Application of Analytic Receiving to the diffCDMA in the High Capacity Land Mobile Communication System and its Transmission Characteristics over Multi-Ray Rayleigh Fading Channel

解析受信の大容量陸上移動通信システム diffCDMA への適用と、そのマルチレイレイレーフェーディング伝播路特性

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Abstract: The differential coding CDMA (diffCDMA) is verified here to conquer the fatal transmission error when communications are carried from high speed running vehicles through such urban environment as rapid multi-ray Rayleigh channels. With bases of introducing differential coding into the primary modulation and employing analytic receiving for primary differential demodulation, the proposing diffCDMA improves transmission characteristics of such similar CDMA systems as IS95 and WcdmaOne IMT2000 when communications carried from more than 100 mile/h mobiles even if through multi-ray environment.

Previously, differential coding primary modulation is not considered to be able to stand on CDMA owing to requiring strict sense synchronization in demodulation, which is suffered from bit error under such rapid multi-ray Rayleigh fading. However, it becomes to be able in high speed and high capacity transmission after the analytic receiving being adopt into CDMA systems. In the demodulation procedure of the diffCDMA, received waves are at first detected after multi-ray propagation with certain inevitable errors both in phase and frequency of the recovered carrier for the pseudo synchronization primary demodulation during the analytic receiving.

1. INTRODUCTION

As known well, the analytic receiving has been developed for high capacity and high speed PSK digital transmission with such advantages as excluding selective frequency fading, superior receiving sensitivity, and etc. These superior characteristics are derived from the analytic signal processing.

When arbitrary real function is given by $f(t)$, the corresponding analytic signal $g(t)$ is spanned on the complex plane as,

$$g(t) = f(t) + j\,\hat{f}(t) \tag{1}$$

Here,

$$\hat{f}(t) \text{ is Hilbert transform of } f(t) \tag{2}$$

$$j \text{ is complex unit, } j = \sqrt{-1} \tag{3}$$

The instantaneous phase $\theta(t)$ and envelope $A(t)$ are given as follows.

$$\theta(t) = \tan^{-1}\frac{\hat{f}(t)}{f(t)} \tag{4}$$

$$A(t) = \sqrt{f^2(t) + \hat{f}^2(t)} \tag{5}$$

The analytic signal $g(t)$ is described as the polar system by using these instantaneous various.

$$g(t) = A(t)\, e^{j\theta(t)} \tag{6}$$

The real part of the analytic signal is corresponding to the existing modulating signals. That is,

$$Re\{g(t)\} = A(t)\cos\{\theta(t)\} \tag{7}$$

Here,

$\theta(t) = \omega_c t + \theta_o(t)$,

ω_c *is carrier angular frequency,*

$\theta_o(t)$ *is phase information at time* t.

Phase difference $\Delta\theta(t)$ between real function $f_i(t)$ and $f_k(t)$ is easily derived from a product of analytic $g_i(t)$ and conjugate $g^*_k(t)$ analytic signals independent with each other as follows.

$$\begin{aligned} & g_i(t)\, g^*_k(t) \\ &= \{A_i(t)e^{j\theta_i(t)}\}\{A_k(t)e^{j\theta_k(t)}\}^* \\ &= \{f_i(t) + j\hat{f}_i(t)\}\{f_k(t) - j\hat{f}_k(t)\} \\ &= \{f_i(t)\, f_k(t) + \hat{f}_i(t)\, \hat{f}_k(t)\} \\ &\quad + j\{\hat{f}_i(t)\, f_k(t) - f_i(t)\, \hat{f}_k(t)\} \end{aligned} \tag{8}$$

Here, i or k is code channel number, respectively.

