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MULTIPLE TONE INTERFERERS IN AN FH-MFSK SPREAD SPECTRUM COMMUNICATION SYSTEM

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ABSTRACT

An analysis of the effect of multiple tone interferers on the bit error rate in a Frequency-Hopped Multiple Frequency Shift Keying (FH-MFSK) spread spectrum communication is given. A constant insertion rate detection strategy has been considered and a matched tuned filtering used in the receiver.

We have obtained results in the 20 MHz (one-way) bandwidth with a data rate of 32 Kb/s and a Rayleigh fading channel. The results show that adequate performance can be achieved even when 40 tone interferers are present with a signal to interference ratio of -10 dB and a signal to noise ratio of 25 dB.

INTRODUCTION

Recently, the use of spread spectrum for mobile radio communications has attracted the attention of many researchers. A multiple access FH-DPSK spread spectrum system design was first proposed by Cooper (1). This system was shown in (2) that it may coexist in the same frequency band with conventional narrow band systems without excesive mutual interference.

In this paper we discuss the performance of another FH-MFSK spread spectrum system, proposed by Goodman (3), in the presence of multiple tone interferers. This new system seems to perform better than the above mentioned FH-DPSK, when used as a multiple access system in an isolated cell case (4). In our analysis we assume as in (2) that there is at most one interfering tone present in a given frequency channel.

SYSTEM DESCRIPTION

In the FH-MFSK system illustred in Fig. 1, the modulator accepts every T seconds one of 2^K K-bit words, X_m, from the source and generates one of 2^K different tones that the system has available. The

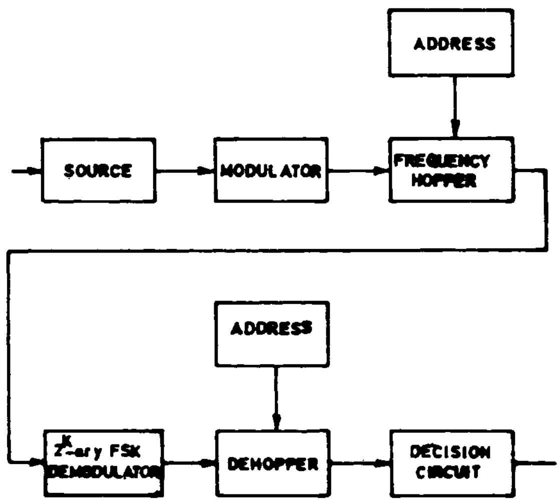

Figure 1. FH-MFSK System.

frequency hopper changes the frequency each time chip $\tau = \frac{T}{L}$ seconds according to the user's adress vector of length L. Each user is assigned a unique address vector $V_{m,q}$ (q=1,..L) which is used to distinguish his message from those of others. The transmitted tone sequence is assigned by the modulo 2^K sum ($\oplus$) of the address and the K-bit code word

$$Y_{m,q} = X_m \oplus V_{m,q}$$

At the receiver, demodulation and modulo 2^K subtraction ($\ominus$) by $V_{m,q}$ are performed in the 2^K-ary FSK demodulator and dehopper respectively, yielding

$$Z_{m,q} = Y_{m,q} \ominus V_{m,q} = X_m.$$

The sequence of operations is illustrated by the matrices of Fig. 2 and Fig. 3 where the 2^K tones have been placed at intervals of $1/\tau$ seconds.

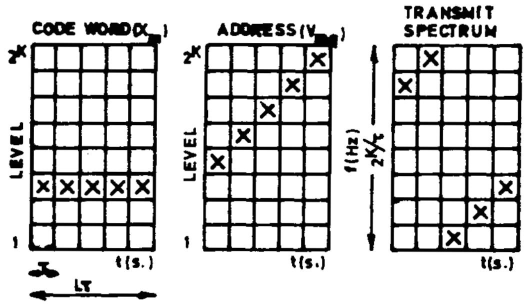

Figure 2. Transmitter Matrices.

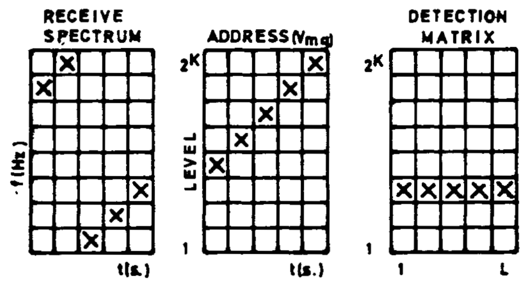

Figure 3. Receiver Matrices.

