# A Modified Delay and Doppler Profiler based ICI Canceling OFDM Receiver for Underwater Multi-path Doppler Channel

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**Summary**
An Orthogonal Frequency Division Multiplexing (OFDM) based wireless communication system has drawn wide attention for its high transmission rate and high spectrum efficiency in not only radio but also Underwater Acoustic (UWA) applications. Because of the narrow sub-carrier spacing of OFDM, orthogonality between sub-carriers is easily affected by Doppler effect caused by the movement of transmitter or receiver. Previously, Doppler compensation signal processing algorithm for Desired propagation path was proposed. However, other Doppler shifts caused by delayed Undesired signal arriving from different directions cannot be perfectly compensated. Then Receiver Bit Error Rate (BER) is degraded by Inter-Carrier-Interference (ICI) caused in the case of Multi-path Doppler channel. To mitigate the ICI effect, a modified Delay and Doppler Profiler (mDDP), which estimates not only attenuation, relative delay and Doppler shift but also sampling clock shift of each multi-path component, is proposed. Based on the outputs of mDDP, an ICI canceling multi-tap equalizer is also proposed. Computer simulated performances of one-tap equalizer with the conventional Time domain linear interpolated Channel Transfer Function (CTF) estimator, multi-tap equalizer based on mDDP are compared. According to the simulation results, BER improvement has been observed. Especially, in the condition of 16QAM modulation, transmitting vessel speed of 6m/s, two-path multipath channel with direct path and ocean surface reflection path; more than one order of magnitude BER reduction has been observed at CNR=30dB.
***Keywords:***
*Underwater Acoustic Communication, OFDM, Channel Estimation, modified Delay and Doppler Profiler*

## I. Introduction

Underwater wireless communication is demanded for many applications such as surveillance of areas as harbors, ports and coastlines or monitoring of fishes and excavation sites such as oil well, trenches and so on to reduce cable cost or time consuming deploy [1]. In order to increase data Bandwidth, Orthogonal Frequency Division Multiplexing (OFDM) based wireless communication system has drawn wide attention for its high transmission rate and high spectrum efficiency even in Underwater applications. The moving of equipment and/or surface ship causes Doppler shift effect during underwater wireless communication, so the receiver of each side is necessary to have the signal processing for Doppler shift compensation capabilities. Previously, Doppler compensation signal processing algorithm for Desired propagation path was proposed [2-4].

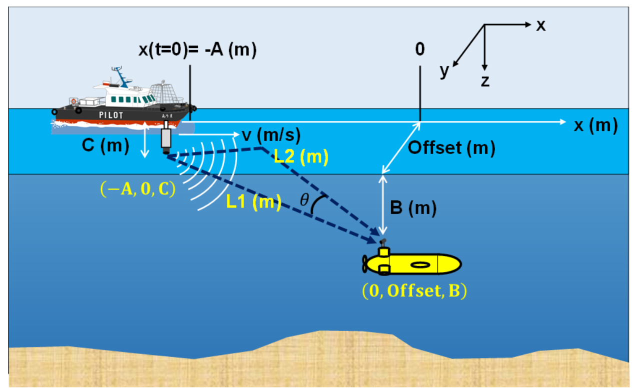


Fig. 1: UWA communication through Multi-path Doppler channel

Fig.1 shows UWA communication scene with multi-path propagation path. Because Desired path L1 and delayed Undesired path L2 are different direction at AUV receiver, different Doppler shifted arriving signals are multiplexed. Even when L1 path Doppler effect is compensated, the difference between L1 and L2 Doppler still remain. Then Receiver Bit Error Rate (BER) is degraded by Inter-Carrier-Interference (ICI) caused by the difference. In the previous paper [5], Delay and Doppler Profiler (DDP), which estimates attenuation, relative delay and Doppler shift of each multi-path component, is used to estimate more accurate Channel Transfer Function (CTF) [6-8].