Therefore, phase difference $\Delta\theta(t)$ between these i and k code channels is given as,

$$\begin{aligned} \Delta\theta_{ik}(t) &= \tan^{-1}\frac{Im\{g_i(t)\, g^*_k(t)\}}{Re\{g_i(t)\, g^*_k(t)\}} \\ &= \tan^{-1}\frac{\hat{f}_i(t)\, f_k(t) - f_i(t)\, \hat{f}_k(t)}{f_i(t)\, f_k(t) + \hat{f}_i(t)\, \hat{f}_k(t)} \end{aligned} \tag{9}$$

The analytic receiving for differential coding PSK has already suggested in equation 9. That is, substituting delayed function $f_i(t-T)$ into eq.9 instead of $f_k(t)$, phase difference gives demodulating information phase values with most precision time resolution as discussed in the next session. The in- or quadrature-channel signal is directly introduced from the real or imaginary part of the product, eq.8, respectively.

$$i_{ik}(t) = Re\{g_i(t)\ g^*_k(t)\} \quad (10)$$

$$q_{ik}(t) = Im\{g_i(t)\ g^*_k(t)\} \quad (11)$$

Substituting eq.1 into eqs.10 and 11 gives the other significant results of analytic receiving for detecting in- and quadrature channel signals as follows.

$$i_{ik}(t) = f_i(t) f_k(t) + \hat{f}_i(t) \hat{f}_k(t) \quad (12)$$

$$q_{ik}(t) = \hat{f}_i(t) f_k(t) - f_i(t) \hat{f}_k(t) \quad (13)$$

2. ANALYTIC RECEIVING IN PSK SYSTEMS

Fig.1 shows a comparison of BER vs. E_b / N_0 response between analytic receiving and existing synchronous detection for QPSK measured after propagation through such two-ray Rayleigh fading environment as 10 dB DUR with 1.0 micro second delay spread. Transmission quality is remarkably damaged over Nyquist limit bandwidth radio channel in the multi-ray propagation environment even if BER being superior to the analytic receiving by 3dB in the static additive white Gaussian noise (AWGN) environment. On the other hand, the multi-ray fading robustness is catastrophically improved by employing the analytic receiving over Nyquist limit bandwidth channel in the same environment from such high speed communications as from 100 mile/h running vehicle or 1,000 km/h flying aircraft.

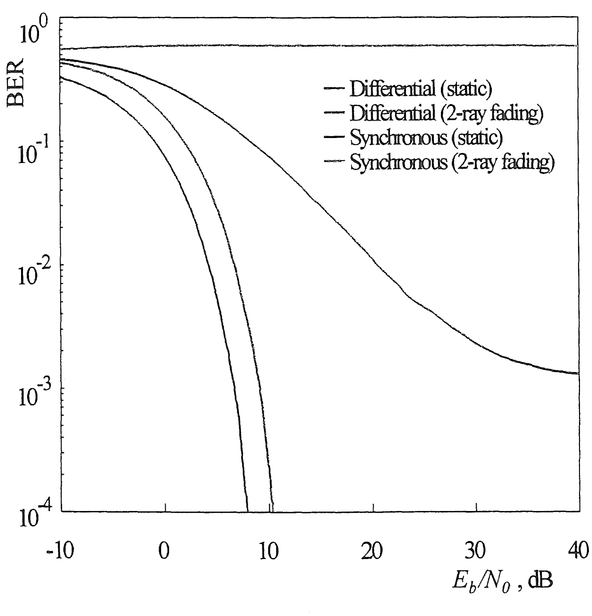

Fig. 1 Bit Error Rate performance of QPSK System

The phase difference $\Delta\theta(t)$ gives the demodulating angular phase by substituting receiving and delayed signals instead of two individual signals as follows,

$$\Delta\theta(t) = \tan^{-1} \frac{Im\{g(t)\ g^*(t-T)\}}{Re\{g(t)\ g^*(t-T)\}}$$

$$= \tan^{-1} \frac{\hat{f}(t)\ f(t-T) - f(t)\ \hat{f}(t-T)}{f(t)\ f(t-T) + \hat{f}(t)\ \hat{f}(t-T)} \quad (14)$$

Here, T is symbol duration.

