Differentially Detected GMSK Signals in CCI Channels for Mobile Cellular Telecommunication Systems

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Abstract—**The performance of conventional and decision feedback differential detection receivers for GMSK signals transmitted in the presence of co-channel interference (CCI) and additive white Gaussian noise (AWGN) is evaluated. For the interference, we adopt a model which includes N statistically independent static as well as faded CCI. Various bit-error-rate (BER) performance evaluation results have indicated that the receiver under investigation performs better as compared to other more conventional receiver structures. Especially significant BER improvements have been obtained for the static CCI channel. For example, it was found that with a carrier-to-interference (C/I) ratio of 14 dB, the performance of a 2-bit decision feedback differential receiver outperforms a conventional 2-bit differential detector by more than 14 dB (at a BER = 10^{-3}). For the faded CCI, the improvement is less: mainly they result in error-floor reductions of about half an order of magnitude. By comparing the performance of static and faded CCI, it was also found that for a given C/I, the performance of the former would depend on the number of interferers, whereas this is not the case for the latter.**

I. Introduction

AMONG the great deal of correlative encoded continuous phase modualtion (CECPM) schemes [1] which have been investigated in the past, Gaussian minimum shift keying (GMSK) [2] has been perhaps one of the most extensively studied. Because of its excellent spectral properties and simple implementation structure, GMSK has been a very popular modulation scheme for mobile radio telecommunication applications. More importantly, GMSK has been selected recently as the new transmission standard for the GSM pan-European digital cellular system [3].

Earlier publications on GMSK include performance analysis and evaluation in flat fading channels (e.g., [4]–[10]) and more recently for frequency-selective fading and co-channel interference (CCI) [11], [12]. In the latter environment, it was found that discriminator detection of GMSK performs better than its counterpart, differential detection [12]. It should be pointed out, however, that the operation of the discriminator detector suffers from so-called "FM-clicks" [13] which tend to occur quite often in the presence of strong interference. In [14] a differential detection technique based on decision feedback, which improves the performance of differentially detected GMSK schemes, has been proposed and evaluated for the AWGN channel. A generalization of this technique to include any type of CECPM scheme is proposed and analyzed in [15]. Furthermore, the performance of this decision feedback technique for GMSK signals operated in the presence of CCI has not been investigated yet. Such an environment has recently become very important because of the tremendous interest and application of frequency reuse wireless telecommunication systems, including land-mobile cellular and indoor telecommunications (see e.g., [16]–[18]). Thus the subject of this paper is the evaluation of these receiver structures for GMSK schemes operated in a combined CCI and AWGN environment.

The outline of the paper is as follows. In Section II we present a detailed system model description which includes the transmitter and channel model, as well as receiver structures. Section III gives the various numerical results in terms of bit-error-rate (BER) performance under various channel conditions. Also, discussion and interpretation of the obtained results are included. Conclusions are drawn in Section IV.

II. System Model Description

A. Transmitter

The GMSK transmitter consists of a differential encoder, a Gaussian premodulation filter, and an FM modulator [4]. The differential encoder is required to prevent the error propagation for 2-bit or higher differential detection [14]. The information bits, a_k, taken from the alphabet $\{\pm 1\}$, are equally probable and independent, and the output of the differential encoder b_k can be expressed as

$$b_k = a_k \prod_{i=1}^{l-1} b_{k-i} \tag{1}$$

for the receiver structure using l-bit ($l \geq 2$) differential detection.

After the Gaussian premodulation filter $H_t(f)$ and the FM modulator, the GMSK signal $s(t)$ is given by

$$s(t) = A_0 \cos[2\pi f_c t + \phi(t)] \tag{2}$$

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where A_0 is a constant amplitude and $\phi(t)$ is the phase containing the information given as [1]:

$$\phi(t) = \frac{\pi}{2T} \sum_{i=-\infty}^{\infty} b_i \int_{-\infty}^{t} g(\tau - iT)\, d\tau. \tag{3}$$

In (3), T is the bit duration and $g(t)$ is the filter response to a rectangular pulse of duration T, which can be expressed as [4]

$$g(t) = Q\left[kB_tT\left(-\frac{1}{2} - \frac{t}{T}\right)\right] - Q\left[kB_tT\left(\frac{1}{2} - \frac{t}{T}\right)\right] \tag{4}$$

where

$$Q(x) = \frac{1}{\sqrt{2\pi}} \int_x^{\infty} \exp\left(-\frac{y^2}{2}\right) dy \tag{5}$$

with $k = 7.546$ and B_t being a single-sided 3-dB bandwidth of the Gaussian filter. In this paper, we will focus on $B_tT = 0.25$, since this value satisfies the requirement that the out-of-band radiation be below 60–80 dB in mobile radio communication [2].

