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Performance Evaluation of Various Decision Schemes for Frequency Demodulation of Narrow-Band Digital FM Signals in Land Mobile Radio

KOJI OHNO, MEMBER, IEEE, AND FUMIYUKI ADACHI, MEMBER, IEEE

Abstract—**Frequency demodulation of narrow-band digital FM signals is attractive for land mobile radio use. However, because of intersymbol interference due to premodulation filtering, the bit error rate (BER) performance is degraded severely. Various improved decision schemes, such as decision feedback equalizer decision, three-level eye decision, and maximum-likelihood sequence estimator decision, have been proposed. In this paper, laboratory experimental results are presented on the BER and block error rate (BKER) performances achieved by the above-mentioned decision schemes for 16-kb/s Gaussian-filtered minimum shift keying (GMSK) signal transmissions. Investigation of BKER performance is especially important for designing appropriate forward error correction. Field experimental results of 16-kb/s GMSK at 1.45-GHz carrier frequency are also presented to compare BER performances achieved using different decision schemes.**

I. Introduction

RECENTLY, DIGITAL SIGNAL transmission has been of growing interest in the field of land mobile radio. A narrow-band modulation scheme is necessary to utilize the limited frequency spectrum as efficiently as possible. Digital FM—such as Gaussian-filtered minimum shift keying (GMSK) [1], generalized tamed frequency modulation (GTFM) [2], and continuous-phase modulation (CPM) [3], [4]—is one of the most promising modulation schemes because of its compact power spectrum and constant envelope property (GMSK and GTFM are special cases of CPM). A narrow-band power spectrum is acheived by introducing premodulation filtering of the baseband waveform of the input to the FM modulator. Bandwidth-efficient digital FM signals can be demodulated by either coherent or noncoherent (differential and frequency) demodulators. Frequency demodulation is of significant interest, because fast multipath fading makes the use of coherent demodulation difficult, and because carrier frequency drift caused by the relatively unstable frequency oscillators used in mobile units precludes application of differential demodulation. However, increased baseband intersymbol interference (ISI) due to premodulation filtering severely degrades the bit error rate (BER) performance with frequency demodulation.

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The authors are with NTT Radio Communication Systems Laboratories, 1-2356 Take, Yokosuka-shi, Kanagawa-ken, 238-03, Japan.
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Various improved decision schemes for frequency demodulation of narrow-band digital FM signals have been proposed. A decision feedback equalizer (DFE) [5]–[7] can be used to reduce ISI. The multilevel decision scheme proposed by Hirono *et al.* [5] for GMSK signal reception is a combination of two-bit integrate-and-dump (I&D) postdetection filtering and two-bit DFE. Three-level eye decision was originally investigated by Chung [2] for differentially encoded GTFM signal reception. Using the correlative property of a three-level eye, Chung [2] also proposed an application of the maximum-likelihood sequence estimator (MLSE). The three-level eye decision and MLSE decision can also be applied to reception of GMSK signals [8]–[10] and duobinary FM signals [12].

In land mobile radio, the signal transmissions are performed within severe multipath Rayleigh fading environments. Therefore, the BER performances achieved using various decision schemes with diversity reception are very interesting. Adachi and Ohno [6] have analyzed theoretically the BER performance of GMSK with postdetection diversity combining and frequency demodulation using the one-bit DFE decision in Rayleigh fading. However, only a few reports concerning diversity combining are available on the measured BER performance of narrow-band digital FM with frequency demodulation [8], [11], [13].

In addition to diversity combining, an attractive technique to reduce fading effects is error control, such as forward error correction (FEC) and automatic repeat request (ARQ). The improvements in signal transmission performance using error control techniques depend on the block error rate (BKER) of more than M-bit errors in a block of N bits (we assume the use of block codes for FEC in this paper). Thus, it is very important to investigate the effects of the different decision schemes on BKER performance as well as BER performance.

This paper experimentally investigates performance of various decision schemes when applied to the frequency demodulation of 16-kb/s GMSK signals. Section II briefly describes the decision schemes considered in this paper. Section III compares the laboratory-measured BER and BKER performances in no-fading and simulated Rayleigh-fading environments. Two-branch diversity reception using postdetection selection combining is considered. Section IV presents the field experimental results conducted in the 1.45-GHz band.

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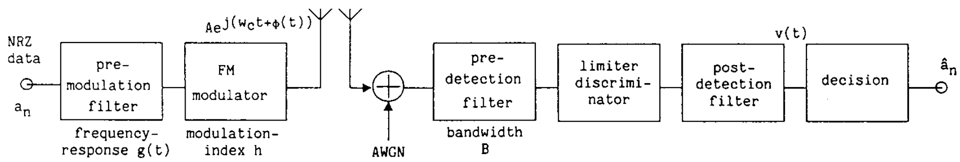

Fig. 1. Block diagram of narrow-band digital FM transmission system.