As shown clearly in equations from 12 to 14, $i(t)$, $q(t)$ and $\Delta\theta(t)$ are successfully defined in excluding carrier frequency, and simultaneously described as full energy form with excluding frequency spectrum elimination. These equations will promise the advantages of tolerance both for frequency shift and multi-ray propagation. Circuitry configuration is directly illustrated for the analytic receiving from eqs.(10) and (11) as shown in fig.2, where mark H, D, MUL, or Σ means Hilbert transformer, delay by symbol duration T, multiplier, respectively.

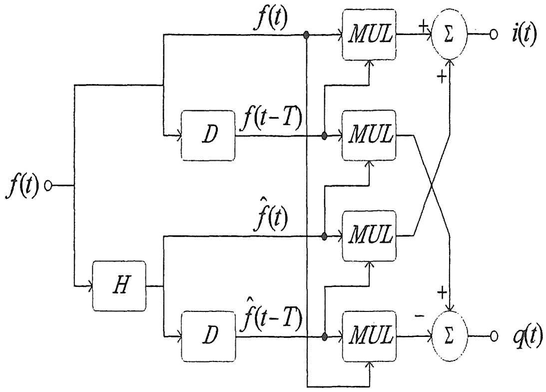

Fig.2 Circuitry configuration of the analytic receiving

Even if the instantaneous phase $\theta(t)$ is shifted by certain phase error $\theta_\delta(t)$ in some successive symbol periods during poor radio channel propagation, the differential demodulating phase $\Delta\theta(t)$ is precisely

$$\begin{aligned}\Delta\theta(t) &= \{\theta_o(t)+\theta_\delta(t)\} \\ &\quad -\{\theta_o(t-T)+\theta_\delta(t-T)\} \\ &= \theta_o(t)-\theta_o(t-T)\end{aligned} \tag{15}$$

Here, $\theta_o(t)$ is a priori information phase value at time t. $\theta_\delta(t) \cong \theta_\delta(t-T)$, because of fading phenomena being sufficiently slow to be quasi static in comparison with the transmission symbol rate.

Analytic receiving is also able to precisely demodulate receiving signal even if the frequency of the receiving signal is offset by ppm order from Doppler shift. That is, since the instantaneous phase being described as $\theta(t)=\theta_o(t)+\omega_c t+\omega_\delta(t)t$ in the PSK modulation, the demodulating phase $\Delta\theta(t)$ is given as,

$$\begin{aligned}\Delta\theta(t) &= \{\theta_o(t)+\omega_c t+\omega_\delta(t)t\} \\ &\quad -\{\theta_o(t-T)+\omega_c(t-T) \\ &\qquad +\omega_\delta(t-T)(t-T)\} \\ &= \theta_o(t)-\theta_o(t-T)+\omega_c T+\omega_\delta(t)T\end{aligned} \tag{16}$$

Here, $\omega_c T$ is given by $2m\pi$, m is a natural number. $\omega_\delta(t) \cong \omega_\delta(t-T)$, because of fading phenomena being also sufficiently slow to be quasi static in comparison with the transmission symbol rate. And $\omega_\delta(t)T$ can be ignored, because it is neglected zero.

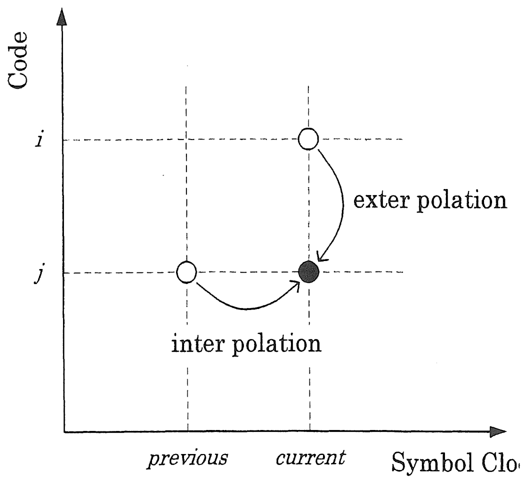

Fig. 3 Various different competition schemes for introducing the analytic receiving to CDMA System

4. ANALYTIC RECEIVING IN CDMA SYSTEMS

Pilot CDMA System

Differential detection shall be generally employed in vehicle communication system in order to avoid the fatal degradation of the synchronized detection in fading environment as shown in fig.1. This advantageous is seemed to induce from employing compensation between such adjacent pivots as adjacent symbols.