B. Channel Model

As illustrated in Fig. 1, the channel model consists of N independent co-channel interferers $i_1(t), i_2(t), \cdots, i_N(t)$, which could be Rayleigh faded so that they become $i_1^f(t), i_2^f(t), \cdots i_N^f(t)$, respectively. Similarly, the transmitted signal $s(t)$ could be also Rayleigh faded, thus resulting in $s^f(t)$. Clearly when fading is absent $i_j^f(t) = i_j(t)$ for $1 \le j \le N$ and $s^f(t) = s(t)$. Furthermore, because of the additive nature of the CCI,

$$i(t) = \sum_{j=1}^{N} i_j(t) \quad \text{and} \quad i^f(t) = \sum_{j=1}^{N} i_j^f(t).$$

The Rayleigh fading is assumed to be a filtered complex Gaussian process $f(t)$, which has an autocorrelation function:

$$R_f(\tau) = J_0(2\pi f_D \tau) \tag{6}$$

where $J_0(\cdot)$ is a zeroth-order Bessel function and f_D is the maximum Doppler frequency. We will consider the case of flat fading so that $s^f(t) = s(t)f(t)$ and $i_j^f(t) = i_j(t)f(t)$ for $1 \le j \le N$. For the CCI, it will be assumed that they employ the same modulation format, i.e., GMSK with $B_tT = 0.25$, and the same bit duration T. Mathematically, the jth CCI can be represented as

$$i_j(t) = B_j \cdot \cos\left[2\pi f_c t + \alpha_j + \beta_j(t)\right] \tag{7}$$

with B_j being a constant amplitude, α_j a carrier random phase shift which is uniformly distributed over $[0, 2\pi)$, and $\beta_j(t)$ given by

$$\beta_j(t) = \frac{\pi}{2T} \sum_{i=-\infty}^{\infty} a_{j_i} \int_{-\infty}^{t} \cdot g(\gamma - iT - \tau_j)\, d\gamma \tag{8}$$

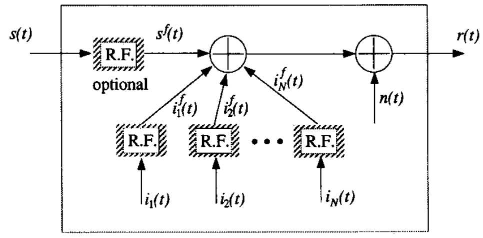

Fig. 1. Block diagram of the channel model.

where a_{j_i} are independent and equally probable bits which are not necessarily differentially encoded, and τ_j is a random time delay uniformly distributed over the time interval $[0, T)$. For the CCI, we will consider the worst-case scenario where, for a given interference power, the amplitude of each interferer is the same, i.e., $B_j = B$ for $1 \le j \le N$. Therefore, in this case, the C/I is given as

$$C/I = \frac{A_0^2}{NB^2}. \tag{9}$$

This definition is extended to the case of Rayleigh fading, although, as will be shown later, the BER performance will not depend on the number of interferers for the Rayleigh faded CCI because of its Gaussian characteristics. In addition, as illustrated in Fig, 1, the composite signal is corrupted by AWGN, $n(t)$, which is a zero-mean Gaussian random process with a double-sided power spectral density $N_0/2$. Clearly, now the received signal $r(t)$ consists of three components:

$$r(t) = \begin{cases} s(t) + i(t) + n(t) \\ \quad \text{(without fading)} \\ s^f(t) + i^f(t) + n(t) \\ \quad \text{(with fading)}. \end{cases} \tag{10}$$

C. Receivers with Decision Feedback [14]

It is well known that, for filtered GMSK signals, conventional differential detectors[1] do not have good performance (see for example [4]). This fact is more prominent for severely filtered GMSK signals (e.g., for $B_tT = 0.25$) because for such signals the transmitted pulse extends over the duration of several symbols, thus creating intersymbol interference (ISI).

