

# Differential Demodulation of GMSK in Rayleigh fading Channels

Joong Ho CHOI\*, Young Yeul HAN\* *Regular Members*

**ABSTRACT** A new demodulation scheme for differentially encoded GMSK signal is introduced and described. In the proposed differential detection method, the signal is sampled at the center of the time slot, at which the effect of the intersymbol interference is relatively smaller than the edge of the time slot. The proposed differential detector takes this advantage. The error rate performance of the differential detector has been numerically calculated in the fast fading encountered in the land mobile radio channels. A comparison of performance with the differential detector for MSK signal is given. Finally the possibility of improving performance employing nonredundant error correction is studied.

## I. INTRODUCTION

In mobile radio communications, modulation schemes that have constant envelope property to reduce the effect of fading and nonlinearity, and a compact power spectrum to reduce adjacent channel interference are preferred. Premodulation Gaussian filtered minimum shift keying (GMSK) has been proposed as a spectrum efficient digital modulation for mobile radio<sup>(1)</sup>. The output power spectrum of MSK can be made compact by introducing a premodulation baseband low-pass filter before modulation. Premodulation Gaussian filtering suppresses the out-of-band spurious power for GMSK, while keeping constant envelope.

Extensive studies have been carried out to improve system performance. One method takes the advantage of the fact that the degradation effect due to the premodulation Gaussian filtering is smaller at the center of time slot than at the edge of the time slot<sup>(2)</sup>. In fact, the effect of the intersymbol interference due to

the premodulation filtering is largest at the edge of time slot. Usually this point has been adopted as a sample timing in conventional coherent detection of GMSK, simply because it is the point of the maximum eye opening.

In this paper, we will take this advantage and use the center of time slot as a sampling instant in differential detection for differentially encoded GMSK signal. The error rate performance of the proposed differential detector in the fading channel has been numerically calculated. The approach used in this paper is similar to the one that calculated the error rate performance of duobinary coded MSK and TFM in(3). The possibility of improving performance for the proposed differential detector employing nonredundant error correction is investigated.

This paper begins in the next section with brief overview of GMSK signal followed in section III by a consideration of differential detection for differentially encoded GMSK signal. In section IV, we calculate the error rate performance in the fading channel. The probability of error is obtained by numerical integration for slow and fast fading. Numerical

\*漢陽大學校 電子通信工學科  
Dept. of Electronic Communication Engineering  
Han Yang University  
論文番號：91-12(接受1990. 9. 24)

results are presented in section V with concluding discussion.

## II. GMSK SIGNAL

GMSK modulation that has a constant envelope and narrow band can be achieved by placing premodulation Gaussian baseband low pass filter before MSK modulator. The impulse response of the baseband Gaussian filter is given by

$$h(t) = A \sqrt{\frac{2\pi}{\ln 2}} B \exp \left\{ -\frac{2(\pi B t)^2}{\ln 2} \right\} \quad (1)$$

where  $B$  is the 3dB bandwidth and  $A$  is constant. Its inverse Fourier transform of  $h(t)$  is given by

$$H(f) = A \exp \left\{ -\left[ \frac{f}{B} \right]^2 \frac{\ln 2}{2} \right\} \quad (2)$$

Gaussian filtered input to the modulator can be calculated as

$$g(t) = \int_{-\infty}^{\infty} h(\tau) a(t-\tau) d\tau \quad (3)$$

where  $a(t)$  is a unit rectangular pulse. The GMSK waveform of the modulator output can be expressed as

$$s(t) = \cos(2\pi f_c t + \phi(t)) \quad (4)$$

where

$$\phi(t) = 2\pi f_d \int_{-\infty}^t \sum g(\nu - nT) d\nu, \quad a_n = \pm 1 \quad (5)$$

$f_c$  is the carrier frequency and  $f_d$  is the phase deviation constant. In the case of GMSK modulation index is 0.5 and  $T$  is the bit

duration. If the excess phase of an MSK signal is plotted versus time, it follows that it is piecewise linear and increase or decrease exactly  $\pi/2$  radian each bit duration. Lines with positive slope represent 1 and lines with negative slope represent 0. A premodulation Gaussian filtering carried out in transmitter would result in degradation of system performance. The digital binary data filtered by a Gaussian low pass filter before modulation experiences severe intersymbol interference (ISI). This leads to degradation in the bit error rate.

