# 레일리 페이딩 채널에서의 SFH를 적용한 2비트 차동검출 GMSK셀룰라 시스템의 성능분석

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# Performance of Cellular System with SFH and 2-bit Differentially Detected GMSK over Rayleigh Fading channel

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요 약

본 논문에서는 레일리 페이딩 채널환경에서 느린 주파수 도약/코드 분할 다중 접속(SFH/CDMA: slow frequency hopping/code division multiple access) 방식을 사용하는 2-비트 차동 검출 GMSK(gaussian minimum shift keying) 시스템의 비트오율 성능에 대해 연구하였다. 주파수 도약시 충돌이 발생하여 동일채널 간섭을 가지게 되는 검출기 출력 신호에 대하여 특성 함수를 유도하여 BER 결과를 얻었으며, 주파수 도약을 적용하지 않는 기존의 TDMA(time division multiple access) 시스템에 비해 주파수 도약을 사용하는 TDMA 시스템이 향상된 성능을 얻을 수 있음을 보였다.

ABSTRACT

In this paper, we study the bit error rate(BER) performance of 2-bit differentially detected GMSK system with the slow frequency hopping code division multiple access(SFH/CDMA) over flat Rayleigh fading channel. We can obtain the BER results using the characteristic function of demodulator's output, which has cochannel interferences(CCI) while a hop being hit, and show that the TDMA system with frequency hopping has improved performance compared to the conventional TDMA system without frequency hopping.


## I. Introduction

Recently, to increase capacity and to improve service features, the digital communication system have made a rapid progress. And, with the development of communication techniques, we can communicate wirelessly through a radio channel. The radio channel is characterized by band-limited and power-limited system, and have fading depairments which are caused by the movement of mobile and the multipath propagation of transmitted signals. Therefore, for band-limited characteristic, we ought to use a high spectral efficient modulation technique, and for power-limited characteristic, modulated signal must have constant envelope to use nonlinear amplifiers. The multipath fading, that is, the distortions of signal's envelope and phase depend on the frequency of carrier and the velocity of mobile. Generally, the peak and valley of fade appear every half a wave length, so these make burst bit errors. In practical system, various diversity schemes are used to overcome hostile fading channel environments. The diversity schemes provide two or more uncorrelatedly faded inputs at receiver, so that the receiver use combining technique and obtain better performance than that of the receiver using a

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논문번호 : T00036-0829, 접수일자 : 2000년 8월 29일

single input.

The CDMA system, which is broadly referred to as DS/CDMA system and FH/CDMA system, has larger system capacity and provide better performance than conventional system in hostile channel environments such as multipath fading, jamming etc. FH/CDMA system provides benign effects by frequency diversity and interferer diversity with using frequency-hopping. By the frequency diversity, we can overcome the delay problem in the receiver when the system uses interleaving technique. Using independent signals which are separated over coherence bandwidth, we can obtain frequency diversity. In FH/CDMA system, the user's hop frequency is shifted to next hop frequency over coherent bandwidth, so the hop-to-hop inputs are independent. Also, the system capacity increases due to the interferer diversity and error correction coding(ECC). If two or more users' frequency is shifted to a specific hop frequence, it is referred to "hit". The hit makes cochannel interference(CCI). But, the error bits of hit hops can be randomized because each user has a specific hopping sequence which differs from others and can be corrected by ECC.

## Ⅱ. System model

### 2.1 Fractional reuse structure with mixed hopping sequences

According to the stochastical property of intercell's sequences or intracell's, frequency hopping sequences are usually classified to orthogonal, random and mixed sequences.