An efficient receiver that improves the performance of a differentially detected GMSK scheme was proposed in [14] and generalized in [15]. Its structure is based upon a decision feedback (DF) method which is easy to implement and very effective to combat a great amount of inherent ISI between successive symbols [14]. This technique partially removes ISI using the phase angle feedback loop with the aid of the previously decoded data bits. Fig. 2 illustrates the general block diagram of this receiver structure applied to the conventional l-bit differential detector. We have chosen the predetection receive filter $H_r(f)$ to be the 4th-order band-pass Butterworth filter. Here, we define the signal-to-noise

[1] With conventional differential detectors, we refer to differential detectors with an l-bit ($l = 1, 2, 3, \cdots$) delay element.

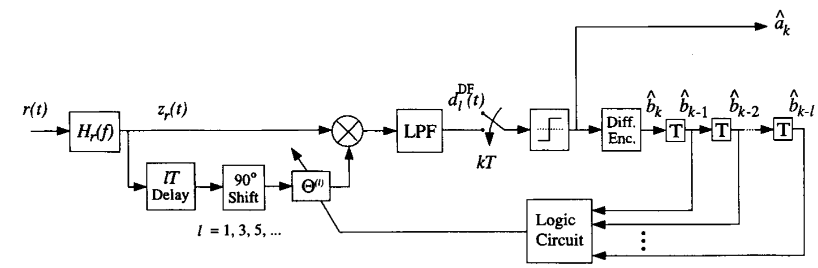

Fig. 2. Block diagram of the l-bit differential detection receiver employing the decision feedback (DF) technique.

ratio (S/N) as

$$S/N = \frac{A_0^2}{2\sigma_n^2} = \frac{E_b}{N_0}\frac{a^2(t)}{B_{rn}T} \tag{11}$$

where E_b is the transmitted bit energy, B_{rn} is the noise bandwidth, σ_n^2 is the noise power, given as

$$\sigma_n^2 = N_0 \int_{-\infty}^{\infty} |H_r(f)|^2\, df \tag{12}$$

and $a(t)$ is the normalized signal amplitude after the signal is filtered [4], i.e.,

$$a(t) = \sqrt{[h_r(t) * \cos(\phi(t))]^2 + [h_r(t) * \sin(\phi(t))]^2}. \tag{13}$$

In (13), $*$ denotes a convolution, and $h_r(t)$ is the inverse Fourier transform of $H_r(f)$. For an nth-order Butterworth filter [19],

$$B_{rn} = \frac{\pi/2n}{\sin(\pi/2n)} B_r \tag{14}$$

where B_r is a double-sided 3-dB bandwidth. Thus for $n = 4$, (11) becomes

$$S/N = \frac{E_b}{N_0}\frac{a^2(t)}{cB_rT} \tag{15}$$

with

$$c = \frac{\pi/8}{\sin(\pi/8)} \approx 1.026. \tag{16}$$

Expressions for the phase shift $\Theta^{(l)}(l = 1, 2)$ as well as the decision variable $d_l^{DF}(t)$ (see Fig. 2) can be found in [14].[2]

III. Numerical Results and Discussion

Using a fourth-order Butterworth as the receiver filter, $H_r(f)$, we have chosen B_rT equal to 1.0 and 0.86, respectively, for all related receivers employing 1- and 2-bit differential detectors. It was found from computer simulation that these values are nearly optimum at BER = 10^{-3} in an AWGN channel [20].

To demonstrate the effects of the number N of CCI on the BER performance, Fig. 3 shows the BER curve of conventional 1-bit differentially detected GMSK with one $(N = 1)$ and four $(N = 4)$ interferers operated in both environments: a static CCI-AWGN at C/I = 18 dB and a Rayleigh faded CCI-AWGN at C/I = 30 dB. For the static case, the BER performance degrades noticeably as N increases, whereas for the fading case, in principle it is not affected by it. This result verifies the well-known fact that the characteristics of a static CCI cannot be considered Gaussian [21]. However, a faded CCI has Gaussian characteristics independent of the number of interferers; therefore, its BER performance depends on the value of C/I and not on N.