Innocently, since CDMA is spanned over two quadrature axes, two different compensation schemes are facilitated over individual axis as shows in fig.3. The one is spanned on the simultaneous clock along to code axis, the other is spanned on the same code axis along to the time axis.
The former will yield the fundamental compensation function of extrapolating pilot signal similar to the IS95 after substituting two individual receiving CDMA signals into eq.12 as follows,

$$i_{ik}(t) =$$

$$\left\{w_i(t)\sum_{l=1}^{m} w_l(t) f_l(t)\right\}\left\{w_k(t)\sum_{n=1}^{m} w_n(t) f_n(t)\right\}$$

$$+\left\{w_i(t)\sum_{l=1}^{m} w_l(t) \hat{f}_l(t)\right\}\left\{w_k(t)\sum_{n=1}^{m} w_n(t) \hat{f}_n(t)\right\} \quad (17)$$

$$Here,\ H\{w_l(t) f_l(t)\} = w_l(t)\hat{f}_l(t), \quad (18)$$

$w_l(t)$ *is the value of DS/SS l-th code at time t.*

Being accumulated within Walsh code length, eq.17 is modified as,

$$\tilde{i}_{ik}(t) = \tilde{f}_i(t)\tilde{f}_k(t) + \tilde{\hat{f}}_i(t)\tilde{\hat{f}}_k(t) \quad (19)$$

Here, $\tilde{f}_i(t)$, $\tilde{f}_k(t)$, $\tilde{\hat{f}}_i(t)$, *or* $\tilde{\hat{f}}_k(t)$ *is approximately given the averaged within the Walsh de-spreading code duration of* $f_i(t)$, $f_k(t)$, $\hat{f}_i(t)$, *or* $\hat{f}_k(t)$, *respectively.*

Quadrature channel component $\tilde{q}_{ik}(t)$ is similarly given as follows.

$$\tilde{q}_{ik}(t) = \tilde{\hat{f}}_i(t)\tilde{f}_k(t) - \tilde{f}_i(t)\tilde{\hat{f}}_k(t) \quad (20)$$

Both $\tilde{i}_{ik}(t)$ and $\tilde{q}_{ik}(t)$ of eqs.19 and 20 are clearly shown to be same to the analytic receiving for the PSK defined by eqs.13 and 14 only with exception of averaged value.

Differential CDMA System

The later will yields the most efficient frequency usage compensation function as the absolute solution for differential coding CDMA after de-spreading by substituting receiving CDMA and its delayed signals into eqs.13 and 14. The de-spreading in-channel signal $i_i(t)$ of code-i is given by

$$i_i(t) = \left\{w_i(t)\sum_{l=1}^{m} w_l(t) f_l(t)\right\} \bullet$$

$$\left\{w_i(t)\sum_{n=1}^{m} w_n(t-T) f_n(t-T)\right\}$$

$$+\left\{w_i(t)\sum_{l=1}^{m} w_l(t)\hat{f}_l(t)\right\}\bullet$$
$$\left\{w_i(t)\sum_{n=1}^{m} w_n(t-T)\hat{f}_n(t-T)\right\} \quad (21)$$

Therefore, averaged $\tilde{i}_i(t)$ is given by

$$\tilde{i}_i(t) = \tilde{f}_i(t)\tilde{f}_i(t-T) + \tilde{\hat{f}}_i(t)\tilde{\hat{f}}_i(t-T) \quad (22)$$

Quadrature channel component $\tilde{q}_i(t)$ is also given as follows.

$$\tilde{q}_i(t) = \tilde{\hat{f}}_i(t)\tilde{f}_i(t) - \tilde{f}_i(t)\tilde{\hat{f}}_i(t-T) \quad (23)$$

These eqs.22 and 23 also guarantee to introduce the novel diffCDMA with bases of the analytic receiving by employing two time different signals on the same code. All of the pilot signals are perfectly excluded in this diffCDMA to yield the most efficient frequency usage.