First, orthogonal sequences have no hit within a cell and are allocated distinct sets of frequencies within neighboring cells. FH/CDMA system with orthogonal sequences is the same as FDMA system, except for the inherent frequency diversity. Second, for FH/CDMA system with random sequence, each active user has a unique hop sequence which is uncorrelated with, but not necessarily orthogonal to, all other users' sequences. Thus, this less stringent constraint of hopping sequences' cross correlation gives interferer diversity. Since hopping sequences

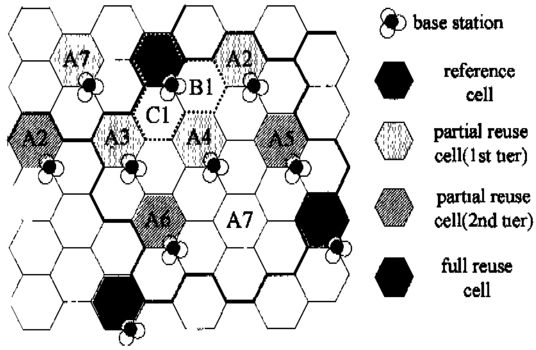



Fig. 1 The cell structure of 3M/L fractional reuse FH/CDMA with mixed sequence pattern.

spread interfered signals evenly over all available frequencies, FH/CDMA system can increase the system capacity with using ECC. Third, mixed sequences are orthogonal within intercell and uncorrelated between intracells. The mixed sequences have the advantage of frequency diversity and interferer diversity, and offer more freedom for system design optimization when used in conjunction with fractional reuse structures.

In this paper, we consider $3M/L$ fractional reuse structure, that is, base station are 3 sectorized by directional antennas, and the reuse cluster contains $3M$ cells arranged as $M$ sub-clusters of size 3. The cells A, B, C in fig. 1 are referred to "color", which use the distinct sets of $N$ frequencies. Each color is divided into the $M$ cells of which cells have the $L$ overlapping subsets of $M$ frequency groups and referred to as "shade" of that color.

Between different color cells, there is no hit and the fractional overlap coefficient $h$ is zero. And, ignoring the interference from the full reuse cells, all fractional overlap coefficient $h$ are 1/3 in considered shade cells. A TDMA frame is constituted of $N_t$ TDM slots. Assuming that all cells have $k$ active users and that silence detection with no transmission during silent periods is implemented to reduce cochannel interferences, the collision probability of a hop can be expressed as

$$P_{coll}(k) = \frac{Mf_s h}{LW/3}\frac{Ak}{N_t} \quad (1)$$

Where $A$ is speech activity ratio and $W$ is the total bandwidth of system. Using eq. (1), we obtain the probability of $m$ hits in $k$ active users and $M$ interference cells.

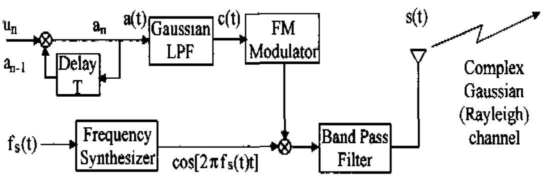
$$P_{hit}(M, m, k) = \binom{M}{m} P_{coll}(k)^{m} \{1 - P_{coll}(k)\}^{M-m} \tag{2}$$

Consequently, the equation of bit error rate in FH/CDMA system is expressed as

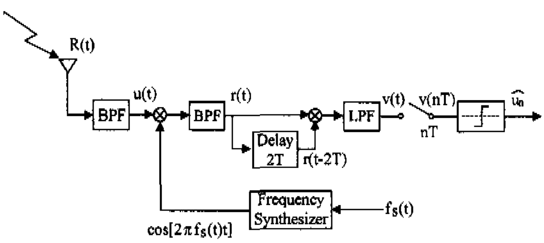
$$\begin{aligned} P_e(k) &= P(\text{error} \mid 0 \text{ hit})(1 - P_{hit}(M, M, k)) \\ &\quad + \sum_{m=1}^{M} P(\text{error} \mid m \text{ hits}) P_{hit}(M, m, k) \end{aligned} \tag{3}$$

## 2.2 Transmitter Model

Fig. 2 shows the transmitter and receiver model of FH/CDMA system with 2-bit differentially detected GMSK.




(a) FH/CDMA-GMSK Transmitter,




(b) FH/CDMA-GMSK Receiver with 2-bit differentially detection,

Fig 2. FH/CDMA-GMSK Transmitter and Receiver model.

