Performance Evaluation of π/4 DQPSK Wireless Communication System with Direct-conversion

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***Abstract*— Recently, there has been a shift from analog to digital wireless communication systems among those used for wireless communication systems. Promotion of the digitization of wireless communication systems has drawn attention to direct-conversion (DC) receivers as means of reducing the cost of digital wireless communication equipment. In this paper, we propose an algorithm for estimating and compensating for Carrier Frequency Offset (CFO) and I/Q imbalance (IQI) caused by DC architecture. This algorithm which uses two preambles improves the algorithm proposed in [9] which is unstable at CFO values near 0 Hz. This paper evaluates the performance of a baseband simulator for an ARIB STD-T61 compliant direct-conversion π/4-DQPSK wireless communication system in a Rayleigh fading.**

***Keywords*— π/4 DQPSK, Direct-conversion, CFO, IQI, ARIB STD-T61**

I. Introduction

Digital wireless communication systems have high tolerance for noise as they are able to tolerate noise and completely reconstruct transmitting signals even in low-SNR environments so long as such noise does not exceed the threshold. Thus, they are superior to analog wireless transmission systems with respect to communication quality. In addition, the transition from analog to digital wireless communication systems is expected to resolve the problems of conventional analog wireless communication systems, such as shortages of channels and insufficient transmission speeds. For this reason, recently the wireless communication systems used for TV and radio broadcasting have been transitioned from analog to digital wireless communication systems.

With the promotion of the digitization of wireless communication system, demand has grown for small and inexpensive digital wireless communication equipment. Under these circumstances, the direct-conversion system has been attracting attention. The super heterodyne system, which is widely used in existing wireless equipment, down-converts received signals in the carrier frequency band to an intermediate frequency (IF) once before converting to the baseband [1]. The direct-conversion system down-converts received signals from the carrier frequency band directly to the baseband.

As the direct-conversion system down-converts received signals without converting to an intermediate frequency, it requires fewer filters and oscillators in its circuits than the super heterodyne system. Thus, it is superior to the super heterodyne system in that smaller systems can be built and power consumption is lower [2]-[5]. On the other hand, unlike the super heterodyne system, the direct-conversion system must process high-frequency signals in the carrier frequency band. For this reason, the direct-conversion system must have high-accuracy analog circuits. If analog circuits of accuracy identical to that of those in the super heterodyne system are used, high-frequency signals cannot be processed and I/Q imbalance (IQI) occurs.

The direct-conversion system uses a quadrature demodulator in the receiver to split received signals into in-phase (I) and quadrature-phase (Q) components. IQI resulting from quadrature errors between the I and Q components distorts the constellation of demodulated signals and degrades bit error rates (BER). In addition, carrier frequency offset (CFO) due to oscillator frequency errors between the transmitter and the receiver also degrades BER.

Recently, algorithms for estimating the CFO and IQI in a receiver have been proposed [6]-[8]. The algorithm in [6] uses special calibration signals to compensate for IQI. However, this method requires complex circuits. [7] proposes a method for compensating for IQI in the presence of CFO. This method estimates spectra by line search in order to estimate CFO and uses a liner least squares (LLS) algorithm to estimate IQI. However, estimation by the frequency domain method involves a complex system and the use of line search limits accuracy. [8] estimates CFO and IQI by the time domain method and is documented to require fewer calculations than the other algorithms discussed above. This method also improves estimation accuracy.

We proposed an algorithm based on [8] in [9], which estimates and compensates for CFO and IQI in narrow band wireless communication systems. However, [9] fails to completely evaluate the algorithm as it does not consider the synchronization or error correction of systems and assumes additive white gaussian noise (AWGN) channels.

This report evaluates performance of a baseband simulator for an ARIB STD-T61 [10] compliant direct-conversion π/4 DQPSK wireless communication system built by using the algorithm for estimating and compensating for CFO and IQI proposed in [9].

ARIB STD-T61 is a standard established by the association of radio industries and businesses in cooperation with wireless equipment manufacturers, telecommunications carriers, and users. ARIB STD-T61 specifies standard specifications and other basic technical requirements for wireless facilities related to the various systems that use radio waves.

