{k,m} \cdot e^{i2\pi \frac{k-m}{N}} \cdot e^{i2\pi \frac{(\delta_n)}{N}} + Z_p \end{aligned} \quad (15)$$

Received signal y_n is time domain signal. Transmitted signals x_n has delay like $n+\delta$ and affected by channel h .

Eq. (15) indicates one band case. Because TDM method separate received two signals in time. So each band has different effect like (15) respectively. But TDM method can't sample two bands at the same time. So it is possible to generate delay at second band. Sample timing offset is like (16).

$$Y_p = H_p \cdot X_p \sum_{m=0}^{N-1} e^{-i2\pi \frac{\delta_m}{N}} \quad (16)$$

Phase error of sample timing offset has different phase degree per sub-carrier because of orthogonality of OFDM. So using synchronization signal, the offset value is estimated [13].

We use constant amplitude zero autocorrelation (CAZAC) sequence as synchronization signals for its good self-correlation to get fine timing and dropper offset. The CAZAC is used to design the training sequence (TS). As the CAZAC sequence does not change its property after IFFT/FFT processing, so the TS is still a CAZAC sequence after OFDM modulation.

Assume S is the TS of the OFDM systems, according to [14], it can be represented as

$$S(k) = e^{j\frac{M\pi k^2}{N}}, k = 0, 1, 2 \dots N-1 \quad (17)$$

M is relative-prime to N . Assume $H(k)=1, M=1$ when there has a frequency offset in the system, the received signal of the TS is

$$r_s(k) = e^{-j\frac{\pi \epsilon^2 \pi}{N}} e^{j\frac{\pi(k+\epsilon)^2}{N}} + z_n, \quad k = 0, 1, 2 \dots N-1. \quad (18)$$

The property of auto-correlation of the TS is used in this part get the timing [13]

$$M(d) = \sum_{k=1}^N r_s(d+k) s^*(k) \quad (19)$$

$$= \left[e^{-j\pi \epsilon_i^2 / N} e^{j\pi(d+k+\epsilon_i)^2 / N} e^{-j\pi \epsilon_f^2 / N} + z(k) \right]^* e^{-j\pi k^2 / N}$$

$$\hat{\delta} = \arg \max (M(d)) \quad (20)$$

where $M(\delta)$ is the cross-correlation of the received TS

and the know TS in the receiver, $\hat{d}$ is the coarse timing start point. Assume ϵ_i and ϵ_f is the integer and fractional CFO. According to the property of the CAZAC sequence,

$M(d)$ will get the max value at $\hat{\delta} = d_{cor} - \epsilon_i$, and d_{cor} is the correct start point of the TS.

V. PROPOSED SUB-SAMPLING METHOD

A. Existing Structure

Existing multi-band system with sub-sampling finds sampling frequency that doesn't overlap signals between multi-band signals according to (10). 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.

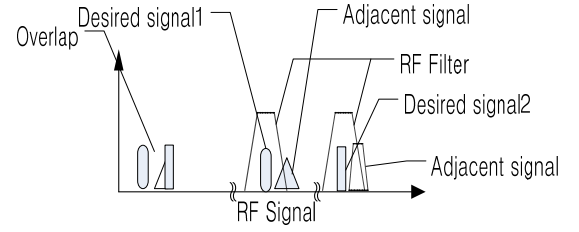


Figure 2. Problem from RF band by sub-sampling.

Sub-Sampling about multi-band of over 2 bands meet condition like (21) [10]. To convert the two signals in low frequency band without interference between signals, $F_{IF,A}$ and $F_{IF,B}$ have to meet (21).

$$0 < F_{IF,A} - BW_A / 2, \quad F_S > F_{IF,A} - BW_A / 2$$

$$0 < F_{IF,B} - BW_B / 2, \quad F_S > F_{IF,B} - BW_B / 2$$

$$\text{if } F_{IF,B} > F_{IF,A}$$

$$F_{IF,B} - BW_B / 2 > F_{IF,A} + BW_A / 2$$

$$\text{if } F_{IF,A} > F_{IF,B}$$

$$F_{IF,A} - BW_A / 2 > F_{IF,B} + BW_B / 2 \quad (21)$$

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 $F_{IF,A}$ ($F_{IF,B} > F_{IF,A}$).

B. Proposed Structure

We propose a method that adds TDM method in Sub-Sampling method.

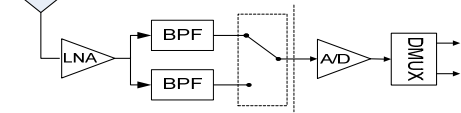


Figure 3. A multi-band receiver structure that joint sub-sampling and TDM method.