II. Improved Decision Schemes

A block diagram of the narrow-band digital FM transmission system (single branch) is shown in Fig. 1. Nonreturn-to-zero (NRZ) data a_n $(= \pm 1)$ is bandlimited by the premodulation filter before input to the FM modulator to achieve a narrow-band power spectrum. At the receiver, the predetection bandpass filter bandlimits the received signal and additive white Gaussian noise (AWGN). The limiter-discriminator (used for frequency demodulation) followed by a postdetection low-pass filter delivers the instantaneous angular frequency deviation of the bandlimited digital FM signal plus AWGN.

A. GMSK Modulation

The transmitted digital FM signals with modulation index h at the carrier angular frequency ω_c can be represented as $s(t) = \mathrm{Re}\,[A \exp j(\omega_c t + \phi(t))]$, where A is the amplitude and $\phi(t)$ is the modulating phase given by

$$\phi(t) = \pi h \sum_{n=-\infty}^{\infty} a_n \int_{-\infty}^{t} g(t-nT)\, dt \tag{1}$$

with T being the bit duration and $g(t)$ the frequency pulse response of the premodulation filter. In this paper, we concentrate on GMSK, where $h = 0.5$ and $g(t)$ is given by

$$g(t) = \frac{1}{2T}\left[\mathrm{erf}\left(\pi B_b T\sqrt{2/(\ln 2)}\left(\frac{t}{T}+0.5\right)\right) - \mathrm{erf}\left(\pi B_b T\sqrt{2/(\ln 2)}\left(\frac{t}{T}-0.5\right)\right)\right] \tag{2}$$

with B_b the 3-dB bandwidth of the premodulation Gaussian filter. Frequency pulses for other CPM schemes are tabled in [14]. The degree to which the spectrum of the modulated signal is compacted can be controlled by changing the bandwidth of the premodulation filter. A B_bT of 0.25 is typical for mobile radio applications. Particularly, the well-known minimum shift keying (MSK) signals can be generated by letting $B_bT \to \infty$, and a digital FM signal similar to tamed frequency modulation (TFM) can also be generated by using $B_bT \approx 0.2$.

B. Investigation of Eye Patterns

Signal transmission is assumed to occur over AWGN channels. Because of the narrow-band spectrum property of generated digital FM signals, the signal distortion caused by the receiver predetection filter is not severe if the filter bandwidth is not too narrow. In this case, the postdetection filter output following the limiter-discriminator can be well approximated as

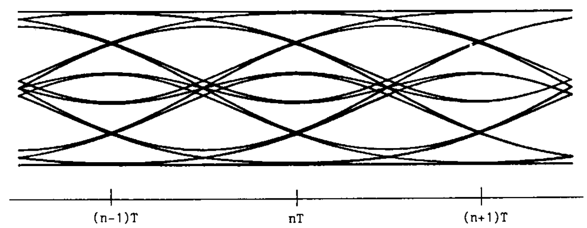

Fig. 2. Eye patterns of GMSK ($B_bT = 0.25$). No predetection filter and one-bit integrator filter as the postdetection filter are assumed.

$$v(t) = \pi h \sum_{n=-\infty}^{\infty} a_n G(t-nT) + \text{noise} \tag{3}$$

where $G(t)$ is the postdetection filter response to $g(t)$ and is defined as

$$G(t) = \int_{-\infty}^{\infty} g(t-\tau) h_B(\tau)\, d\tau \tag{4}$$

with $h_B(t)$ being the impulse response of the postdetection filter. An example of the eye patterns of a band-limited GMSK signal with $B_bT = 0.25$ and a one-bit integrator postdetection filter ($h_B(\tau) = 1/T$ for $|\tau| < T/2$ and 0 elsewhere)[1] is shown in Fig. 2. Because ISI is introduced by the premodulation filter, the binary eye aperture at $t = nT$ is reduced severely. If the ISI caused by the receiver predetection filter is taken into account, the eye aperture is reduced further.

In the following, first, we describe the DFE decision scheme that enlarges the eye aperture and so improves BER performance. Next, the three-level eye decision and MLSE decision schemes are described.

C. DFE Decision

We assume a_n to be detected without loss of generality. The sampling instant is to be taken at $t = (n + \alpha)T$, where $-0.5 \leqq \alpha \leqq 0$. From (3), the postdetection filter output v_n can be written as

$$\begin{aligned} v_n &= \pi h \sum_{m=-\infty}^{\infty} a_{n-m} G_{m+\alpha} + \text{noise} \\ &= a_n \pi h G_{+\alpha} + \pi h \sum_{m=-\infty, \neq 0}^{\infty} a_{n-m} G_{m+\alpha} + \text{noise} \end{aligned} \tag{5}$$

[1] Throughout this section, we assume a one-bit integrator filter, for convenience, whose output at time t is the integration of the limiter-discriminator output over an interval from $t - T/2$ to $t + T/2$. A one-bit I&D filter also can be used. In this case, the filter output at t is an integration of its input from $t - T$ to t. Hence, the same filter output can be obtained when the sampling instant is offset by $+T/2$ from that for the integrator filter in Sections II-C, II-D, and II-E.