II. SYSTEM MODEL

This section outlines the system noted above. This system is compliant with ARIB STD-T61, a standard for frequency division multiple access (FDMA) narrow band digital mobile communication systems. We will describe the standard later. The system uses a direct-conversion quadrature demodulator in the receiver.

First, we will describe the specifications and frame formats of ARIB STD-T61. Next, we will describe the configurations of the transmitter and the receiver. In the description of the receiver configuration, we will note the differences between the direct-conversion system and the super heterodyne system, which is widely used by quadrature demodulators in existing wireless equipment. We also discuss the CFO and IQI that occur in quadrature demodulators when the direct-conversion system is adopted.

An algorithm for estimating and compensating for the CFO and IQI that occur in quadrature demodulators that adopt the direct-conversion system has been proposed in [9]. This algorithm is described in the next section.

A. ARIB STD-T61[10]

ARIB STD-T61 is a standard for FDMA narrow band digital mobile communication systems established by the association of radio industries and businesses in cooperation with wireless equipment manufacturers, telecommunications carriers, and users. ARIB STD-T61 specifies standard specifications and other basic technical requirements for wireless facilities related to the various systems that use radio waves. The specifications of this system are shown in Table 1.

TABLE 1. ARIB STD-T61 SYSTEM SPECIFICATIONS

Access method	FDMA
Modulation method	π/4 DQPSK
Carrier frequency band	400 [MHz]
Interval between adjacent channels	6.25 [kHz]
Transmission rate	9.6 [kbps]
Roll-off rate	0.2
Coding rate	4/5
Tolerable CFO range	-1000 to 1000 [Hz]

B. ARIB STD-T61 Frame Formats

The frame formats used by this system are shown in Fig. 1. The system uses 384-bit, 40-ms long frames. There are two types of 40-ms long frames. Figure 1 (a) shows the frame format for service channels and Figure 1 (b) shows the frame format for multipurpose channels. Service channels are two-way channels consisting of a radio information channel (RICH) and a traffic channel (TCH1) employed by users for communication. Multipurpose channels consist of an unmodulated carrier or an arbitrary modulated bit sequence and are employed in 1) through 5) below:

1) High-speed adjustment of received frequencies (AFC : Automatic Frequency Control)
2) Establishment of symbol synchronization for received signals
3) Reduction of delay distortion in received signals
4) Alleviation of nonlinear distortion characteristics in transmitting signals
5) Supplementing insufficient transmission volume with information bit sequences
6) Compensation for CFO and IQI in received signals

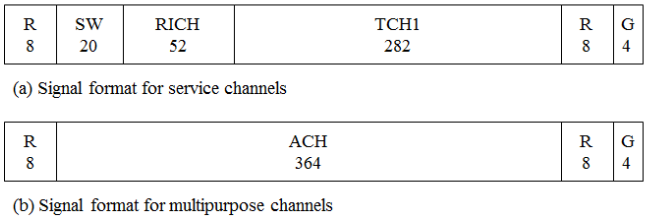

Figure 1. Frame Formats

R, SW, RICH, TCH1, and ACH in Figure 1 are described below :

- Transient response (R): When frames are transmitted in burst mode, this softens the rising and falling edges of the transmission waves in order to prevent the spread of spectrum into adjacent channels. Bit pattern and shape are not specified
- Guard time (G): When signals are transmitted from different mobile stations, the received adjacent frames overlap each other. A guard time in which no transmission occurs is set to guard against the overlapping of signals.
- Synchronous word (SW1): This bit sequence is used for frame synchronization. It is placed in the center of a frame and may be used for training for waveform equalization. In this system, $1E56F_{16}$, specified in ARIB STD-T61 ver. 1.2 [10], is used to synchronize both frames and symbols.
- Radio information channel (RICH): This is used to identify radio channels or when transferring between transfer operation mode, communication mode, and etc.
- Traffic channel 1 (TCH1): This is a two-way point-to-point or point-to-multipoint channel for transferring user information.
- Assist channel (ACH): This is a channel consisting of an unmodulated carrier and an arbitrary bit sequence. An unmodulated carrier and an arbitrary modulated bit sequence can be combined to make an assist channel.