As a specific example, let the input message sequence be 0111001... represented by  $m_0, m_1, m_2, \dots$ , and differentially encoded sequence be 11010001... represented by  $a_0, a_1, a_2, \dots$ . The resulting phase path  $\phi_k$  is shown in Figure 1. Intersymbol interference from preceding and following symbols is largest at the sampling time  $nT$  ( $n=0, 1, 2, \dots$ ). On the other hand, the effect of ISI is relatively small at the center of time slot, which is denoted by black dot. We will take this advantage in detecting GMSK signal. Figure 2 shows the phase transition of the worst case i. e., ..., 0, 0, 0, 1, 1, 1, ..., which causes the maximum effect of intersymbol interference.

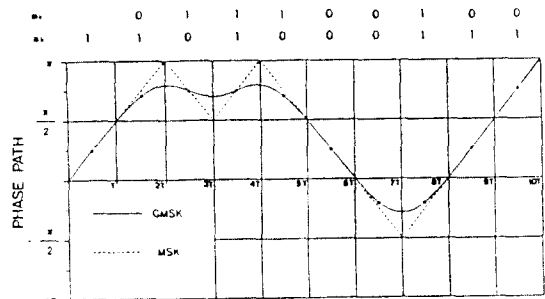


Fig. 1. Phase path in GMSK.

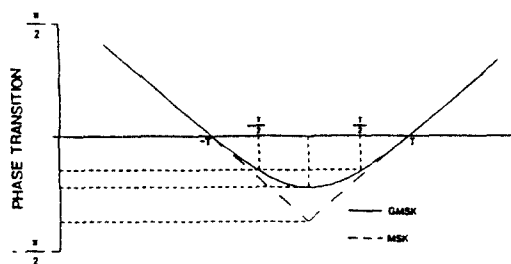


Fig. 2. Phase transition of the worst case.

### III. SYSTEM CONFIGURATION

The GMSK modulator with differential encoding is shown in Figure 3. Gaussian baseband filter suppresses well out-of-band spurious power for GMSK, while still keeping the constant envelope property. Figure 4 shows the differential detector with sampling instant  $(m+0.5)T$ , this sample timing is less sensitive to the effect of the intersymbol interference resulting from the Gaussian filter.

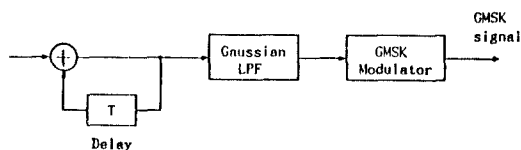


Fig. 3. GMSK modulator with differential encoding.

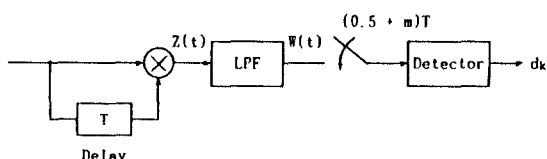


Fig. 4. GMSK demodulator with differential encoding sampled at  $t=(m+0.5)T$ .

In the absence of noise and fading, input to the low-pass filter is

$$\begin{aligned} Z(t) &= \cos[\omega_c t + \phi(t)] \cos[\omega_c(t-T) + \phi(t-T)] \\ &= \frac{1}{2} \cos[\omega_c T + \phi(t) - \phi(t-T)] \\ &\quad + \frac{1}{2} \cos[2\omega_c(t-T/2) + \phi(t) + \phi(t-T)] \end{aligned} \quad (6)$$

The double frequency terms are rejected by the low-pass filter leaving only the leading term,

$$W(t) = \frac{1}{2} \cos[\phi(t) - \phi(t-T)] \quad (7)$$

where  $\omega_c T = \text{multiple of } 2\pi$ , and the sampled output at the time instant  $(m+0.5)T$  is

$$W((m+0.5)T) = \frac{1}{2} \cos[\phi((m+0.5)T) - \phi(mT)] \quad (8)$$

The output of differential detector for differentially encoded MSK signal at the sampling instant  $(m+0.5)T$  is

$$\begin{aligned} W((m+0.5)T) &= 0 && \text{when } m_k = 0 \\ W((m+0.5)T) &= \frac{1}{2} && \text{when } m_k = 1 \end{aligned} \quad (9)$$

The decision threshold becomes 0.25. This decision threshold will be used later to calculate the probability of error for MSK signaling in the presence of fading and noise. If MSK signal shown in figure 1 is demodulated, phase of MSK signal, phase difference,  $W((m+0.5)T)$ , and output of detector are resulted as shown in Table 1.