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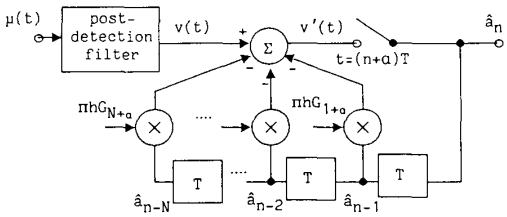

Fig. 3. Block diagram of N-bit DFE decision scheme.

where $G_{m+\alpha} = G((m + \alpha)T)$. The second term of (5) represents the ISI due to premodulation filtering. The ISI from the past bits a_{n-m} $(m = 1, 2, \cdots)$ can be canceled using the previous decision results, $\hat{a}_{n-m}$. A block diagram of the N-bit DFE decision is shown in Fig. 3. In the DFE, $\eta_{n-m} = \hat{a}_{n-m}\pi hG_{m+\alpha}$ is generated and subtracted from the limiter-discriminator output prior to the decision for a_n. If all a_{n-m} are correct, then perfect cancellation of the ISI from past N bits is possible. The DFE decision is based on the polarity of

$$v_n' = v_n - \pi h \sum_{m=1}^{N} \hat{a}_{n-m} G_{m+\alpha}. \tag{6}$$

We consider a negative offset decision instant α in order to reduce the ISI from the future bits. The ISI from the past bits may increase, but it can be canceled. The eye aperture obtained by one-bit DFE was calculated; it was found to be maximized at around $\alpha = -0.25$, which is used in the experiment. In the case of the multilevel decision proposed by Hirono *et al.* [5], a two-bit DFE with $\alpha = -0.5$ is used.

DFE produces error propagation since the present decision depends on the past decision results. This error propagation may affect FEC performance.

D. Three-Level Eye Decision

From (3), we obtain

$$v_n = (a_n + a_{n-1})\pi hG_{1/2} + \pi h \sum_{m=-\infty, \neq 0,1}^{\infty} a_{n-m} G_{m-1/2} + \text{noise} \tag{7}$$

at $t = (n - 1/2)T$. A three-level eye is obtained. Observing the eye patterns of Fig. 2 shows that this three-level eye aperture is larger than the binary eye aperture at $t = nT$, because $G_{1/2} \gg G_{-3/2}$ for $B_bT \approx 0.25$, which is typical for mobile radio applications of GMSK. To perform a decision on the three-level eye, the transmitted data must be encoded differentially. The decision rule at the receiver is that "1" is sent if $|v_n| \leq \pi hG_{1/2}$; "0" otherwise.

The eye aperture of the three-level eye decision is identical to that of the one-bit DFE decision with $\alpha = -0.5$. The DFE decision scheme can achieve larger eye aperture than that of the three-level eye decision, thus providing better BER performance. However, an advantage of the three-level eye decision is that it produces no error propagation because it is a bit-by-bit decision. This will be discussed in Section III.

E. MLSE Decision

It is well known that if there is dependence between the data sequence, MLSE can be applied to improve BER performance over a bit-by-bit decision in the Gaussian noise environment. The three-level eye v_n obtained at $t = (n - 1/2)T$ can be used, where the dependence exists only between pairs of adjacent bits, a_{n-1} and a_n, if the ISI from other bits is negligible. Chung [2] implemented the MLSE decision algorithm to use this correlative property of the three-level eye to improve the BER performance of GTFM signal reception. Chung's MLSE is based on the assumption of white Gaussian noise at the limiter-discriminator output. However, the discriminator output is characterized by non-Gaussian noise such as clicks, and, thus, this MLSE is not expected to be optimum. Nevertheless, it significantly improves BER performance. Herein, we apply this MLSE to GMSK signal reception [10]. Different from the bit-by-bit three-level eye decision, differential encoding at the transmitter is not necessary.

Three parameters, U, V, and W_n, are introduced to implement the MLSE:

$$U = W_n + v_{n+1} - A, \qquad V = W_n + v_{n+1} + A \tag{8}$$

where $A = \pi hG_{1/2}$. The decision rule is $\hat{a}_n = -1$ if $V < 0$, and $\hat{a}_n = 1$ if $U \geqq 0$. If neither $U \geqq 0$ nor $V < 0$, then a definite decision cannot be possible, and it is necessary to store two alternating bit sequences $(\pm 1, \mp 1, \pm 1 \cdots)$ in shift registers until the condition that $U \geqq 0$ or $V < 0$ happens (for detail, see [2]). In our experiments, two 16-bit output shift registers are used. W_{n+1} can be updated using the present decision result:

$$W_{n+1} = \begin{cases} v_{n+1} - A, & \text{if } U \geqq 0 (\hat{a}_n = 1) \\ v_{n+1} + A, & \text{if } V < 0 (\hat{a}_n = -1) \\ -W_n, & \text{if } U < 0 \text{ and } V \geqq 0. \end{cases} \tag{9}$$

In the MLSE, the present decision is based on the value of W_n determined by the previous decision; therefore, similar error propagation is expected, as in the DFE decision scheme.