This system uses 364-bit ACH as preambles for estimating CFO and IQI. Figure 2 shows the preambles for estimating

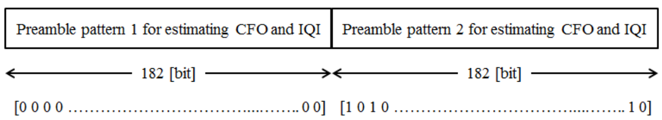

Figure 2. Frame Formats

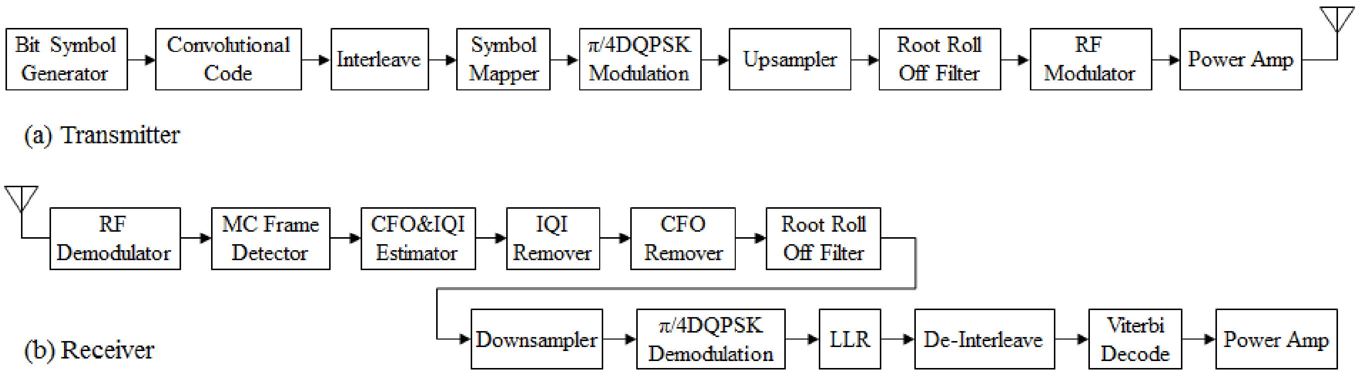

Figure 3. Transmitter and Receiver Configurations

CFO and IQI in detail. Two different preamble patterns, (1) and (2), are used to estimate CFO and IQI. If only a single preamble pattern is used to estimate CFO, the estimation accuracy may be reduced markedly for some CFO values. Thus, this system uses two different preamble patterns according to the CFO value in order to prevent a reduction in estimation accuracy. Section 3 describes the detection of the multipurpose channel frames described in this section. Section 4 describes an algorithm for estimating and compensating for CFO and IQI in [9].

C. Transmitter Configuration

Figure 3 (a) shows the transmitter configuration. First, the transmitter performs convolution coding of transmitting signals. Next, it interleaves the convolution-coded signals. It then performs π/4 DQPSK modulation of the interleaved signals and transmits the signals up-converted to a carrier frequency.

Figure 3 (b) shows the receiver configuration. In existing wireless equipment, the super heterodyne system, as shown in Figure. 4, is used by the RF demodulator, which is the front end of the receiver. On the other hand, this system adopts the direct-conversion system. The circuit configuration is shown in Figure 5. In Figure 5, the parameter α is the IQI gain value, ϕ is the IQI phase imbalance, and Δf is the CFO value.

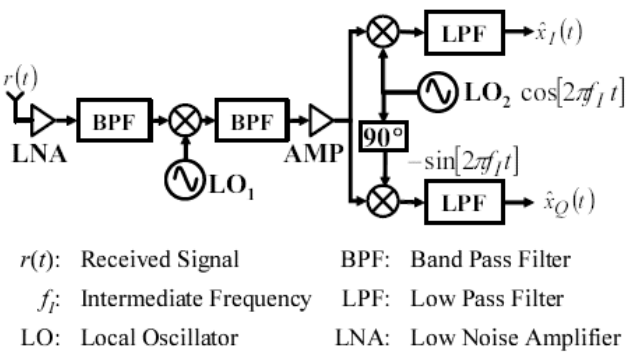

Figure 4. Super Heterodyne Receiver

As previously described, the super heterodyne system downconverts signals in the carrier frequency band to an intermediate frequency once, and then further down-converts them to the baseband, while in contrast the direct-conversion system down-converts signals in the carrier frequency band directly to the baseband. For this reason, the direct-conversion system requires fewer filters and oscillators in its circuits, resulting in smaller circuits and reduced costs. By comparing the circuits in Figures 4 and 5, one can verify that the circuits for the direct-conversion system are smaller.