The diffCDMA is also significantly shown in the above to guarantee their accuracy by excluding vagueness both of phase and frequency induced during propagation over poor multi-ray channel owing to the genius of analytic receiving. In eqs.19 and 20, or eqs.22 and 23, the pair of $\tilde{i}_{ik}(t)$ and $\tilde{q}_{ik}(t)$, or pair of $\tilde{i}_i(t)$ and $\tilde{q}_i(t)$ holds the Hilbert transform relation with each other, respectively.

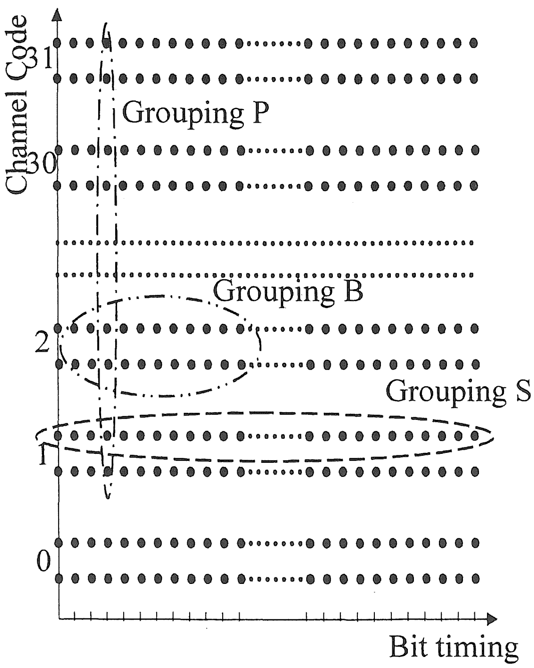

Fig. 4 ECC bit map comparison among typical grouping schemes on the plane spanned by time and code

Table 1 Radio Link Parameters.

		Differential System	Pilot System
Chip rate		4.096Mcps	4.096Mcps
Symbol rate		32ksps	32ksps
Carrier frequency		2GHz	2GHz
Modulation	Data	$\pi/4$ shifted DQPSK	QPSK
	Spreading	Walsh32	Walsh32
Error Correct		BCH(63,51)	BCH(63,51)

4. SIMULATION RESULTS

The analytic receiving is verified in such two CDMA systems as shown in fig.3 through computer simulation. Simulations are performed under conditions listed in table 1. All 1.66 Mbps signals are carried by 4.096Mcps on 4.096MHz frequency bandwidth at the 2GHz domain from 100mile/h vehicles running through such three environments as two-ray Rayleigh fading of DUR= 10dB with 1.0 micro second delay spread, flat fading, and static AWGN. The all codes of 32 length Walsh sequence are employed to span 32 code space. The BCH (63,51) ECC is employed to correct double errors within 63bit block along individual bit string of 32k symbol/second.