III. Measured BER and BKER Performance

A. Experimental Setup

The laboratory experiment block diagram is shown in Fig. 4. A 16-kb/s $(2^9 - 1)$-bit pseudonoise (PN) sequence was generated for the transmitted data. It was band-limited by a premodulation Gaussian filter having a 3-dB bandwidth of 4 kHz $(B_bT = 0.25)$ and was fed into the 920-MHz FM modulator to generate a GMSK signal.

A Rayleigh fading simulator was used to generate two independent Rayleigh fading signals for two-branch postdetection diversity reception. The simulator outputs were fed into a two-branch diversity receiver using postdetection selection combining (SC), in which each of the two GMSK signals was band-limited at the 455-kHz IF stage by a predetection filter (Gaussian-type bandpass characteristics) with a 3-dB bandwidth $B = 17$ kHz $(BT = 1.1)$, before application to the limiter-discriminator. The discriminator output with the larger received signal envelope was selected as the receiver output

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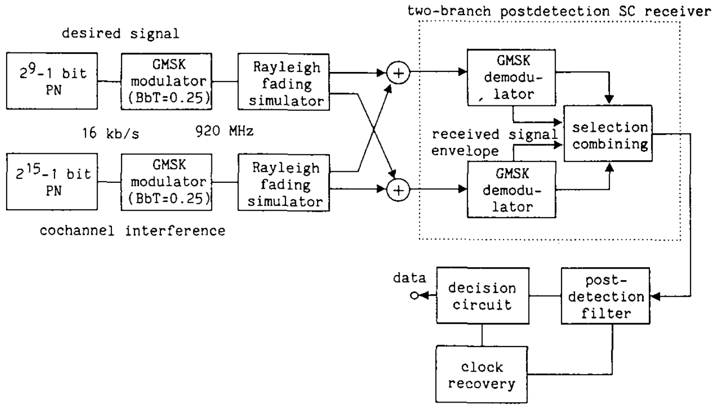

Fig. 4. Laboratory experiment block diagram.

(postdetection SC). The receiver output was fed into a postdetection filter, followed by the various decision circuits described in Section II. In the experiment, the one-bit I&D postdetection filter was used.

In cellular mobile radio systems, to realize efficient frequency utilization, the same carrier frequency is reused in spatially separated cells. The reuse distance is determined from the signal-to-interference ratio (SIR) necessary to obtain a system-determined BER. Hence, the BER performance in cochannel interference-limited channels is as important as in the AWGN channels. Cochannel interference was identical to the desired signal for a GMSK signal with a $B_b T$ of 0.25, but was generated using a $(2^{15} - 1)$-bit PN sequence.

B. BER Performance

First, the measured BER performances in a no fading environment are presented and discussed. Next, the performances in a Rayleigh fading environment are discussed.

The decision algorithms described in Section II were performed in software on a computer to allow a valid comparison of their BER performance. In this case, the postdetection filter output was sampled and stored in memory. For the two-bit DFE using a two-bit I&D filter, two consecutive one-bit I&D filter outputs were added to form a two-bit I&D filter output. For one-bit DFE, an offset decision instant $\alpha = -0.25$ was used since the BER was minimized, as expected from Section II. The measured BER's in no fading environments are plotted in Figs. 5 and 6 for AWGN and cochannel interference-limited channels, respectively. In the figures, the BER performance of MSK is also plotted for comparison. The results for MSK transmission agree well with the experimental results of Tjhung and Wittke [15].

It can be seen that two-bit DFE with two-bit I&D filtering achieves the best BER performance, and a BER of 10^{-3} can be obtained at a signal energy bit-to-noise power spectrum density ratio (E_b/N_0) of 11.7 dB, with a loss of 2 dB in E_b/N_0 compared with MSK signal reception. The BER performance of the MLSE decision scheme is slightly inferior to that of two-bit DFE, and the E_b/N_0 necessary for a BER of 10^{-3} is

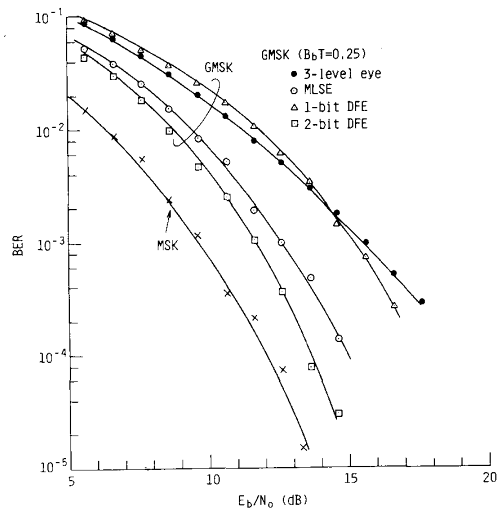

Fig. 5. Measured BER performance in AWGN channel. No fading.