Let  $\varphi_E(mT)$  be the phase difference between MSK signal and GMSK signal for the worst signal pattern at the sampling instant  $mT$  and  $\varphi_c((m+0.5)T)$  be the phase difference between MSK signal and GMSK signal for the worst signal pattern at the sampling instant  $(m+0.5)T$ .  $\varphi_E(mT)$  and  $\varphi_c((m+0.5)T)$  can be calculated with parameter  $BT$  as shown in

Table 1. Possible values of MSK signal in Figure 1.

Phase $\phi(t)$	$\Delta\phi(t)$	$W(m+0.5T)$	Output $d_k$
45			
135	90	0	0
135	0	1	1
135	0	1	1
135	0	1	1
45	90	0	0
15	90	0	0
15	0	1	1
15	90	0	0
135	90	0	0

Figure 5. Without loss of generality, we assumed that  $\phi(mT)$  and  $\phi((m+0.5)T)$  are  $\pi/2$  and  $\pi/4$  respectively for the case of MSK.

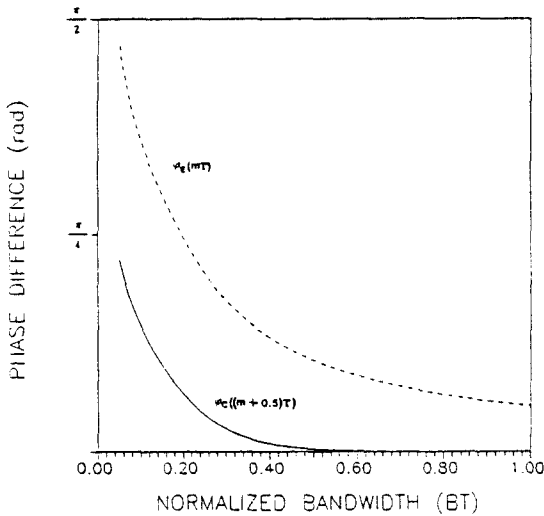
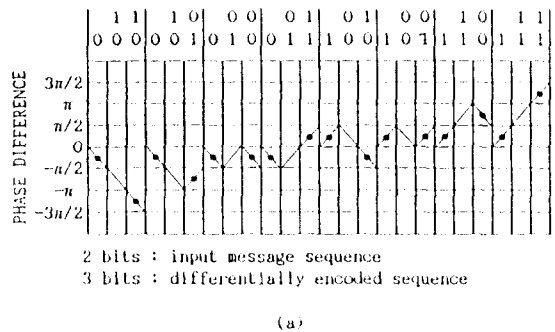


Fig. 5. Phase difference compared with MSK versus BT.

It was shown in(4) that MSK signal have the inherent properties that the outputs of a conventional differential detector and a differential detector with two time slot delay circuit sampled at  $mT$  correspond to the data and parity, respectively, of a single error correcting self orthogonal convolutional code. We will

show that the proposed demodulation scheme also has this inherent nonredundant error correcting property. Figure 6 shows the relationship between the input message sequence and output of the 2 bit differential detector sampled at  $t=(m+0.5)T$  denoted by black dot for the differentially encoded sequence. It is clear that when the output of 2 bit differential detector is  $\cos(\pm\pi)=-1$ , input message 11 is sent. When the output of the 2 bit differential detector is  $\cos 0=1$ , in input message sequence 00 is sent. Likewise when the output of the differential detector is  $\cos(\pm\pi/2)=0$ , input message sequence is 01 or 10.

We can obtain the parity of a single error correcting self orthogonal convolutional code from the output of 2 bit differential detector passed through the absolute value device and inverter ( $0 \rightarrow 1, 1 \rightarrow 0$ ).