In this paper, in order to mitigate the ICI, a modified Delay and Doppler Profiler (mDDP), which estimates

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Table 1: UWA OFDM System Parameters

| Parameters | Value |
|---|---|
| Sampling Frequency Fs | 102.4kHz |
| Band Width | 8 kHz |
| Passband frequency | 16 kHz |
| FFT size | 2048 |
| OFDM symbol length T | 20.0 ms (2048 points) |
| Guard Interval length Tg | 5.0 ms (512 points) |
| Sub-carrier spacing | 50 Hz |
| Number of sub-carrier | 161 |
| Scattered pilot | 81 every 2 OFDM symbols |
| Continuous pilot | 13 |
| Carrier modulation | QPSK/16QAM/64QAM |

not only attenuation, relative delay and Doppler shift but also sampling clock shift of each multi-path component, is proposed. Based on the outputs of mDDP, an ICI canceling multi-tap equalizer is also proposed. In section II, first, UWA OFDM Communication System is shown. Then proposed mDDP method is explained. Computer simulation results are shown in section III. Finally, in section IV, conclusions will be given.

## II. mDDP based ICI Canceling UWA OFDM Communication system

UWA OFDM Communication system is basically extension of the system proposed in paper [3]. Detail system parameters are listed in Table I. The sub-carrier spacing of OFDM is 50Hz, then effective OFDM symbol duration is 20ms with 2048 points IFFT/FFT as modulation/demodulation. To estimate Channel Transfer Function (CTF), two kinds of pilot signals are inserted in Time-Frequency sub-carriers' arrangement as shown in Fig. 2. The blue circles for CTF estimation are scattered pilots which is inserted every 2 row and every 2 columns. The other yellow circles are continuous pilots for tracking phase-change in time domain.

### A) *Time-domain pre-processing of Receiver*

The upper side of Fig. 3 (before 3rd FFT) is the time-domain pre-processing block of the Receiver. Since 8kHz of OFDM signal bandwidth is not enough, smaller

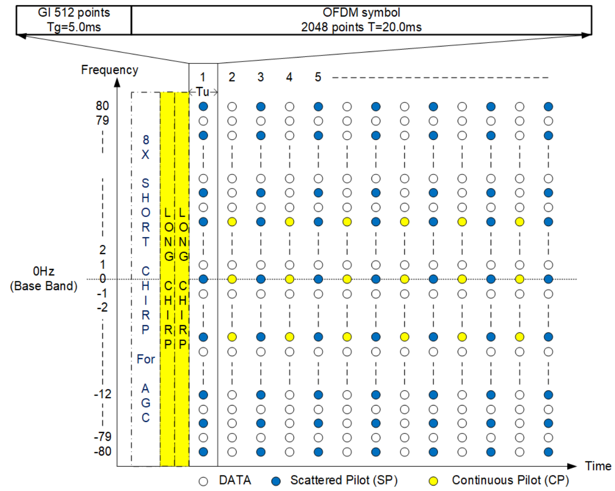



Fig. 2: Time-Frequency structure of OFDM signal

than 16kHz of Passband frequency narrowband approximation cannot be assumed. Then to compensate Doppler effect, the length of received Baseband signal needs to be shrunk to expanded. 1st dashed line block named Resampled and De-rotated is the functions for detecting shrink-expansion factors $\beta 1$ for coarse and $\beta 2$ for fine and signal resample and de-rotate. Detail explanation can be found in [3]. 2nd dashed line block named Phase shift compensation is the function for time-domain phase change detect and its compensation.

Fig.4 shows two cases of OFDM signal after 1st dashed line Shrink and Expansion processing. In Fig.4(a) of Single-path case, the 1st row shows 3 OFDM symbols with no movement and the 2nd row shows the case of received signal that appears to be stretched by moving reception. 3 descriptions in the rectangle box corresponds to the function of 1st dashed line block. After the 1st processing, the received OFDM signal is shrunk to the same length of no move case. In Fig.4(b) case of