12.6 dB.[2] It is seen that a 3-dB improvement in E_b/N_0 over the three-level eye decision scheme is obtained using MLSE. This improvement is consistent with the result of GTFM signal reception [2]. For cochannel interference-limited channels, the best performance was obtained with MLSE. The SIR necessary for BER $= 10^{-3}$ is 11.8 dB, which is a loss of 3.6 dB compared with MSK signal reception. Two-bit DFE is 2 dB inferior to MLSE in cochannel interference situations.

The results for Rayleigh fading and postdetection diversity reception are plotted in Figs. 7 and 8. A fading maximum Doppler frequency f_D of 40 Hz was used (this corresponds to a mobile speed of 47 km/h at a 920-MHz carrier frequency). In the figures, measured BER's of MSK are also plotted for comparison. For AWGN channels, the MLSE and two-bit DFE decision schemes have quite similar performance and achieve a BER of 10^{-3} at an average E_b/N_0 of 18.5 dB with

[2] In the case of GTFM signal reception ($B = 0.62$ and $r = 0.36$) [2, fig. 21], the measured signal-to-noise ratio necessary for a BER of 10^{-3} is about 12.1 dB. This value is close to the measured value for the GMSK case ($B_b T = 0.25$).

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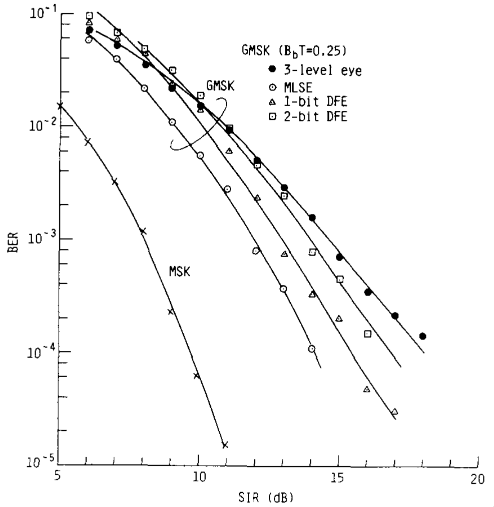

Fig. 6. Measured BER performance in cochannel interference-limited channel. No fading.

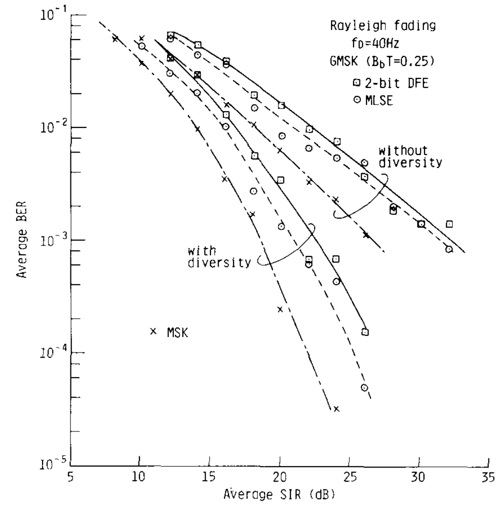

Fig. 8. Measured average BER performance in cochannel interference-limited channel. Rayleigh fading.

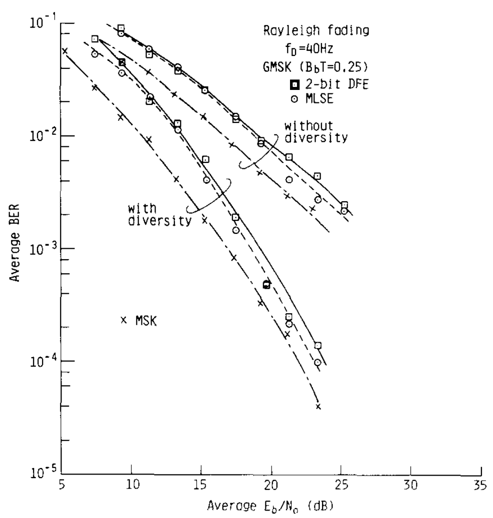

Fig. 7. Measured average BER performance in AWGN channel. Rayleigh fading.

diversity reception, which is about a 1.5-dB loss in the average E_b/N_0 compared with MSK signal reception. For cochannel interference-limited channels however, MLSE achieves better BER performance, and a BER of 10^{-3} was obtained at an average SIR of 21 dB with diversity reception, which is a 2.5-dB loss compared with MSK signal reception.

The figures also show that postdetection diversity reception is effective to reduce the impact of multipath fading and, hence, can reduce the transmitter power and cochannel reuse distance in a narrow-band digital FM mobile radio.

It can be concluded from Figs. 5–8 that MLSE could be the most efficient decision scheme from the BER performance point of view. However, its algorithm is more complicated than the DFE decision scheme, which can be implemented easily in hardware and, thus, has a practical advantage over MLSE. Although there are some differences in the BER performances of the various decision schemes, the differences are much less in Rayleigh fading channels (compare Figs. 5 and 7). This is because, in these channels, most bit errors are produced at deep fades, where the differences are naturally small.