Noise, interfering signals, etc., can influence the detection by causing a tone to be detected when none has been transmitted (insertion). In addition, the receiver can omit a transmitted tone (miss). To allow for this possibility, we use the majority logic decision rule: choose the code word associated with the row containing the greatest number of entries.

INTERFERENCE ANALYSIS

The receiver signal has the form

$$S(t) = \sum_{i=-\infty}^{\infty} R \ \mathrm{rect}_{\tau}(t-i\tau) \cos (\omega_i^j t+\theta_i^j)$$

where, R is a Rayleigh distributed random variable, ω_i^j is the frequency corresponding to the frequency channel j and assigned to an ith. time chip, θ_i^j is a $(0,2\pi)$ uniform random variable and

$$\mathrm{rect}_{\tau}(t) = \begin{cases} 1 & \text{if } 0<t<\tau \\ 0 & \text{otherwise} \end{cases}$$

On the other hand, the interfering signal that affects the jth frequency channel

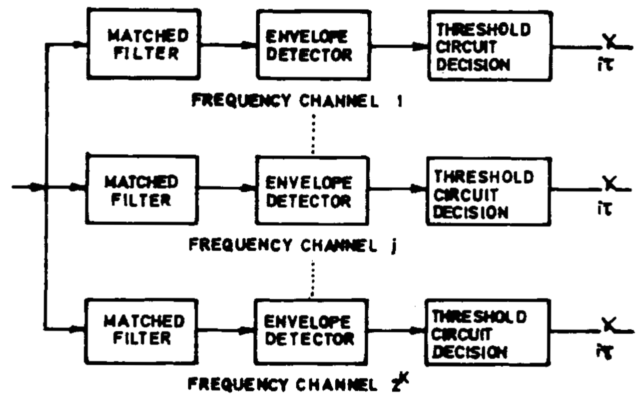

Figure 4. 2^K-ary FSK Demodulator.

is denoted as

$$x_j(t) = R_{I_j} \cos (\omega^j t+\phi_I^j)$$

where R_{I_j} is a Rayleigh distributed random variable, ω^j is a $(\omega_i^j - \frac{1}{2\tau}, \omega_i^j + \frac{1}{2\tau})$ uniform random variable and ϕ_I^j is a $(0,2\pi)$ uniform random variable. Hence, the total interfering signal is represented as

$$I_T(t) = \sum_{j=1}^{2^K} a_j \ x_j(t)$$

where $a_j \epsilon (0,1)$ is a binary random variable defined as

$$\mathrm{Prob}\ \{a_j=1\} = \frac{M}{2^K}$$

M is the number of interfering tones. As previously mentioned, at most only one interfering signal in every frequence channel has been allowed.

The transmitted signal suffers from the Rayleigh and Log-normal fading encountered in the mobile environnement. However, in our analysis a control power is assumed which maintains the mean power received to a preassigned value, so only the Rayleigh fading is considered.

The receiver signal, S(t), is then processed by a 2^K ary FSK demodulator as shown in Fig. 4. Each branch of this demodulator would perform as a noncoherent OOK optimum receiver if only their corresponding frequency channels were active. At the output, a majority decision rule is made in order to obtain the desired binary data. The received complex envelope signal in the frequency channel j (for sake of conciseness, the subindex j is omitted in the sequel) is formulated as:

$Z(t) = R\exp(-j\theta)+a\,R_I\exp(-j\phi_I)+n(t)$

$i\tau \leq t<(i+1)\tau.$

The first term in the formula corresponds to the desired signal, the second to the interfering signal and n(t) is white noise present at the receiver input, with N_0 being its one side spectral power density.

If h(t) is denoted as the complex envelope impulse response of the receiver matched filter, (Fig. 4) we can formulate the complex envelope signal in its output as

$$Z_0(t)=Z(t)*h(t)+$$

$$\left[\sum_{\substack{c=-IL\\c\neq 0}}^{IR} a_c\,R_{I_c}\exp(-j\phi_{I_c}).\exp(-j2\pi f_c t)\right]*h(t)$$

where we have taken into account the interference due to the IR higher and IL lower adjacent channels. Furthermore, a_c. R_{I_c} and ϕ_{I_c} are equally distributed as a, R_I and ϕ_I respectively, and f_c is a $(\frac{c}{\tau}-\frac{1}{2\tau},\frac{c}{\tau}+\frac{1}{2\tau})$ uniform random variable. Since the receiver filter is matched to a rectangular envelope signal, we have