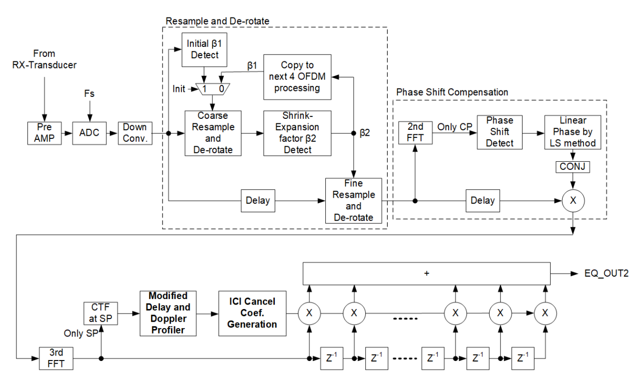
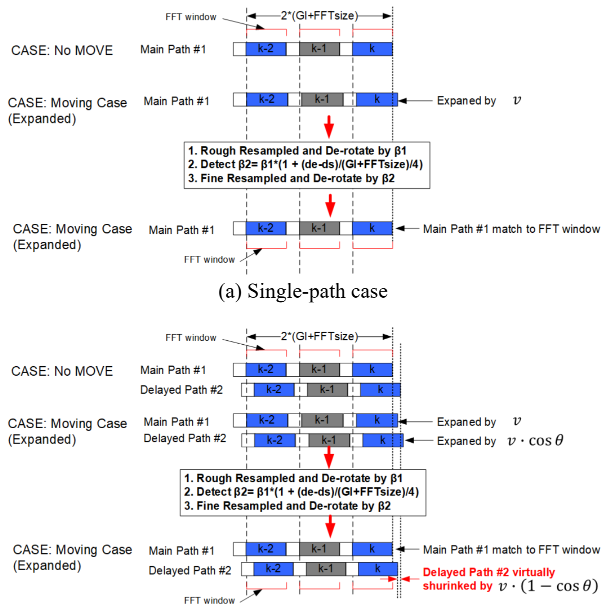



Fig. 3: Receiver Block Diagram with ICI Canceling Equalizer




(b) Case of Multi-path with different Doppler Shift

Fig. 4: OFDM signal after the Shrink and Expansion processing

multi-path with different Doppler Shift, Delayed path #2 is added to Fig.4(a). Since expansion factor of path #2 is different from that of path #1 by $\cos\theta$, Delayed path #2 OFDM symbols shown in the bottom row does not match with that of no move case. Then the Delayed path #2 disrupts orthogonality between OFDM sub-carriers and Inter-Carrier-Interference (ICI) between data sub-carriers are introduced.

### B) *ICI canceling by modified Delay and Doppler Profiler*

References [6-8] describes detail function of Delay and Doppler Profile (DDP) using pilot signals embedded in OFDM sub-carriers. DDP estimates attenuation, relative delay and Doppler shift of each multi-path component.

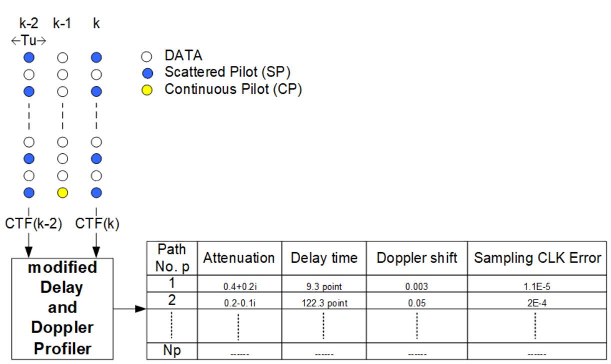



Fig. 5: Modified Delay and Doppler Profiler

However, as explained in Fig.4(b), the Delayed Path #2 causes ICI by its remaining shrink or expansion. Then for UWA system needs additional shrink or expansion parameter of each multi-path component.

Consequently, modified DDP is proposed to estimate not only attenuation, relative delay and Doppler shift but also sampling clock shift of each multi-path component.

Fig. 5 is a block diagram of proposed mDDP. The upper side of the figure represents the Time-Frequency structure of sub-carriers with time-domain symbol index k. The mDDP estimates each Np multipath component waves to receiver. Each analyzed component can be characterized using four parameters such as Attenuation $r_p$, Propagation Delay time $\tau'_p$, normalized Doppler shift $fd_p$ and Sampling CLK error $\beta_p$ for wave component index $p$. Here the Attenuation $r_p$ is complex value including amplitude attenuation and phase rotation and Doppler shift is normalized by sub-carrier spacing $f_0$ such as $\alpha_p = fd_p/f_0$. Using the symbol k measured CTF(k) and symbol k-2 measured CTF(k-2), mDDP detects $N_P$ sets of those four parameters.