Fig.4 shows 3 kind bit mapping schemes where ECC being employed. Grouping S means the case of 12 redundant bits of BCH ECC being interpolated after every 51 information bits. Individual two BCH (63,51) block codes are simultaneously adopted into every spectrum spreading code along time bases. Grouping B is rather simple to adopt BCH with somewhat loss in transmission efficient, where 12 redundant bits are interpolated after every 50 information bits. Every shortened BCH (62,50) is employed for every DS/SS code along to time axis. The resulting grouping P is performed along to DS/SS code axis. That is, shortened BCH (62,50) is also convenient to introduce into diffCDMA, because one of 32 code channels being devoted to so-called controlling signal. Therefore, $f_d T$ and Doppler shift are set to be 0.008 and 0.16 ppm in the all simulations, respectively. Such compensations as RAKE receiving and power control are easily facilitated in the diffCDMA, which are excluded from simulations to make the analytic receiving effect clear. E_b / N_0 is measured on the maximum transmission capacity of 1.66Mbps when all the 32 code being employed in the differential CDMA system as the both cases for up and down link. The pilot CDMA system is measured on the down link as the maximum transmission capacity of 1.60Mbps when one code being devoted for common pilot and the left 31 codes being employed for 31 individual information channel. Simulation results are shown in fig.5 to be theoretically expected in receiving reliability through static environment, where the differential CDMA is degraded by 3 dB in comparison with the pilot CDMA owing to doubled noise power interfered over adjacent symbols.

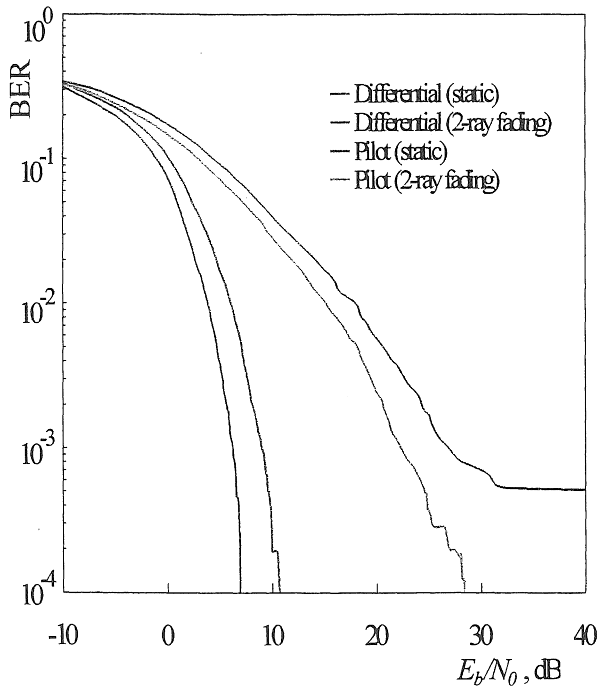

Fig. 5 Bit Error Rate performance of CDMA System

Simultaneously, BER vs. E_b / N_0 responses are splendid as shown in the same figure measured from 100mile/h running vehicles passing through flat fading or two-ray fading environment. Computer simulations are successfully verified to be almost error free at E_b / N_0=10dB for 1.66 Mbps high capacity and 4.096 Mcps of the diffCDMA through such poor propagation channel as two-ray Rayleigh fading of *DUR*=10dB and 1.0 micro-sec delay spread from 100 mile/h running mobile with employing BCH(63,51) double error correction. The frequency usage efficiency is observed to be 0.40 bit/Hz, and BER is zero at E_b / N_0 =10dB compensated for spreading gain where 64 bits being devoted to error free transmission.

CONCLUSION

The diffCDMA has been successfully proposed with emphasis on maximization both the transmission reliability and capacity. Such high reliability as error free is examined from 100mile/h running vehicles through poor radio propagation channel of transmitting 1.6 Mbps CDMA to high speed running vehicles. It is clearly shown in the computer simulations that the proposing diffCDMA is able to put megbit/second order high capacity and 100 mile/h high-speed mobile CDMA systems on the developing stages. The transmission capacity has extended by twice with owing to implicit interpolating signal based on differential coding instead of employing pilot signals of extrapolation. The frequency usage efficiency will be improved with employing such techniques as continuous phase primary modulation, continuous chip shaping secondary modulation, virtual segment interleaving, power control, and RAKE receiving. The circuitry complexity will be effectively reduced in diffCDMA, which is implemented in the isomorphic topology between up and down links owing to the novel analytic receiving.

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