Output of differential detector	Input bits
$\cos(\pm\pi) = -1$	1 1
$\cos(\pm\pi/2) = 0$	0 0
$\cos 0 = 1$	0 1, 1 0

(b)

Fig. 6. Relationship between the input message sequence and the output of the 2 bit differential detector sampled at  $t=(m+0.5)T$ . (a) Possible value of phase difference. (b) Output of 2 bit differential detector versus input bits.

#### IV. ERROR RATE PERFORMANCE

The received signal at the receiver input through the fast fading radio channel,  $s(t)$  is given by

$$s(t) = x_s(t)\cos[\omega_c t + \phi(t)] - y_s(t)\sin[\omega_c t + \phi(t)] \quad (10)$$

where  $x_s(t)$  and  $y_s(t)$  are independent zero mean Gaussian low pass processes. With the assumption of additive Gaussian noise at the receiver input and using Gaussian bandpass filtering before differential detection, the total signal plus noise to the differential detector is

$$e(t) = s(t) + n(t) \quad (11)$$

where

$$n(t) = x_n(t)\cos\omega_c t - y_n(t)\sin\omega_c t \quad (12)$$

and  $x_n(t)$ ,  $y_n(t)$  are independent zero mean Gaussian lowpass processes of Gaussian spectrum. Then it can be shown that the output of the sampler at the time  $t = (m+0.5)T$  is given by<sup>(5)</sup>

$$W((m+0.5)T) = \frac{1}{2} [R_1^2 - R_2^2] \quad (13)$$

where  $R_1$  and  $R_2$  are jointly Rayleigh distributed<sup>(6)(7)</sup>.

$$P(R_1, R_2) = \frac{4R_1R_2}{\sigma^2(1-|\rho|^2)} \exp\left\{-\frac{(1-\rho_1)R_1^2 + (1+\rho_1)R_2^2}{\sigma^2(1-|\rho|^2)}\right\}$$

$$\cdot I_0\left[\frac{2\rho_r R_1 R_2}{\sigma^2(1-|\rho|^2)}\right], \quad R_1, R_2 > 0 \quad (14)$$

$I_0(\cdot)$  is the modified Bessel function of the first kind and zeroth order defined by

$$I_0(\nu) = \frac{1}{2\pi} \int_0^{2\pi} \exp(\nu \cos\varphi) d\varphi \quad (15)$$

and

$$\rho = \rho_r + j\rho_i \quad (16)$$

$$\sigma^2\rho = \sigma_s^2\rho_s \exp(j(\theta_1 - \theta_2)) + \sigma_n^2\rho_n \quad (17)$$

where  $\theta_1 - \theta_2$  is the change in the desired signal phase,  $\sigma_s^2$  and  $\sigma_n^2$  are average signal and noise powers respectively, and  $\rho_s$  and  $\rho_n$  are corresponding normalized autocorrelation functions, and

$$\sigma^2 = \sigma_s^2 + \sigma_n^2 \quad (18)$$

$$\sigma_s^2\rho_s = \int_{-\infty}^{\infty} W_s(f) \exp(j2\pi fT) df \quad (19)$$

$$\sigma_n^2\rho_n = \int_{-\infty}^{\infty} W_n(f) \exp(j2\pi fT) df \quad (20)$$

where  $W_s(f)$  and  $W_n(f)$  are the equivalent baseband power spectra of  $s(t)$  and  $n(t)$  respectively. In order to analyze the error probability, let us assume that  $W_s(f)$  and  $W_n(f)$  in (19) and (20) are respectively given by

$$W_s(f) = \begin{cases} \frac{\sigma_s^2}{\pi(f_D^2 - f^2)} & , |f| \leq f_D \\ 0 & , |f| > f_D \end{cases} \quad (21)$$

$$W_n(f) = \begin{cases} \frac{\sigma_n^2}{\ln 2} \exp\left\{-\left[\frac{f}{B}\right]^2 \ln 2\right\} & , |f| \leq f_D \\ 0 & , |f| > f_D \end{cases} \quad (22)$$

where  $f_D$  is the maximum Doppler frequency. We assumed that 3dB bandwidth of IF filter and premodulation baseband filter are same.  $\rho_s$  and  $\rho_n$  can then be obtained as