UWA OFDM baseband transmitting signal $s_B(t)$ can be expressed as (1).

$$s_B(t) = \sum_{m=-\infty}^{\infty} g(t - mT_s) \cdot \sum_{n=0}^{N-1} d(m,n) e^{j2\pi n f_0 (t - mT_s)} \cdots (1)$$

$$T_s = \frac{1}{f_0} + T_g \cdots (2)$$

$$g(t) = \begin{cases} 1 & -T_g \leqq t < 1/f_0 \\ 0 & otherwise \end{cases} \cdots (3)$$

where, N is the number of sub-carriers, $f_0$ is the sub-carrier spacing, $T_s$ is symbol length as defined in (2), $d(m,n)$ is data symbol of sub-carrier index n at the $m^{th}$ OFDM symbol. By assuming the transmission channel has $N_P$ delay paths, the received baseband signal can be written as (4).

$$r_B(t) = \sum_{p=1}^{N_P} r_p\, s_B(t - \tau_p) e^{j2\pi \Delta f_p (t - \tau_p)} \cdots (4)$$

where, $r_p$, $\tau_p$ and $\Delta f_p$ are attenuation, relative delay and Doppler-shift of $p^{th}$ path respectively.

$$t = \frac{i}{N f_0}(1 + \beta) + kT_s(1 + \beta) \cdots (5)$$

After sampling of $r_B(t)$ by using equation (5) with sampling index $i$ and OFDM symbol number $k$, the block of samples is de-modulated by DFT to generate data symbol $\hat{d}(k,l)$ as expressed in (6).

$$\hat{d}(k,l) = h(k,l,l)d(k.l) + \sum_{\substack{n=0 \\ n\neq l}}^{N-1} h(k,l,n)d(k.n) + w(k,l) \cdots (6)$$

where $w(k,l)$ is additive noise that corresponds to the $l^{th}$ sub-carrier in the $k^{th}$ OFDM symbol and $h(k,l,n)$ is the transfer function from symbol $d(k,n)$ to the $l^{th}$ sub-carrier. If $n \neq l$, $h(k,l,n)$ represents the influence of ICI. $h(k,l,n)$ can be expressed as (7), (8a) and (8b).

$$h(k,l,n) = \sum_{p=1}^{N_P} h_p(k,l,n) \cdots (7)$$

$$h_p(k,l,n) = \frac{sinc(A)}{sinc\left(\frac{A}{N}\right)} \cdot e^{j\frac{\pi(N-1)A}{N}} \cdot r_p \cdot e^{-j\frac{2\pi(n+\alpha_p)\tau_p'}{N}} \cdot e^{j\frac{2\pi k(N+GI)(\alpha_p+n\beta_p)}{N}} \cdots (8a)$$

$$A = n - l + n\beta_p + \alpha_p(1+\beta_p) \cong n - l + n\beta_p + \alpha_p \cdots (8b)$$

where $\alpha_p = \Delta f_p / f_0$ is normalized frequency offset of $p^{th}$ path.

The parameters of channel transfer function $h(k,l,n)$ are estimated so as to minimize the mean square error shown in (9).

$$E(k) = \sum_{l=\boldsymbol{P}} \left\{ \left| \frac{x(k,l)}{d(k,l)} - h(k,l,l) \right|^2 + \left| \frac{x(k-2,l)}{d(k-2,l)} - h(k-2,l,l) \right|^2 \right\} \cdots (9)$$

where $\Sigma_{l=\boldsymbol{P}}$ means that the summation is performed as long as $l$ is scattered pilot symbol.