However, in the direct-conversion system, which ownconverts signals directly to the baseband, high-frequency signals in the carrier frequency band must be processed. This causes degradation of the analog circuits, which in turn causes CFO and IQI. For this reason, circuits to compensate for CFO and IQI must be included.

The MC frame detector is a circuit that detects the multipurpose channels shown in Figure. 1 (b). The CFO & IQI estimator is a circuit that estimates IQI and CFO from signals distorted by IQI and CFO. The estimated values are then used to compensate for IQI and CFO in the IQI and CFO removers. The π/4 DQPSK demodulation performs π/4 DQPSK demodulation of the IQI- and CFO-compensated signals. Finally, the LLRs for the π/4 DQPSK-demodulated signals are calculated and Viterbi decoding is performed.

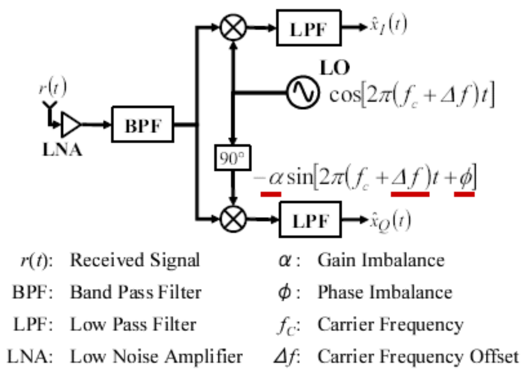

Figure 5. Direct-conversion Receiver

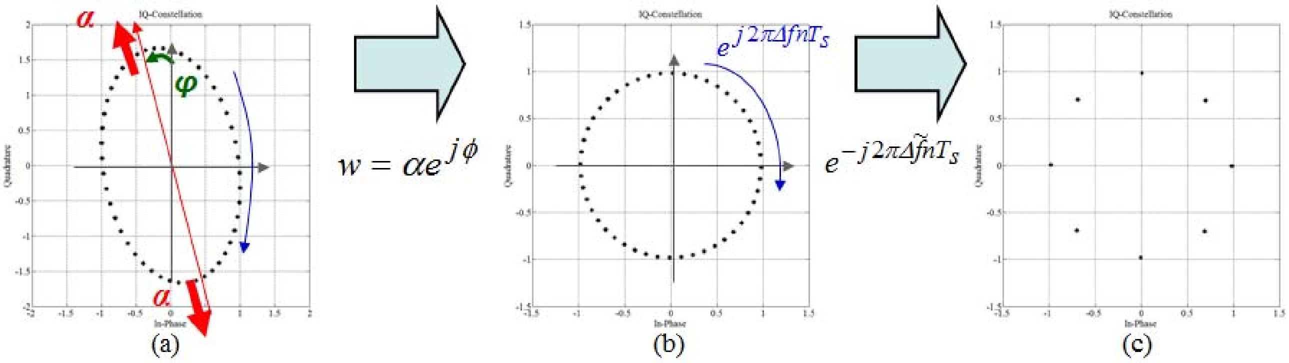

Figure 6. Process of Compensating for CFO and IQI

III. MULTIPURPOSE CHANNEL FRAME DETECTION

Next we will describe multipurpose channel frame detection. The MC frame detector shown in Figure 3 detects the multipurpose channel frame shown in Figure 1 (b). In this system, preambles for compensating for CFO and IQI are present within the multipurpose channel's ACH.

However, according to ARIB STD-T61, the receiver enters areas where data is always transmitted in asynchronous mode. In such an area, as the starting points of frames cannot be found, the receiver must detect multipurpose channel frames from received signals alone. In addition, as CFO and IQI cannot be compensated for until the multipurpose channel frames are detected, this system must detect multipurpose channel frames from received signals affected by CFO and IQI.