$u_{s,n}$ is the $n$th antipodal input bit of users, and $a_{s,n}$ is the output of differentially encoder. For 2-bit differential detection, the differential encoding rule is given by

$$a_{s,n} = - u_{s,n} a_{s,n-1} \tag{4}$$

The output of Gaussian low pass filter(GLPF) can be expressed as,

$$p(t) = \sum_{n=-\infty}^{\infty} a_{s,n}\, g(t - nT), \tag{5}$$

where $T$ is symbol duration and $g(t)$ is the response of GLPF to a unit rectangular pulse as follows;

$$\begin{aligned} g(t) &= \Omega\left(\frac{t}{T}\right) * h(t) \\ &= \sqrt{\frac{2\pi}{\ln 2}} B_t \int_{t-T}^{t} \exp\left(-\frac{2\pi^2 B_t^2}{\ln 2} x^2\right) dx \\ &= \frac{1}{2}\left[ erf\left\{\sqrt{\frac{2}{\ln 2}}\, \pi B_t T\left(\frac{t}{T}\right)\right\} \right. \\ &\quad \left. - erf\left\{\sqrt{\frac{2}{\ln 2}}\, \pi B_t T\left(\frac{t}{T} - 1\right)\right\}\right], t > 0 \end{aligned} \tag{6}$$

where $B_t$ is the 3dB-bandwidth of GLPF and

$$erf(y) = \frac{2}{\sqrt{\pi}} \int_0^y \exp(-u^2) du = -erf(-y) \tag{7}$$

The output of FM modulator is given by

$$s(t) = \cos(2\pi f_c t + \phi_s(t)) \tag{8}$$

The user's signal is transmitted through complex gaussian channel(that is, Rayleigh fading channel) after the carrier frequency is shifted to the frequency $f_s(t)$ by frequency synthesizer according to his own hopping sequences. Thus, the transmitted signal can be expressed as

$$s(t) = \cos\{2\pi(f_c + f_s(t))t + \phi_s(t)\} \tag{9}$$

Likewise, the interference signal from $l$th user can be expressed as

$$\begin{aligned} i_l(t) &= \cos\{2\pi(f_c + f_{i,l}(t))t + \phi_{i,l}(t)\} \\ &= \cos\{2\pi(f_c + f_s(t))t + \phi_{i,l}(t)\} \end{aligned} \tag{10}$$

where $f_{i,l}(t)$ is equal to $f_s(t)$ since hit happens. And, the modulated phase of interference signal is given by

$$\phi_{i,l}(t) = 2\pi f_d \int_{-\infty}^{t} \sum_{n=-\infty}^{\infty} a_{l,n}\, g(\mu - nT) d\mu \tag{11}$$

where the differential encoded data of interference signal $a_{l,n}$ are supposed to be independent with one another.

### 2.3 Receiver

The received signal is despreaded by the dehopper with the same hopping sequence as the user hopping sequence. Assuming that white Gaussian noise is added to received signal and that Gaussian bandpass filter is used before differential detection, the total input to the differential detector is given by

$$r(t) = s(t) + \sum_{l=0}^{m} i_l(t) + n(t) \tag{12}$$

where

$$s(t) = \mathrm{Re}\{Z_s(t)\exp[j2\pi f_c t + \phi_s(t)]\} \tag{13}$$

$$i_l(t) = \mathrm{Re}\{Z_{i,l}(t)\exp[j2\pi f_c t + \phi_{i,l}(t)]\} \tag{14}$$

$$n(t) = \mathrm{Re}\{Z_n(t)\exp[j2\pi f_c t]\} \tag{15}$$

and $Z_s(t)$, $Z_{i,l}(t)$, $Z_n(t)$ are the complex envelope notations of user's signal, $l$th intereference and noise respectively which are complex Gaussian processes with mean zero. For simplicity, eq. (12) is rewritten as

$$r(t) = \mathrm{Re}\{Z(t)\exp[j2\pi f_c t]\} \tag{16}$$

where

$$Z(t) = Z_s(t)\exp[j\phi_s(t)] + \sum_{l=0}^{m} Z_{i,l}(t)\exp[j\phi_{i,l}(t)] + Z_n(t) \tag{17}$$

Thus, we can obtain 2-bit differentially detected signal $v(t)$ as following;

$$\begin{aligned} v(t) &= \mathrm{LPF}[\mathrm{Re}\{Z(t)\exp[j2\pi f_c t]\} \\ &\quad \times \mathrm{Re}\{Z(t-2T)\exp[j2\pi f_c(t-2T)]\}] \\ &= \frac{1}{2}\mathrm{Re}\{Z(t)\cdot Z^*(t-2T)\} \end{aligned} \tag{18}$$

where LPF[ · ] means low pass filtering. The differentially detected signal $v(t)$ is sampled at time $t=nT$, and the detector decides $u_n=1$ if $v(nT)>0$, otherwise $u_n=-1$, where $u_n$ is a information bit at time $t=nT$.