In order to simplify the problem, it is supposed that the channel model contains one path at the first. In this case, criterion is rewritten as (10).

$$E_1(k) = \sum_{l=\boldsymbol{P}} \left\{ \left| CTF(k,l) - f(\alpha_1,\beta_1,l) r_1 \cdot e^{-j\frac{2\pi(l+\alpha_1)\tau_1'}{N}} \right|^2 + \left| CTF(k-2,l) - f(\alpha_1,\beta_1,l) r_1 \cdot e^{-j\frac{2\pi(l+\alpha_p)\tau_1'}{N}} \cdot e^{j\frac{-2\pi 2(N+GI)(\alpha_1+l\beta_1)}{N}} \right|^2 \right\} \cdots (10)$$

where $f(\alpha_1,\beta_1,l)$ is written as (11)

$$f(\alpha_1,\beta_1,l) = \frac{sinc(l\beta_1+\alpha_1)}{sinc\left(\frac{l\beta_1+\alpha_1}{N}\right)} \cdot e^{j\frac{\pi(N-1)(l\beta_1+\alpha_1)}{N}} \cdot e^{j\frac{2\pi k(N+GI)(\alpha_1+n\beta_1)}{N}} \cdots (11)$$

One of the necessary conditions to minimize $E_1(k)$ with regard to $r_1$, $\tau_1'$ (relative delay time in sampling points) and $\alpha_1$ is that its partial derivative of $r_1$ has zero value. With this condition, $r_1$ can be shown as (12a) and (12b).

$$r_1 = \frac{S_k^* + \cdot e^{j\frac{2\pi 2(N+GI)\alpha_1}{N}} S_{k-2}^*}{2 \cdot e^{-j\frac{2\pi\alpha_1\tau_1'}{N}} \sum_{l=\boldsymbol{P}} |f(\alpha_1,\beta_1,l)|^2} \cdots (12a)$$

$$S_k = \sum_{l=\boldsymbol{P}} CTF^*(k,l) f(\alpha_1,\beta_1,l)\, e^{-j\frac{2\pi l\tau_1'}{N}} \cdot e^{j\frac{2\pi(k)(N+GI)l\beta_1}{N}} \cdots (12b)$$

By calculate more as shown in reference [6-8], $\alpha_1$ is calculated from $\tau_1'$ as (13).

$$\alpha_1 = \frac{1}{\frac{2\pi 2(N+GI)(1+\beta_1)}{N}} arctan\left[\frac{\Im(S_k^* \cdot S_{k-2})}{\Re(S_k^* \cdot S_{k-2})}\right] \cdots (13)$$

Consequently, once $\tau_1'$ is determined, $r_1$ and $\alpha_1$ are easily obtained from (12a) and (13), respectively. In this proposed method, $\tau_1'$ is changed by $1/(2Nf_0)$ (twice of sampling frequency) and rough estimation is first obtained. And this rough estimation is modified to more precise value using Newton method as described in references [6-8].

According to the computation above, $h_1(k,l,n)$ can be obtained using equations (8a) and (8b). Then, measured $CTF(k,l)$ and $CTF(k-2,l)$ are updated to $CTF(2,k,l)$ and $CTF(2,k-2,l)$ by subtracting $h_1(k,l,n)$ and $h_1(k-2,l,n)$ respectively. By removing the influence of the first estimated path from the cost function$E_1(k)$, the following criterion $E_2(k)$ can be obtained such as (14).

$$E_2(k) = \sum_{l=\boldsymbol{P}} \left\{ \left| CTF(2,k,l) - f(\alpha_2,\beta_2,l) r_2 \cdot e^{-j\frac{2\pi(l+\alpha_1)\tau_2'}{N}} \right|^2 + \left| CTF(2,k-2,l) - f(\alpha_2,\beta_2,l) r_2 \cdot e^{-j\frac{2\pi(l+\alpha_2)\tau_2'}{N}} \cdot e^{j\frac{-2\pi 2(N+GI)(\alpha_2+l\beta_2)}{N}} \right|^2 \right\} \cdots (14)$$

Since $E_2(k)$ contains the influence of other paths, the second path can be estimated using the same procedure in the previous section. By repeating this operation by the number of paths in the channel model, $N_p$ sets of those four parameters can be obtained as shown in Fig. 5.