C. BKER Performance

The improvements in BER performance with the use of FEC depends on the BKER $P(M, N)$ of more than M bit errors in a block of N bits. In the experiments, we used values of N less than 48 bits. The reason is that in existing mobile radio systems, the random error-correcting FEC codes, such as the Bose–Chaudhuri–Hocquenguem codes, of relatively short bit lengths (~ 48 bits long) are widely used.

Measured BKER values are plotted against M for various values of N for the MLSE and three-level eye decision schemes in Fig. 9(a) and for the DFE decision scheme (1-bit and 2-bit) in Fig. 9(b). Although the BKER's of the three-level eye decision scheme decrease rapidly, those of the MLSE and DFE decision schemes decrease more slowly due to error propagation effects. The BER of the three-level eye decision at $E_b/N_0 = 10.6$ dB was 1.3×10^{-2}. The $P(M, N)$ values calculated from the binomial distribution are plotted for $N = 16$ bits in Fig. 9(a). Quite good agreement is observed. Since three-level eye decision is a bit-by-bit decision, the bit errors can be assumed to be independent. Conversely, it can be seen from Fig. 9 that most of the errors produced by the DFE and MLSE decision schemes are double-bit errors. To show this more clearly, Fig. 10 compares the probability of any bit error falling into n-bit solid (or consecutive) burst error for the three-level eye and two-bit DFE decision schemes at $E_b/N_0 = 10.6$ dB. The probability is equal to the ratio of the number of bit errors contained in n-bit solid errors to the total number of bit errors. Double-bit errors are predominant in the DFE decision scheme, whereas most bit errors are single errors in the three-level eye decision scheme.

Measured BKER performances in an AWGN channel with

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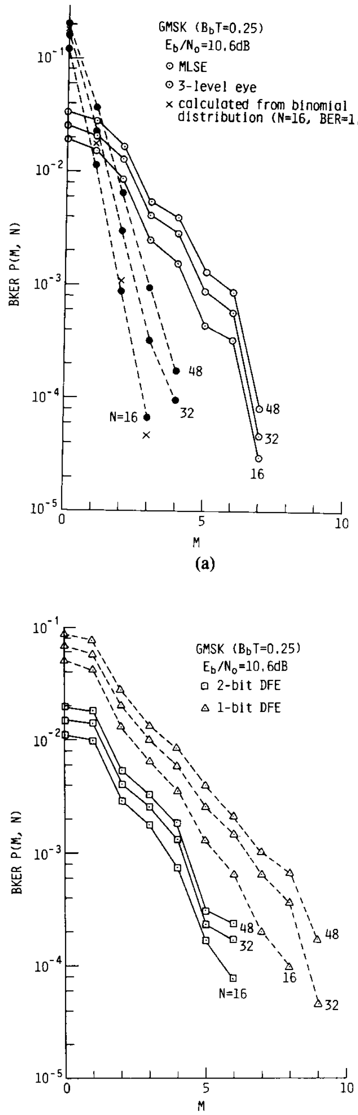

Fig. 9. Measured BKER performance in AWGN channel. No fading. (a) MLSE and three-level eye decision. (b) DFE.

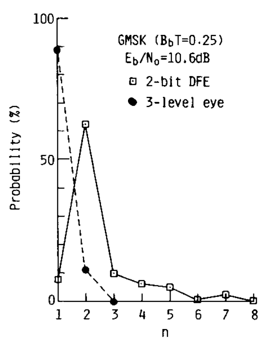

Fig. 10. Probability of any bit error falling into n-bit solid burst error.