## Ⅲ. BER performance analysis

To obtain the probability of error, we require the detected signal's probability density function, which can be derived from the characteristic function $G_m(v)$ of output $v(nT)$ using inverse fourier transformation. If we assume the average value of the received signal's complex envelope $Z(t)$ is zero, the characteristic function of $v(nT)$ is given by

$$G_m(\nu) = [1 - 2j\nu\sigma_m^2(\rho_m + \rho_m^*) + 4\nu^2\sigma_m^4(1 - |\rho_m|^2)]^{-1} \tag{19}$$

where the received signal's correlation coefficient $\rho_m$ and variance $\sigma_m{}^2$ are obtained from covariance equation of $Z(nT)$ and $Z((n-2)T)$. And,

$$\sigma_m^2\rho_m = \sigma_s^2\rho_s(2T)\exp(j\Delta\phi_s) + \sum_{l=0}^{m}\sigma_{i,l}^2\rho_{i,l}(2T)\exp(j\Delta\phi_{i,l}) + \sigma_n^2\rho_n(2T) \tag{20}$$

where $\sigma_k^2$ is the average power of the received signal,

$$\sigma_m^2 = \sigma_s^2 + \sum_{l=0}^{m}\sigma_{i,l}^2 + \sigma_n^2 \tag{21}$$

In above equation, $\Delta\phi_s$, $\Delta\phi_{i,l}$ are the 2 bits duration phase differences of signal and interferences, respectively. It is assumed that the receiver has an omnidirectional antenna and the IF filter has a Gaussian shape and 3dB-bandwidth $B_i$. The baseband power spectrums of signal, interferences, and noise are given by

$$W_s(f) = \begin{cases} \dfrac{\sigma_s^2}{\pi\sqrt{f_{D,s}^2 - f^2}}, & |f| < f_D \\ 0, & |f| > f_D \end{cases} \tag{22}$$

$$W_{i,l}(f) = \begin{cases} \dfrac{\sigma_{i,l}^2}{\pi\sqrt{f_{D,i,l}^2 - f^2}}, & |f| < f_D \\ 0, & |f| > f_D \end{cases} \tag{23}$$

$$W_n(f) = \frac{\sigma_n^2}{B_i\sqrt{\frac{\pi}{\ln 2}}}\exp\left\{-\left(\frac{f}{B_i}\right)^2\ln 2\right\} \tag{24}$$

Thus, the autocorrelation coefficients of signal, interferences, and noise are given by

$$\rho_s(2T) = J_0(4\pi f_D T) = A_1 \tag{25}$$

$$\rho_{i,l}(2T) = J_0(4\pi f_D T) = A_1 \tag{26}$$

$$\rho_n(2T) = \exp\{-\frac{(2\pi B_t T)^2}{\ln 2}\} = A_2 \tag{27}$$

where $J_0(\cdot)$ is the zero$^{th}$ order Bessel function of the first kind. Defining the signal to noise ratio as $\Gamma_s = \sigma_s^2/\sigma_n^2$ and interference to noise ratio as $\Gamma_{i,l} = \sigma_{i,l}^2/\sigma_n^2$, then the correlation coefficient of the received signal is given by

$$\begin{aligned}\rho_m &= \rho_m(\Delta\phi_s, \{\Delta\phi_{i,m}\}) \\ &= \frac{1}{\Gamma_s + \sum_{l=0}^{m}\Gamma_{i,l} + 1}\{\Gamma_s A_1 \cos(\Delta\phi_s) \\ &\qquad + \sum_{l=0}^{m}\Gamma_{i,l} A_1 \cos(\Delta\phi_{i,l}) + A_2\} \\ &\quad + j\frac{1}{\Gamma_s + \sum_{l=0}^{m}\Gamma_{i,l} + 1}\{\Gamma_s A_1 \sin(\Delta\phi_s) \\ &\qquad + \sum_{l=0}^{m}\Gamma_{i,l} A_1 \sin(\Delta\phi_{i,l})\} \end{aligned} \tag{28}$$