After all, complete channel transfer function $h(k, l, n)$ can be generated. Since most energy of ICI is concentrated in the neighborhood of sub-carrier index, the ICI terms which do not significantly affect $\hat{d}(k, l)$ in (6) can be neglected and it is assumed as (15).

$$h(k, l, n) = 0 \quad when \;\; |l - n| > \frac{NumCOL - 1}{2} \quad \cdots (15)$$

where, $NumCOL$ is the number of neighborhood sub-carrier taken into account against $l^{th}$ sub-carrier. By considering the reverse operation of the ICI effect by $h(k, l, n)$, Multi-Tap ICI Canceling Equalizer has been implemented as shown in Fig. 3. as a Fixed Impulse Response Filter.

## III. Computer Simulations

In order to verify the proposed mDDP based ICI canceling, Fig.1 situation is modeled by Matlab.

Table 2: Simulation Parameters

| Parameters | Value |
|---|---|
| Vessel starting position | A=30m |
| RX transducer depth | B=20m |
| TX transducer depth | C=2m |
| TX-RX minimum horizontal distance | Offset=15m |
| Vessel Speed | v= 6, 3 m/s |
| Desired(L1) Undesired(L2) Ratio | 6dB |

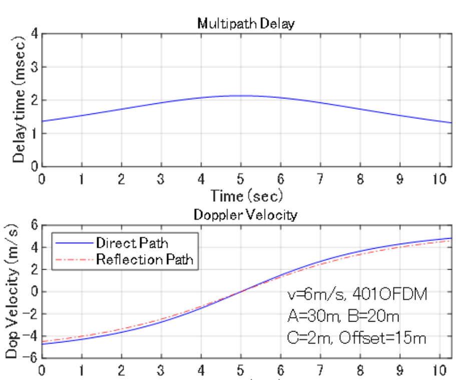
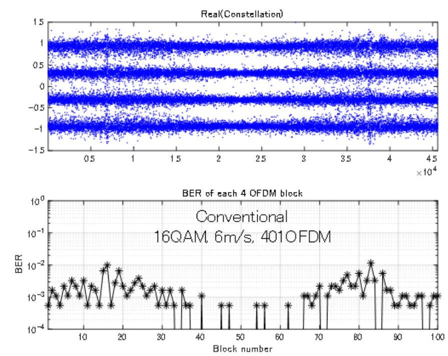


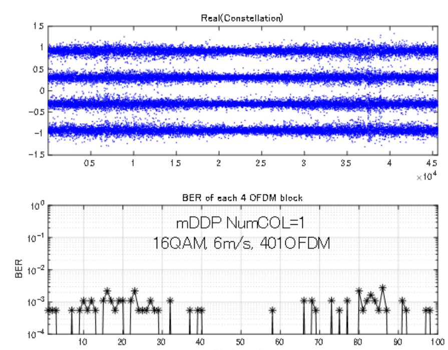

Fig. 6: Multi-path delay and Doppler velocities

Transmitting transducer of depth C=2m moves x-axis direction with speed of v(m/s). Receiving transducer is located at the depth of B=20m. More detail simulation parameters are summarized in Table 2.

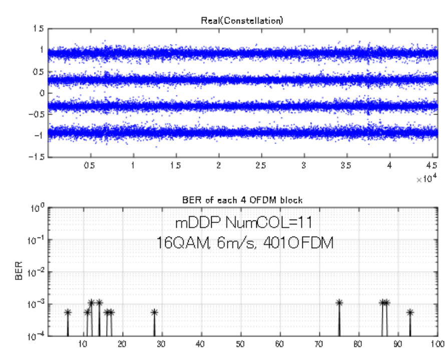
Fig.6 shows delay time between Direct path L1 and Reflection path L2 path and each Doppler velocity, which is effective direction-oriented velocity observed at receiver, for v=6m/s with 401OFDM symbol simulation. The delay time is roughly 1.4 to 2.1ms. Direct path Doppler velocity is slightly larger than it of Reflection path. Fig.7 shows the changes in time direction of real component of 16QAM constellation and BER. Fig.7(a)