Thus, this system uses the preambles for estimating CFO and IQI within the multipurpose channel's ACH as the preambles for detecting multipurpose channel frames. It follows that the ACH shown in Figure 2 serves as the preambles for estimating CFO and IQI and as the preambles for detecting multipurpose channel frames.

The RF demodulator in Figure 3 (b) detects multipurpose channel frames from received signals affected by CFO and IQI by using autocorrelation. The autocorrelation value of a received signal affected by CFO and IQI, $z(k)$, can be expressed by equation (2), where N is the autocorrelation length and $\hat{x}(t)$ is the received signal affected by CFO and IQI. This system calculates 182-bit autocorrelation for received signals to detect multipurpose channel frames. As the received signal $\hat{x}(t)$ is a complex signal upsampled eightfold, the autocorrelation length N is the number of samples calculated by dividing 182 by the 2 symbols and multiplying the resulting quotient by 8.

$$N = (182/2)\times 8 \qquad (1)$$

$$z(k) = \frac{1}{N}\left|\sum_{n=0}^{N-1}\hat{x}(k-n)\hat{x}(k-N-n)\right| \qquad (2)$$

Thus, the MC frame detector performs autocorrelation for received signals affected by CFO and IQI by using equation (2), passing the autocorrelation values' peaks as the points where multipurpose channel frames were detected to the CFO & IQI estimator in Figure 3 (b).

IV. ALGORITHM FOR ESTIMATING AND COMPENSATING FOR CFO AND IQI

This section describes the algorithm for estimating and compensating for CFO and IQI proposed in [9]. First, we express the received baseband signal with CFO and IQI distortion as an equation. Then, we will describe the algorithm for compensating for CFO and IQI.

A. CFO and IQI

When a signal in the carrier frequency band is down-converted with the direct-conversion system to a signal in the baseband by the RF demodulator of the receiver shown in Figure 5, CFO and IQI occur. The signal affected by CFO and IQI is expressed by equation (3).

$$\hat{x}(t) = \left\{e^{-j2\pi\Delta ft}x(t)\right\}\frac{1}{2}\left(1+\alpha e^{-j\phi}\right) + \left\{e^{j2\pi\Delta ft}x^*(t)\right\}\frac{1}{2}\left(1-\alpha e^{j\phi}\right) \qquad (3)$$

$x(t)$ is the baseband signal created by the transmitter. It can also be expressed using $x_I(t)$ and $x_Q(t)$, which are the I and Q components, respectively, of $x(t)$, as follows: $x(t) = x_I(t) + jx_Q(t)$. α is the gain imbalance and ϕ is the phase imbalance of the I and Q components, while Δf is the CFO value.

B. CFO and IQI Imbalance Compensation Scheme

When the algorithm proposed in [9] is used to compensate for the influences of CFO and IQI on a received baseband signal, the CFO- and IQI-compensated signal $\hat{S}(n)$ can be expressed by equation (4).

$$\hat{S}(n) = \{\mathrm{Re}[\hat{x}(n)]w + j\,\mathrm{Im}[\hat{x}(n)]\}e^{j2\pi\Delta\tilde{f}nT_s} \qquad (4)$$

[9] proposes an algorithm that uses two periodic preambles to estimate CFO and IQI parameters. As previously described, this system also adopts two periodic preambles as shown in Figure 2. Equation (5) is an algorithm for estimating CFO values. N is the number of symbols contained in the periodic preamble signal. η indicates whether the CFO is positive or negative. η is determined by the imaginary component of the autocorrelation of the received signal.

$$\Delta\tilde{f} = \frac{\tan^{-1}\left[\eta(\mathrm{a}^T\mathrm{b})^{-1}\sqrt{(\mathrm{a}^T\mathrm{a})^2-(\mathrm{a}^T\mathrm{b})^2}\right]}{2\pi T_S N} \qquad (5)$$

a and **b** in equation (5) are vectors expressed as follows:

$$\mathbf{a} = \begin{bmatrix} 2\hat{x}_I(n+N) \\ 2\hat{x}_Q(n+N) \end{bmatrix} \tag{6}$$

$$\mathbf{b} = \begin{bmatrix} \hat{x}_I(n)+\hat{x}_I(n+2N) \\ \hat{x}_Q(n)+\hat{x}_Q(n+2N) \end{bmatrix} \tag{7}$$