[2]Note that as the order n increases, the difference between the noise bandwidth and the 3-dB bandwidth decreases.

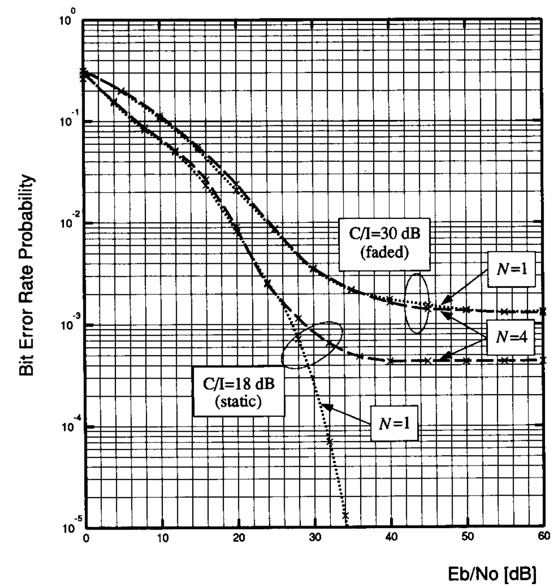

Fig. 3. BER performance of conventional 1-bit differentially detected GMSK scheme operated in the presence of static and Rayleigh faded CCI with $N = 1$ and $N = 4$.

Fig. 4 illustrates the obtained BER performances of the proposed receivers (denoted as "DF") for a static CCI at C/I = 14 dB ($N = 1$ and $N = 4$). For comparison purposes, in the same figure, the performance of conventional 1- and 2-bit differential receivers (denoted as "C") operated in the same environment are also given. From these results, it is clear that the proposed receivers result in significant BER improvements. For example, the 2-bit DF receiver results in gains of more than 14 dB as compared to the performance of a conventional 2-bit

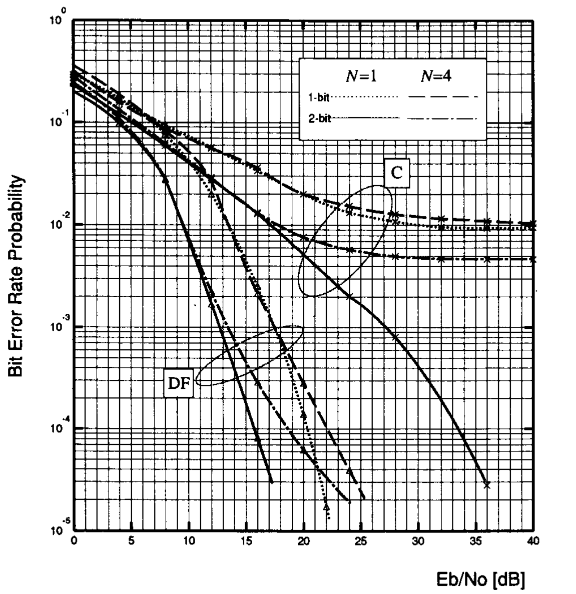

Fig. 4. BER performance of the conventional (C) and decision feedback (DF) 1- and 2-bit differential detection receiver in the presence of static CCI (C/I = 14 dB) with $N = 1$ and $N = 4$.