$$\begin{aligned} &= \rho_{mr}(\Delta\phi_s, \{\Delta\phi_{i,m}\}) + j\rho_{mj}(\Delta\phi_s, \{\Delta\phi_{i,m}\}) \\ &\triangleq \rho_{mr} + j\rho_{mj}\end{aligned}$$

where $\rho_{mr}$ and $\rho_{mj}$ are the real and imaginary part of correlation coefficient $\rho_m$. Inserting $\rho_m$ into $G_m(V)$ and using inverse fourier transformation, we can obtain the conditional probability density function of $v(nT)$ expressed as

$$p(V|m \text{ hits}) = \frac{1}{2\pi}\int_{-\infty}^{\infty} G_m(\nu) e^{-j\nu V} d\nu$$

$$= \begin{cases} \frac{1}{4\sigma_m^2\sqrt{1-\rho_{mj}^2}} \exp\{\frac{\sqrt{1-\rho_{mj}^2}+\rho_{mr}}{2\sigma_m^2(1-|\rho_m|^2)} V\}, & V<0 \\ \frac{1}{4\sigma_m^2\sqrt{1-\rho_{mj}^2}} \exp\{-\frac{\sqrt{1-\rho_{mj}^2}-\rho_{mr}}{2\sigma_m^2(1-|\rho_m|^2)} V\}, & V>0 \end{cases} \tag{29}$$

Integrating the conditional pdf of $v(nT)$, we can obtain conditional bit error probability such as

$$\begin{aligned} &P_e\ (u_n = 1, m \text{ hits}) \\ &= P(V<0|u_n = 1, m \text{ hits}) \\ &= \int_{-\infty}^{0} p(V|u_n = 1, m \text{ hits}) dV \\ &= \frac{1}{2}(1 - \frac{\rho_{mr}}{\sqrt{1-\rho_{mj}^2}}) \end{aligned} \tag{30-a}$$

$$\begin{aligned} &P_e\ (u_n = -1, m \text{ hits}) \\ &= P(V>0|u_n = -1, m \text{ hits}) \\ &= \int_{0}^{\infty} P(V|u_n = -1, m \text{ hits}) dV \\ &= \frac{1}{2}(1 + \frac{\rho_{mr}}{\sqrt{1-\rho_{mj}^2}}) \end{aligned} \tag{30-b}$$

To analyze eq. (30), we must consider the phase differences $\Delta\phi$ and their probabilities. At GMSK modulator, there arise inter symbol interferences(ISI) by GLPF, and the phase difference depends on the bandwidth of premodulation Gaussian filter. In this paper, we consider that $B_tT$ is 0.25 because for lower values of bandwidth the error probability is increased, while for higher values of bandwidth the spectral efficiency is reduced. Assuming the probabilities of all user's data(±1) are equiprobable, we are able to obtain phase differences and their probability shown in table 1.

We can derive error probabilities with respect to the desired user's mark or space, which are the function of the correlation coefficients and phase differences, also the probabilities of phase differences.

Table 1. Phase differences and probability distributions.

| Phase difference | | Probability | |
|---|---|---|---|
| $\Delta_1$ | 0° | $P_{\Delta 1} = P(\Delta\phi = \Delta_1)$ | 1/4 |
| $\Delta_2$ | 37.6° | $P_{\Delta 2} = P(\Delta\phi = \Delta_2)$ | 1/4 |
| $\Delta_3$ | 103.6° | $P_{\Delta 3} = P(\Delta\phi = \Delta_3)$ | 1/8 |
| $\Delta_4$ | 141.2° | $P_{\Delta 4} = P(\Delta\phi = \Delta_4)$ | 1/4 |
| $\Delta_5$ | 178.8° | $P_{\Delta 5} = P(\Delta\phi = \Delta_5)$ | 1/8 |