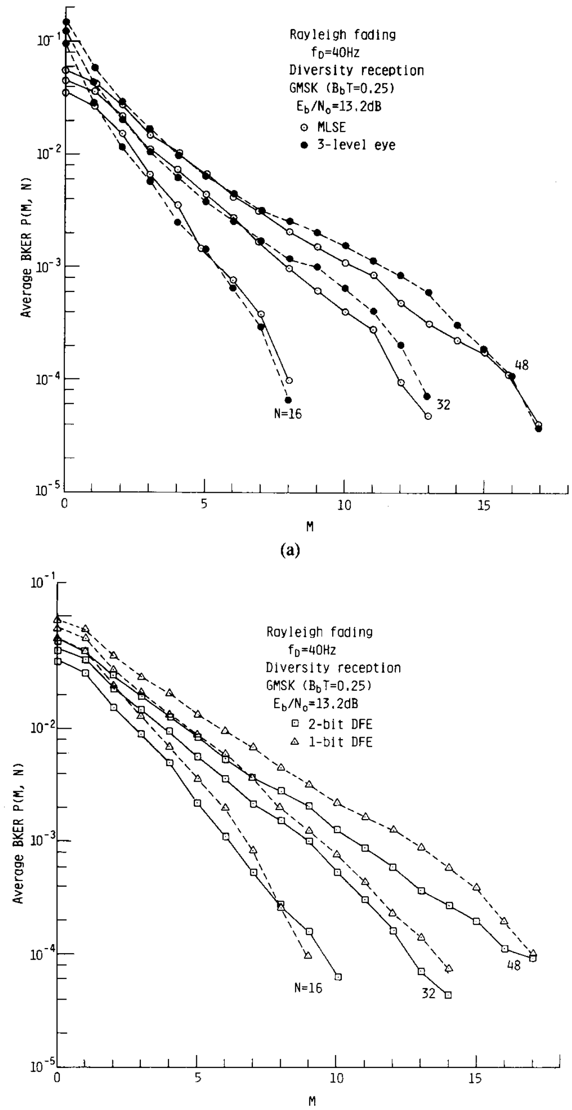

Fig. 11. Measured BKER performance in AWGN channel. Rayleigh fading. (a) MLSE and three-level eye decision. (b) DFE.

Rayleigh fading are shown in Fig. 11 for an average $E_b/N_0 =$ 13.2 dB and a maximum Doppler frequency f_D of 40 Hz. Diversity reception was used. Similar BKER curves were obtained (except for $M = 0$) for all decision schemes, and the BKER values decrease very slowly with M for all decision schemes. The reason for this is attributed to the fact that in Rayleigh fading, burst errors produced by deep fade predominate over error propagation effects. Similar results were obtained from cochannel interference-limited channels.

Two important findings have been obtained. First, if a simple ARQ incorporating error detection is employed, the throughput is determined by the value of $P(0, N)$. The DFE or MLSE decision scheme can achieve slightly better throughput than the three-level eye decision, because errors tend to be caused in bursts (most of them are double-bit errors) due to

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error propagation. Second, if an FEC technique correcting up to M (≥ 2) bit errors is employed, the three-level eye decision scheme provides the best performance in a no fading environment. However, it should be noted that if a bit interleaving technique to randomize bursty errors is used, the DFE and MLSE decision schemes provide better performance than the three-level eye decision scheme, even in a system using an FEC technique.

IV. Field Experiments

A. Experimental Setup

Fig. 12 shows the experiment setup of the base and mobile stations. The transmitting site was located on top of the NTT Research & Development Center, Yokosuka, Japan. A 16-kb/s ($2^2 - 1$)-bit PN sequence was used as the data to be transmitted. A GMSK signal with $B_bT = 0.25$ generated at 1.45-GHz-band carrier frequency was amplified by a 20-W class-C power amplifier. The transmitting antenna was a corner-reflector type with a 10-dB gain over the dipole. The test area was residential, located approximately 1.3km from the base station. A measuring vehicle was driven at a speed of around 23 km/h. Two $\lambda/4$ whip antennas separated by about 1.3 carrier lengths were mounted on top of the vehicle for space diversity reception. The separation used was large enough to achieve low fading correlations. The antenna outputs were fed to the postdetection SC receiver, described in Section III. A signal strength of 0 dB-μV (– 113 dBm at 50 Ω) corresponds to E_b/N_0 = 14.0 dB. The three-level eye decision and the one- and two-bit DFE decision schemes were adopted for the experiments. For the one-bit DFE and three-level eye decision schemes, the postdetection filter used was a Gaussian low-pass filter with a 3-dB bandwith, B_0, of 7 kHz instead of the one-bit I&D filter. however, two-bit DFE was followed by a two-bit I&D postdetection filter. The average BER was determined by counting the number of bit errors produced during a 5-s interval and dividing by 8×10^4. The measured average BER's and median signal strengths over a 5-s interval, and the vehicle speed, were stored on a magnetic disk for later processing. The 5-s interval corresponds to about 150 wavelengths at a vehicle speed of 23 km/h, which is a long enough period to measure the median signal strength.

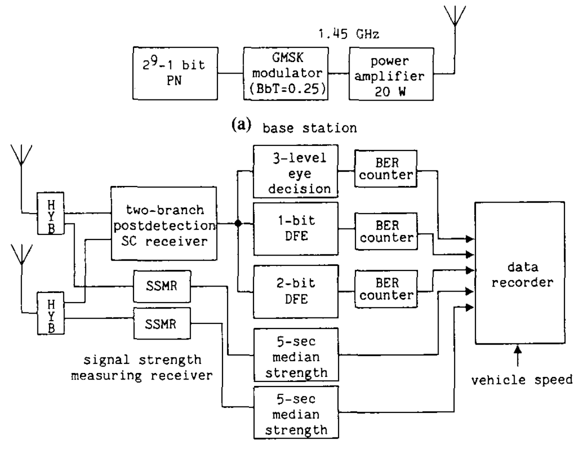

Fig. 12. Field experiment setup.