(a) Conventional interpolator CTF equalizer

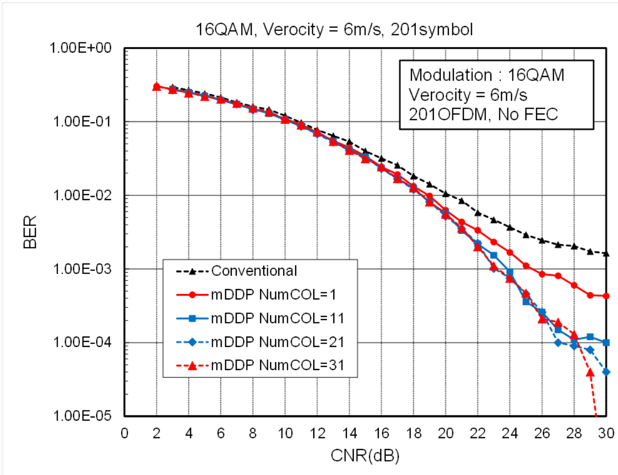



(b) mDDP with NumCOL=1 (No ICI Cancel)

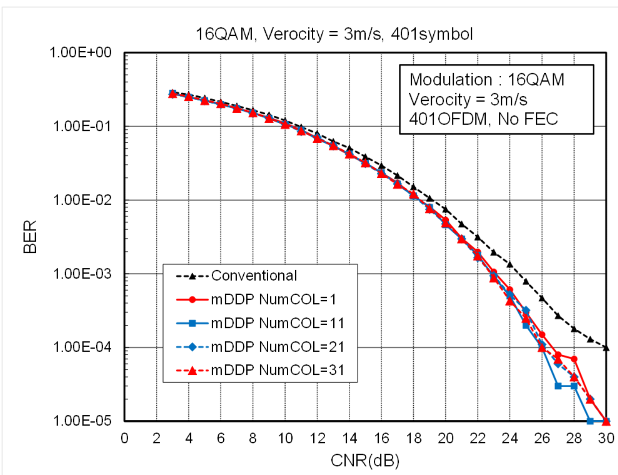



(c) mDDP with NumCOL=11 (ICI Cancel)