$$\rho_s = J_0(2\pi f_D T) \tag{23}$$

$$\rho_n = \exp\left\{-\frac{(\pi BT)^2}{\ln 2}\right\} \tag{24}$$

where  $J_0(\cdot)$  is the zeroth order bessel function of the first kind.  $\rho$  is reduced to

$$\rho = \frac{\Gamma}{\Gamma+1} \exp\left\{-\frac{(\pi BT)^2}{\ln 2}\right\} \tag{25}$$

$$\Gamma = \frac{\sigma_s^2}{\sigma_n^2}$$

Assuming a priori probability of mark transmission in  $s(t)$  is denoted by  $p$ , the probability of error is given by

$$\begin{aligned} P_e &= p \cdot \Pr[R_1^2 - R_2^2 < a^2 | \text{mark}] + (1-p) \\ &\quad \cdot \Pr[R_1^2 - R_2^2 \geq a^2 | \text{space}] \\ &= p \int_0^{\infty} P(z | \text{mark}) dz + (1-p) \\ &\quad \int_0^{\infty} P(z | \text{space}) dz \\ &= \frac{1}{2} \int_0^{\infty} \int_{a^2-z}^{\infty} P(R_1, R_2) dR_1 dR_2 \\ &\quad + \frac{1}{2} \int_0^{\infty} \int_{-\infty}^{a^2-z} P(R_1, R_2) dR_1 dR_2 \end{aligned} \tag{26}$$

where  $z = R_1^2 - R_2^2$  and  $a^2$  is the value of threshold.

The integral in(26) is numerically calculated by integrating the density function  $P(R_1, R_2)$  over appropriate integration region in Figure 7. Assuming a priori probabilities of mark and space to be equally likely, using the numerical method, we calculated the error probability for the worst signal pattern which causes the maximum ISI effect<sup>(6)</sup>. The value of threshold  $a^2$  changes as the parameter  $BT$  changes. However the change of  $a^2$  is relatively small,

since the sampling time of  $(m+0.5)T$  is less sensitive to the effect of the intersymbol interference. The threshold value becomes 0.5 for MSK signaling. Calculated results are shown in figure 8, 9, 10 for  $BT=0.05, 0.25$  and  $0.5$  respectively. MSK results are also shown in Figure 11 for comparison. Fading rate  $f_D T$  is taken as a parameter for all plotting with values ranging from 0 to 0.005 corresponding to the vehicle speed ranging from 0 to 100 km/h at a bit rate of 16 kbps in 900 MHz band.

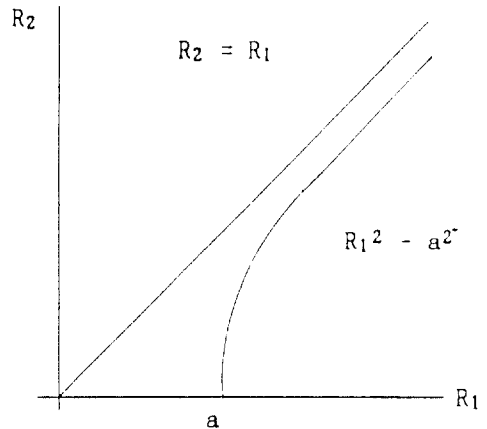


Fig. 7. Integration Region.

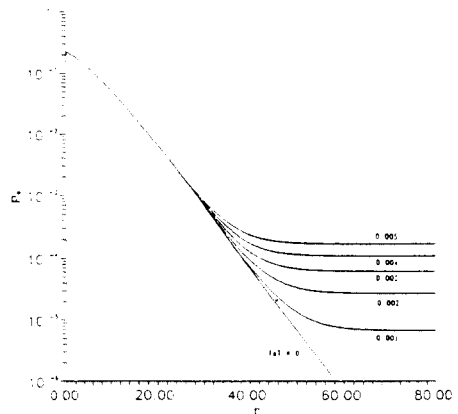


Fig. 8. Error rate performance of GMSK with differential detection with  $BT = 0.05$ .