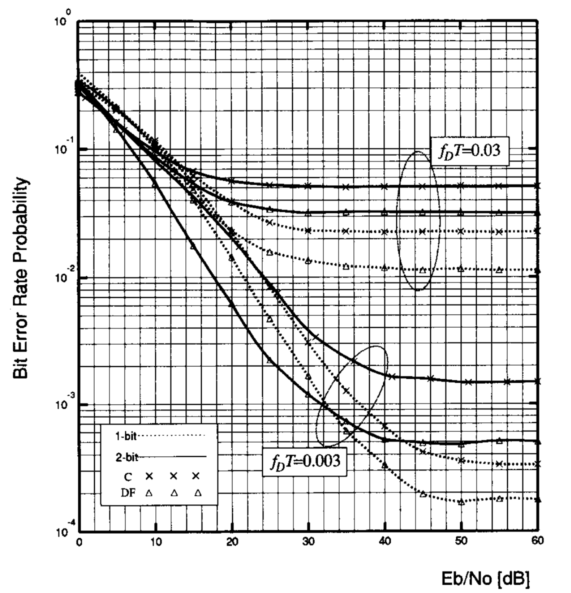

Fig. 6. BER performance of the conventional (C) and decision-feedback (DF) 1- and 2-bit differential detection receivers in the presence of faded CCI (C/I = 40 dB).

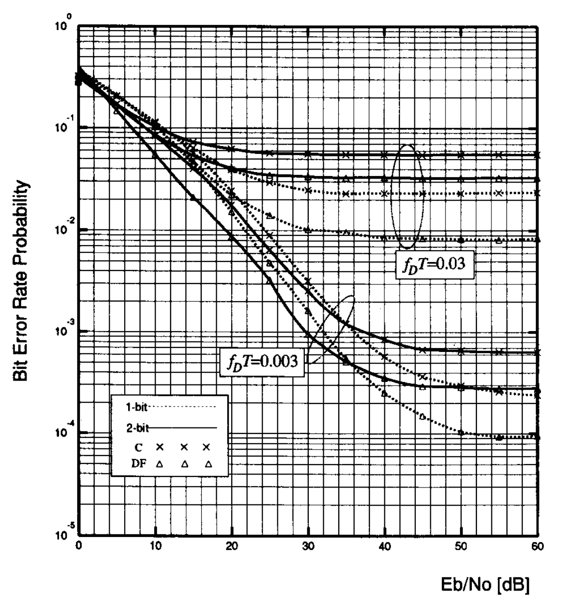

Fig. 5. BER performance of the conventional (C) and decision-feedback (DF) 1- and 2-bit differential detection receivers in the presence of fading (C/I $\rightarrow \infty$).

differential detector (at BER = 10^{-3}). At BER = 10^{-4}, the performance improvements are in excess of 18 dB. It is also expected that as C/I decreases, the overall performance gains will increase even further. The following two figures (Figs. 5 and 6) illustrate the performance of the proposed receivers in a Rayleigh fading environment at C/I $\rightarrow \infty$ (no CCI) and C/I = 40 dB with $N = 1$. Additional peformance evaluation results (e.g., for other values of the C/I) can be found in [20]. For all the graphs in Figs. 4–6, the dotted line and the solid line represent, respectively, 1-bit and 2-bit differential detection for $N = 1$, and the conventional receiver is denoted by an "x" and DF by a "triangle." The decision threshold is "0" for all cases, except a conventional 2-bit detection in a static environment where an optimized dc bias ν is applied [4]. It should be pointed out that the performance of the conventional receivers is in agreement with the analytical BER performance reported in [5]–[7], [12]. We conclude our discussion with the following observations:

1) 2-bit differential detection with the optimal threshold outperforms 1-bit in the "static" case. Indeed, this has also been the case in an AWGN and a Rician-fading environment [4]. However, as reported in [5]–[7], [12], 1-bit differential detection offers a better BER performance than 2-bit in a typical Rayleigh-faded CCI mobile channel.
2) The DF works well in an additive noise and/or interference environment, i.e., static CCI and AWGN (Fig. 4). However, in the fading environment, DF offers some reduction in the error floor (Figs. 5 and 6)—relatively little improvement compared to the static environment.

IV. Conclusions

The performance of several decision feedback differential receivers for GMSK signals transmitted over static and faded CCI channels has been evaluated. Our results have shown that the proposed receivers perform better than other more conventional receivers such as 1- or 2-bit differential detectors.

The obtained gains are more significant for the static CCI channel. For the faded CCI, error-floor reductions have been observed.

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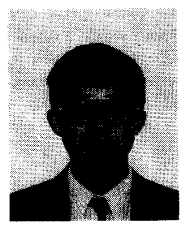

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P. Takis Mathiopoulos (S'79–M'89) For biography and photograph see page 257 of this issue.