Fig. 7: Constellation (Real) and BER dependence on time

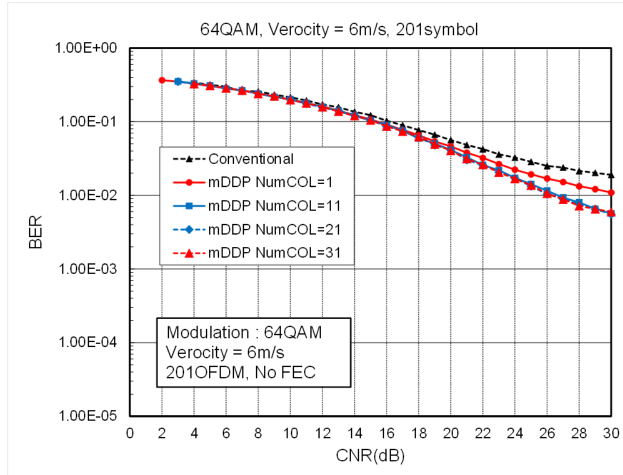



(a) v=6m/s

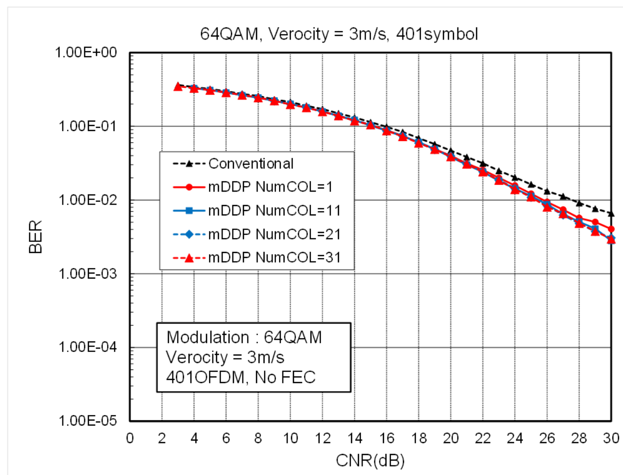



(b) v=3m/s

Fig. 8: BER vs CNR (dB) for 16QAM modulation




(a) v=3m/s




(b) v=3m/s

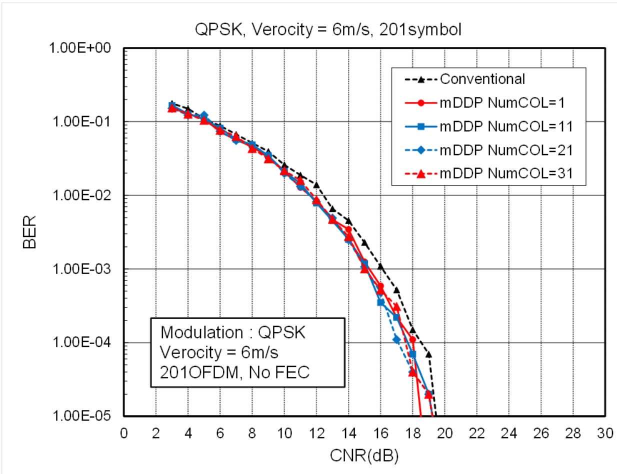
Fig. 9: BER vs CNR (dB) for 64QAM modulation

corresponds to the case of conventional linear interpolated CTF and one-tap equalizer. mDDP with NumCOL=1 means that CTF is estimated by mDDP but ICI cancel is not used as shown in Fig.7(b). Fig.7(c) shows the case of ICI canceling by NumCOL=11 multi-tap equalizer with mDDP. Obviously, BER is drastically improved by introducing mDDP and ICI canceling.

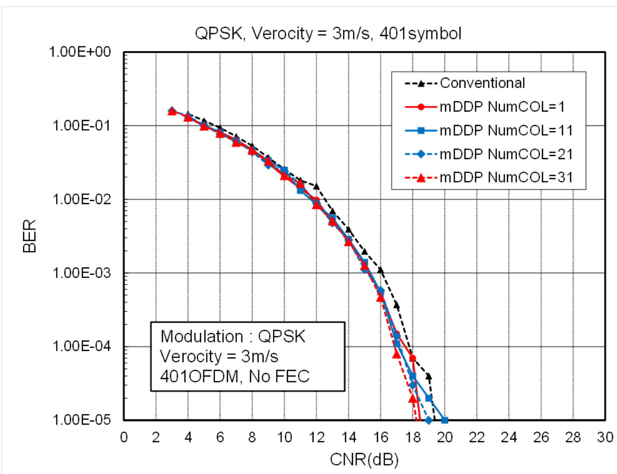
Fig.8 shows BER vs CNR(dB) for 16QAM with v=6m/s and 3m/s. Number of OFDM symbols used in the simulation are 201 and 401, which means the vessel moves from x= -30m to 0m. For v=6m/s, ICI canceling with 11 tap (NumCOL=11) effectively decrease BERs. However, larger NumCOL of 21 or 31 is not effective. For lower speed of v=3m/s, mDDP channel estimation without ICI cancel shows large BER reduction compare with conventional. Fig.9 and 10 show the same cases for 64QAM and QPSK modulation. For 64QAM, similar effectiveness of mDDP with ICI cancel is observed. For QPSK, only mDDP effectiveness is observed because of relatively low BER situation.

## IV. Conclusion

modified Delay and Doppler Profiler (mDDP), which estimates not only attenuation, relative delay and Doppler shift but also sampling clock shift of each multi-path component, is proposed. Based on the outputs of mDDP, an ICI canceling multi-tap equalizer is also proposed. Computer simulated performances of one-tap equalizer with the conventional Time domain linear interpolated Channel Transfer Function (CTF) estimator, multi-tap equalizer based on mDDP are compared.




(a) v=6m/s




(b) v=3m/s

Fig. 10: BER vs CNR (dB) for QPSK modulation

According to the simulation results, BER improvement has been observed. Especially, in the condition of 16QAM modulation, transmitting vessel speed of 6m/s, two-path multipath channel with direct path and ocean surface reflection path; more than one order of magnitude BER reduction has been observed at CNR=30dB.

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