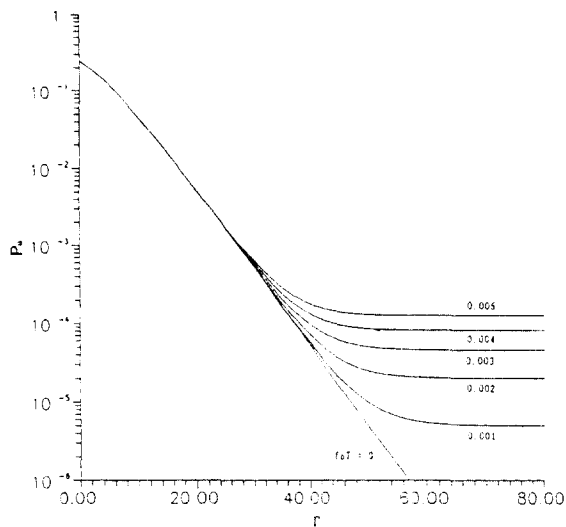


Fig. 9. Error rate performance of GMSK with differential detection with  $BT=0.25$ .

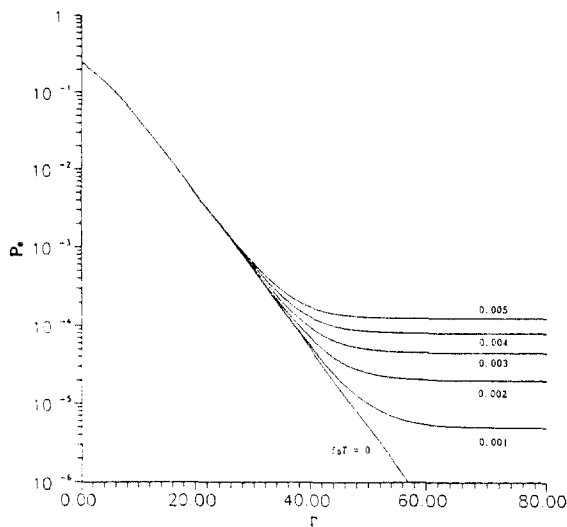


Fig. 10. Error rate performance of GMSK with differential detection with  $BT=0.5$ .

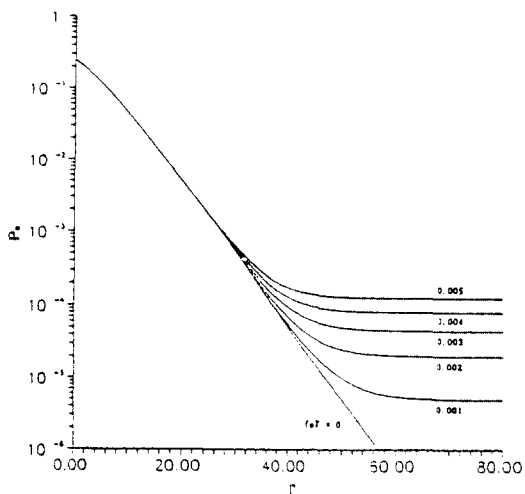


Fig. 11. Error rate performance of MSK with differential detection.

## V. CONCLUSION

An improved differential demodulation method for GMSK signal has been presented. This method take advantage of the fact that the sample timing at the center of time slot is less sensitive to the effect of the intersymbol

interference resulting from the Gaussian baseband low pass filtering prior to modulation. Error rate performance of the proposed differential demodulation of GMSK was calculated numerically in the mobile radio communication channel characterized by Rayleigh fading for the worst signal pattern. The plotted results show that the error rate of GMSK is always higher than that of MSK because of intersymbol interference. The degradations are slightly higher for the smaller values of  $BT$ , which are expected in the proposed demodulation method.

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Joongho Choi was born in Seoul, Korea on June 5, 1966. He received the B.S. degree in electronic communication engineering from Hanyang University, Seoul, Korea in 1990. Since 1990 he has been studying for the M.S. degree at Hanyang University. His research interests include digital communications and mobile communications.



韓榮烈 Young Yeul HAN 正會員

1938年6月10日生

1960年2月：서울대학교電子工學科卒業

1970年5月：美國加州大學校工學碩士

1979年5月：美國加州大學校工學博士

1961年8月～64年8月：西德Siemens社勤務

1969年8月～70年9月：KAIST勤務

1980年1月～80年9月：ADD勤務

IEEE Senior member, member of sigma Xi

現在：漢陽大學校電子通信工學科 教授