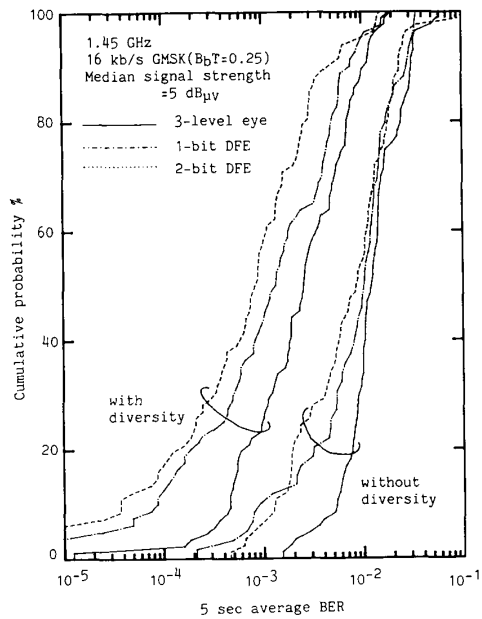

Fig. 13. Measured cumulative probability distribution of 5-s average BER's at median signal strength of 5 dB-μV.

B. Results

The measured average BER's scattered possibly because of variations in the fast fading characteristics with different vehicle locations. Fig. 13 shows an example of the cumulative probability distribution of measured average BER, calculated from 5-s average BER's obtained at median signal strengths of within 5 dB-μV $\pm$ 1 dB. The two-bit DFE decision scheme offers the best performance. With diversity reception, the 50 percent BER (below which are 50 percent of measured average BER's) is 8.4×10^{-4} for two-bit DFE, 1.4×10^{-3} for one-bit DFE, and 2.6×10^{-3} for the three-level eye decision.

To evaluate BER performance of the DFE and three-level eye decision schemes, the values of the 50 percent BER are plotted in Fig. 14 with 2-dB increments in the median signal strength. It shows that the two DFE decision schemes provide similar BER performance, but two-bit DFE is slightly superior. At a 50 percent BER of 10^{-2}, about 6-dB diversity improvement was obtained, which agrees with laboratory experiments.

V. Conclusion

This paper has evaluated the decision schemes of the decision feedback equalizer (DFE), the three-level eye, and the maximum-likelihood sequence estimator (MLSE) for frequency demodulation of 16-kb/s GMSK signals. First, laboratory experimental results on bit error rate (BER) and block error rate (BKER) performances were presented. In additive white Gaussian noise channels, two-bit DFE can achieve the best BER performance, whereas MLSE is the best for

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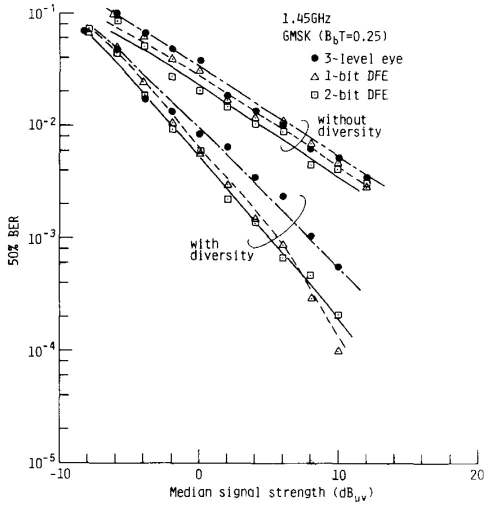

Fig. 14. Measured 50% BER performance of DFE and three-level eye decision schemes.

cochannel interference-limited channels. BKER performance was also examined. The three-level eye decision is a bit-by-bit decision and, thus, has superior performance because there is no error propagation. In fading environments, however, this superiority tends to diminish because bursty errors due to deep fades predominate rather than error propagation effects. Some of the laboratory experimental results have been confirmed with the field experiments at a 1.45-GHz carrier frequency. The results presented in this paper may be useful for designing a suitable error control technique and choice of appropriate decision scheme for mobile radio applications.

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Koji Ohno (M'83) was born in Gifu, Japan, in 1958. He received the B.S. and M.S. degrees from Nagoya Institute of Technology, Nagoya, Japan, in 1981 and 1983, respectively.

In 1983 he joined NTT Laboratories. He has been engaged in the research and development of digital mobile radio communication systems. He is now a Research Engineer, NTT Radio Communication Systems Laboratories.

Mr. Ohno is a member of the Institute of Electronics and Communication Engineers of Japan.

Fumiyuki Adachi (M'79) graduated from Tohoku University, Japan, in 1973, and received the Ph.D. degree in 1984.

Since 1973 he has been with the Nippon Telegraph & Telephone Corporation (NTT) Laboratories in Japan, where he has carried out various analytical and experimental studies concerning diversity techniques in fading digital FM. He has also been concerned with the studies of man-made radio noise, voice transmission quality, syllabic compander effects, statistical estimation of the signal strength in Rayleigh fading. Currently, he is responsible for the research on digital transmission techniques including modulation/demodulation, diversity reception and error control, and on random access protocols for cellular digital land mobile radio. During the academic year of 1984/1985, he was U.K. SERC Visiting Research Fellow at the Department of Electrical Engineering and Electronics of Liverpool University, U.K.

Dr. Adachi was awarded the IEEE Vehicular Technology Society 1980 Paper of the Year Award. He is a member of Institute of Electronics, Information and Communication Engineers of Japan.

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