The proposed structure is like Fig 3.

$$0 < F_{IF,A} - BW_A / 2, \quad F_S > F_{IF,A} - BW_A / 2$$

$$0 < F_{IF,B} - BW_B / 2, \quad F_S > F_{IF,B} - BW_B / 2 \quad (22)$$

Multi-band signals are received with one antenna and the signals pass through LNA. Afterward, the multi-band signals are divided into two signals by filter. And each signal is sampled as 2 times faster than the existing sub-sampling frequency. At sampling & holder and ADC, TDM and sub-sampling are performed at the same time. The signals that are received with TDM has no interference between receiving singles because the signals is divided in time. Therefore, the converted signals just satisfy (21) instead of (21). So it is possible to decrease sampling

frequency. In the (22), there are no equations about interference between converted signals.

VI.SIMULATION AND DISCUSSION

Table 1. Simulation Parameters

OFDM system	
The number of Subcarriers	64
Bandwidth	20MHz
Symbol Period	4 μ s
The number of Sensors	8
CP Length	0.8 μ s
Pilot Type	Block Type Pilot
Modulation	QAM
Channel	AWGN

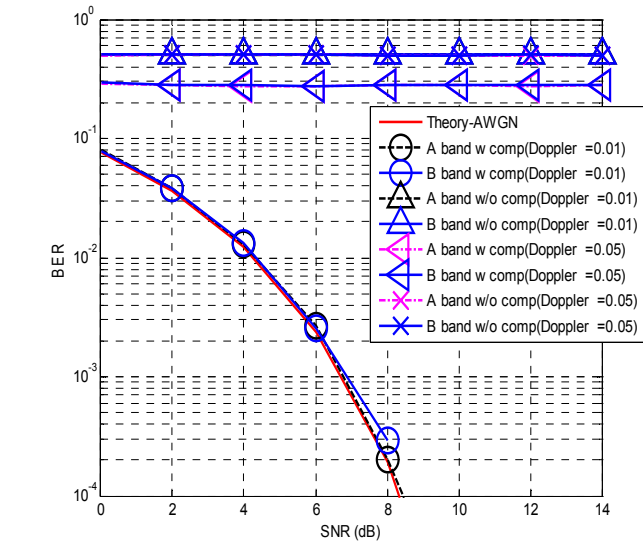


Figure 4. BER performance with Doppler.

Fig. 4 indicates BER performance when Doppler 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 case of A,B band w/o comp for Doppler=0.01 , we can't communicate because of phase rotation. And after compensating phase rotation, there is small performance degradation because of ICI. And when the Doppler=0.05, we can't communicate because we use block type pilot and do linear interpolation. It is difficult to estimate fast phase

rotation.

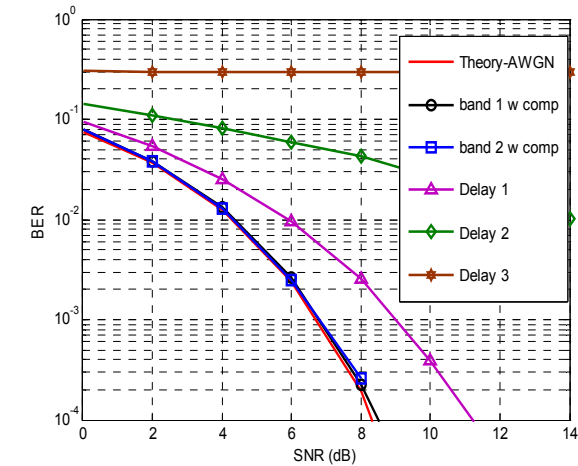


Figure 5. BER performance with timing offset.

Fig. 5 indicates result about timing offset. Here, Delay1 is 3ns, Delay 2 : 0.125us, Delay 3 : 0.5us. We can see two bands can communicate due to Delay 1 and Delay 2 after compensation. And in case of Delay 3 we can't communicate with comp for it is difficult to estimate fast phase rotation.

VII. CONCLUSIONS

In this paper, we propose a downlink OFDMA receiver with multi-band and multi-mode signals using sub-sampling method with TDM combined with beamforming technology. Sub-sampling is method that can convert high frequency band signals to low frequency band signals. It also is able to receive over 2 bands signals with one sample & holder and ADC. The sampling frequency need not very high due to (16). It also has an advantage that each signal is divided which means undesired signals that occurs due to RF filter characteristic don't affect the others signals. Simulations illustrate the BER performance by compensating the frequency offset and timing offset of the system

ACKNOWLEDGEMENT

This research was supported by Basic Science Research Program through the National Research Foundation of Korea(NRF) funded by the Ministry of Education, Science and Technology(No. 2010-0007567)

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