$$P[\text{ Error },\ u_n=+1|\ m\ \text{hits}] \triangleq P_{e+}(m)$$
$$= \sum_{j=1}^{2} P_{\Delta j} \sum_{\substack{\forall n1, n2, n3, n4, n5 \\ n1+n2+n3+n4+n5=m}} \{P_{\Delta 1}^{n1} P_{\Delta 2}^{n2} P_{\Delta 3}^{n3} P_{\Delta 4}^{n4} P_{\Delta 5}^{n5} \times P_{e+}(m\ \text{hits}, j, n1, n2, n3, n4, n5)\} \tag{31}$$
$$= \sum_{j=1}^{2} P_{\Delta j} \sum_{\substack{\forall n1, n2, n3, n4, n5 \\ n1+n2+n3+n4+n5=m}} \{P_{\Delta 1}^{n1} P_{\Delta 2}^{n2} P_{\Delta 3}^{n3} P_{\Delta 4}^{n4} P_{\Delta 5}^{n5} \times \frac{1}{2}(1-\frac{\rho_{mr}(j, n1, n2, n3, n4, n5)}{\sqrt{1-\rho_{mj}(j, n1, n2, n3, n4, n5)^2}})\}$$

$$P[\text{ Error },\ u_n=-1|\ m\ \text{hits}] \triangleq P_{e-}(m)$$
$$= \sum_{j=3}^{5} P_{\Delta j} \sum_{\substack{\forall n1, n2, n3, n4, n5 \\ n1+n2+n3+n4+n5=m}} \{P_{\Delta 1}^{n1} P_{\Delta 2}^{n2} P_{\Delta 3}^{n3} P_{\Delta 4}^{n4} P_{\Delta 5}^{n5} \times P_{e-}(m\ \text{hits}, j, n1, n2, n3, n4, n5)\} \tag{32}$$
$$= \sum_{j=3}^{5} P_{\Delta j} \sum_{\substack{\forall n1, n2, n3, n4, n5 \\ n1+n2+n3+n4+n5=m}} P_{\Delta 1}^{n1}\{P_{\Delta 2}^{n2} P_{\Delta 3}^{n3} P_{\Delta 4}^{n4} P_{\Delta 5}^{n5} \times \frac{1}{2}(1+\frac{\rho_{mr}(j, n1, n2, n3, n4, n5)}{\sqrt{1-\rho_{mj}(j, n1, n2, n3, n4, n5)^2}})\}$$

Finally then, the error probability of fractional reuse/mixed sequence FH/CDMA system which use 2-bit differentially detected GMSK is given by

$$P_e(k) = \{P_{e+}(0)+P_{e-}(0)\}(1-P_{hit}(M, M, k)) + \sum_{m=1}^{M}\{P_{e+}(m)+P_{e-}(m)\}P_{hit}(M, m, k) \tag{33}$$

Assuming all error bits are randomized by interleaving and frequency diversity, RS coded bit error rate $P_{ec}$, is given by

$$P_{ec} = \sum_{j=t+1}^{n_c} \frac{j}{n_c}\binom{n}{j} P_e^j (1-P_e)^{n_c-j} \tag{34}$$

where $t=\lfloor(n_c - k_c)/2\rfloor$ is error correct ability and $\lfloor x \rfloor$ is maximal integer which is lower than $x$.

In a speech-oriented communication system, each user is allowed to transmit data continuously after setting up a call. A very important design issue for such a system is the user capacity. A service provider will be interested in the maximum number of users that can be served simultaneously by the system with required performance. In a digital communication system, the BER is an appropriate performance measure. In FH/CDMA system, the growing of active users increases the hit probability and the effect of cochannel interferences is considerable. And so, it increases BER and deteriorates the performance of system. We can define the user capacity as

$$C = \max_k \{\ P_e(k) < BER_{req}\} \tag{35}$$

## Ⅳ. Results

We consider seven 150 kHz wide channels to hop and this channel is shared by 10 TDMA users. A gross data rate of 200 kbps is chosen to give each user a data rate of 20 kbps. To analyze the results, we define capacity to be the number of users/cell/MHz.