Next, an algorithm for estimating the compensated IQI value, w, is shown in equation (8). The compensated IQI value, w, represents $\alpha e^{j\phi}$

$$w = (\mathbf{c}^H \mathbf{c})^{-1} \mathbf{c}^H \mathbf{d} \tag{8}$$

c and **d** in equation (8) are vectors expressed as follows:

$$\mathbf{c} = \left[\hat{x}_I(n) e^{j2\pi\Delta\tilde{f}T_S N} - \hat{x}_I(n+N)\right] \tag{9}$$

$$\mathbf{d} = \left[j\hat{x}_Q(n+N) - j\hat{x}_Q(n) e^{j2\pi\Delta\tilde{f}T_S N}\right] \tag{10}$$

These compensation algorithms can be used to compensate for CFO and IQI. The process of using the algorithm proposed in [9] for estimating and compensating for CFO and IQI to compensate for received baseband signals with CFO and IQI distortion is shown in Figure. 6.

In Figure 6, (a) shows the constellation of received baseband signals with CFO and IQI distortion, while (b) shows the constellation of the same signals after compensation for the influence of IQI. (c) shows the constellation of the same signals after compensation for the influence of CFO. Regarding the influence of IQI, note that while IQI indicates quadrature imbalance of the I and Q components of received signals, this algorithm indicates to what extent Q components deviate from I components.

V. Performance Evaluation

A. *Simulation Specifications*

The simulation specifications set for performance evaluation are shown in Table 2.

TABLE 2. Simulation Specifications

Modulation method	π/4 DQPSK
Carrier frequency band	400 [MHz]
Transmission rate	9.6 [kbps]
Coding rate	4/5
Channel	Rayleigh Fading
Mobile unit speed	60 [km/h]
Maximum Doppler frequency	22.22[Hz]
CFO	-500 [Hz]
Gain imbalance α	0.5
Phase imbalance ϕ	10 [deg]
Transmitting data length	384 [bit]
Number of transmitted bit	3,840,000 [bit]

B. *Multipurpose Channel Frame Synchronous Characteristics*

Multipurpose channel frame detection rate characteristics at mobile unit velocities of 20 km/h, 60 km/h, and 100 km/h are shown in Figure 7. According to Figure 7, the MC frame error rates at an SNR of 20 dB and mobile unit velocities of 20 km/h, 60 km/h, and 100 km/h were approximately 2 × 10^{-2}, 5 × 10^{-3}, and 2 × 10^{-3}, respectively. When the autocorrelation adopted by this system was used to detect multipurpose channel frames, detection accuracy improved as the mobile unit velocity increased. This is because as mobile unit velocity decreases, Doppler frequency also decreases and consequently the Doppler period becomes longer. When the Doppler period is longer than the multipurpose channel frames, the frames are significantly affected by the Doppler effect. If the Doppler effect is significant, multipurpose channel frames cannot be detected. However, when the Doppler period is short, multipurpose channel frames, even if significantly affected by the Doppler effect, may be synchronized in the part not affected. This is why the accuracy of multipurpose channel frame detection improved as mobile unit velocities increased in the range of velocities from 20 km/h to 100 km/h.