Fig. 3 shows the BERs of single cell, cellular with CCI, and fractional reuse FH celluar with A=1 and A=0.5 for CIR=10 dB. Except for single cell, increasing SNR over 35 dB has no effects on improvement of performance. For the improvement, we use (8, 4) RS code and show it in Fig. 4. The

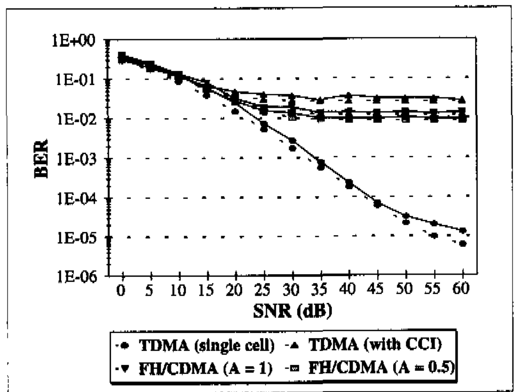



Fig. 3 BER of TDMA and FH/CDMA systems over Rayleigh Channel.

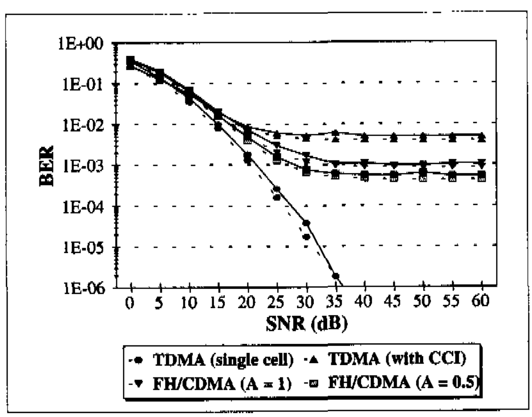



Fig. 4 BER of TDMA and FH/CDMA systems with ECC over Rayleigh Channel.

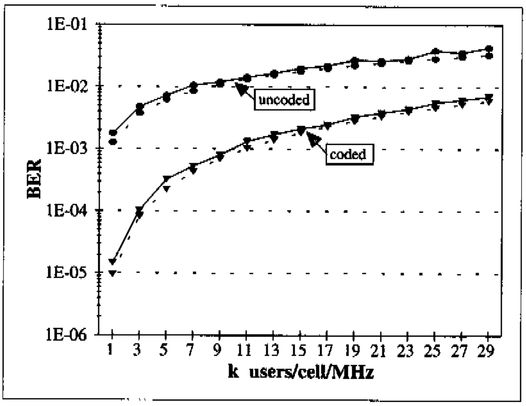



Fig. 5 System capacity of FH/CDMA systems without speech activity detector.

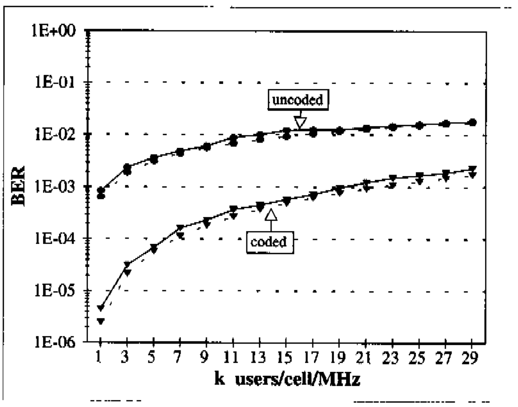



Fig. 6 System capacity of FH/CDMA systems with speech activity detector : A = 0.5.

Fig. 4 indicates that we can obtain required performance $10^{-3}$ using with frequency hopping. Fig. 5 and 6 show the user capacity of FH/CDMA with and without speech activity detector. Comparing Fig. 5 and 6 indicates that user capacity is increased about 2 times by using speech activity detector.

## V. Conclusion

In this paper we present the BER of fractional reuser FH/CDMA with 2-bit differentially detected GMSK and analyze numerically and obtain user capacity. Speech activity detector and RS coding were considered. We obtain about 20 users/cell/MHz for FH/CDMA.

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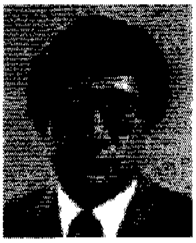

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