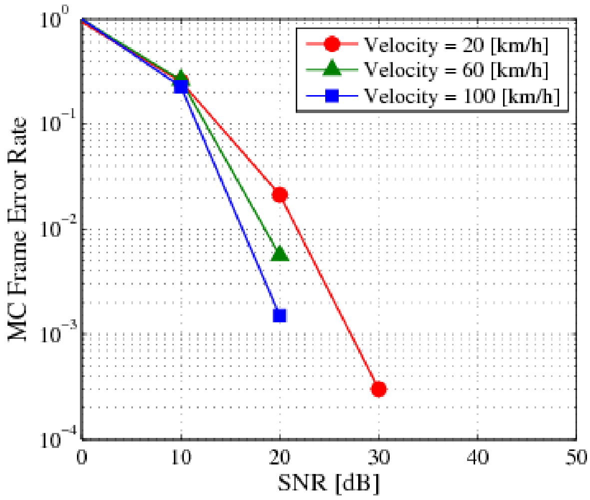

Figure 7. MC Frame Error Rate

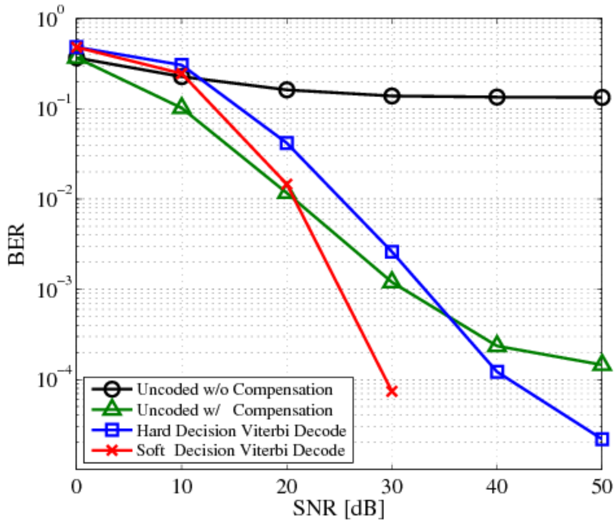

Figure 8. BER Characteristics Under the Rayleigh Fading Environment (Mobile Unit Velocity: 60 km/h, CFO: -500 Hz)

C. BER Characteristics

Figure 8 shows the BER characteristics under the Rayleigh fading environment with a mobile unit velocity of 60 km/h and a CFO value of -500 Hz. According to Figure 8, signals without compensation for CFO and IQI (Uncoded w/o Compensation) cannot be communicated because the BER converges at around 10^{-1} due to the CFO and IQI. On the other hand, the BER for signals with compensation for CFO and IQI (Uncoded w/Compensation) can be improved to approximately 10^{-4} at an SNR of 50 dB. Figure 8 shows that soft decision Viterbi decode improves SNR approximately 5 dB when performed without error correction (Uncoded w/ Compensation) at a BER of 10^{-3}.

VI. CONCLUSION

This report evaluated performance of a baseband simulator for an ARIB STD-T61 compliant direct-conversion π/4 DQPSK wireless communication system built by using the algorithm for estimating and compensating for CFO and IQI proposed in [9].

The MC frame error rates at an SNR of 20 dB and mobile unit velocities of 20 km/h, 60 km/h, and 100 km/h were approximately 2×10^{-2}, 5×10^{-3}, and 2×10^{-3}, respectively. When the autocorrelation adopted by this system was used to detect multipurpose channel frames, detection accuracy improved as mobile unit velocity increased.

We found that the algorithm proposed in [9] is also effective for estimating and compensating for CFO and IQI in this system. According to Figure 8, it was also found that soft decision Viterbi decode improves SNR approximately 5 dB when performed without error correction (Uncoded w/Compensation) at a BER of 10^{-3}. Figure 9 is the π/4 DQPSK wireless communication prototype system which is developed by Kyushu TEN co.ltd, based on the proposed our DC compensation techniques.

In the future, we will conduct field tests with this prototype system.

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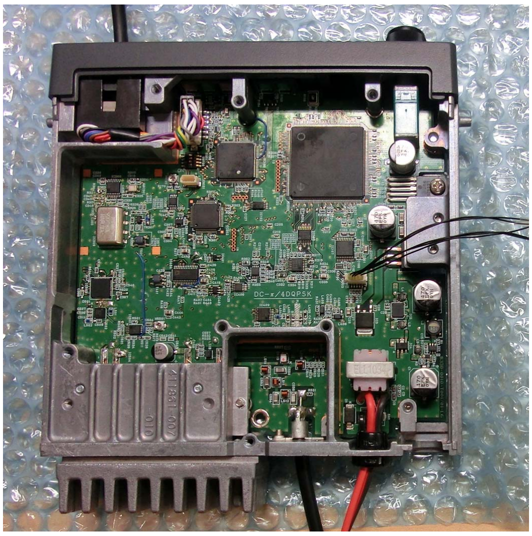

Figure 9. π/4 DQPSK prototype system