$$h(t) = \mathrm{rect}_\tau(t)$$

and

$$Z_0(\tau) = \tau.R\exp(-j\theta) +$$

$$\sum_{c=-IL}^{IR} a_c.\tau.R_{I_c}\exp(-j\phi_{I_c})\,\mathrm{sinc}(f_c\tau)+n_f(\tau)$$

where

$$\mathrm{sinc}(x) = \frac{\sin\pi x}{\pi x}\ ;\ a_0=a\ ;\ R_{I_0}=R_I\ ;\ \phi_{I_0}=\phi_I$$

and $n_f(t)$ is the filtered white gaussian noise, which is represented as

$$n_f(t) = n_x(t)+j\,n_y(t)$$

$n_x(t)$ and $n_y(t)$ being statistically independent, and

$$E\{n_x(t)\} = E\{n_y(t)\} = 0$$

$$\mathrm{Var}\{n_x(t)\} = \mathrm{Var}\{n_y(t)\} = N_0\tau$$

BIT ERROR RATE CALCULATION

The bit error rate (BER) requires the previous computation of the insertion and miss probabilities. Once these probabilities are known, the BER can be easily obtained (3).

The insertion probability is formulated as

$$P_I= \mathrm{Prob}\{|Z_0(\tau)-\tau R\exp(-j\theta)|\geq C_0\} =$$

$$= \mathrm{Prob}\left\{\left|\sum_{c=-IL}^{IR} a_c\tau.R_{I_c}\exp(-j\phi_{I_c}).\mathrm{sinc}(c+b_c)+n_f(\tau)\right|\geq C_0\right\}$$

where $b_c=f_c\tau-c$ is a $(-\frac{1}{2},\frac{1}{2})$ uniform distributed random variable and C_0 is the threshold decision value. By defining

$$r = \sum_{c=-IL}^{IR} a_c\,R_{I_c}\exp(-j\phi_{I_c})\ \mathrm{sinc}(c+b_c)$$

we can write (5).

$$\mathrm{Prob}\,\{Z_0(\tau)\geq C_0|a_c,b_c\} = \exp(-\beta_1^2/2)$$

where

$$\beta_1^2 = (\frac{C_0}{\tau})^2\frac{2}{E\{|r|^2\}+\frac{1}{\tau^2}.E\{|n_f(\tau)|^2\}} =$$

$$= \beta^2\frac{1}{1+G.\rho.\left[\sum_{c=-IL}^{IR} a_c.\mathrm{sinc}^2(c+b_c)\right]}$$

and

$$\beta^2= (\frac{C_0}{\tau})^2/\frac{N_0}{\tau}$$

$$G = \frac{E\{R_{I_c}^2\}}{E\{R^2\}} = \frac{E\{R_I^2\}}{E\{R^2\}}$$

$$\rho = \frac{1}{2}\frac{E\{R^2\}.\tau}{N_0} = \frac{E_c}{N_0}$$

The insertion probability could now be formulated as the multiple integral

$$P_I=\int_{-\frac{1}{2}}^{\frac{1}{2}}\cdot\cdot\int_{-\frac{1}{2}}^{\frac{1}{2}}\int_{-\infty}^{\infty}\cdot\cdot\int_{-\infty}^{\infty}\exp\{-\frac{\beta^2}{2}.\frac{1}{1+G.\rho.\left[\sum_{c=-IL}^{IR} a_c.\mathrm{sinc}^2(c+b_c)\right]}\}$$

$$.\,db_{IR}\cdots db_{IL}.f_{a_{IR}}(x_{IR})\cdots f_{a_{IL}}(x_{IL}).dx_{IR}\cdots dx_{IL}$$

where

$$f_{a_c}(x_c) = \frac{M}{2^K}\,\delta(x_c-1)+(1-\frac{M}{2^K})\,\delta(x_c)$$

However, the direct calculation of P_I becomes too cumbersome and another method is envisaged; so, by denoting $f_V(x)$ the probability density function of the random variable

$$V = \sum_{c=-IL}^{IR} a_c \; \text{sinc}^2(c+b_c)$$

then

$$P_I = \int_0^{\infty} \exp\left(-\frac{\beta^2}{2} \cdot \frac{1}{1+G.\rho x}\right) . \; f_V(x) \; dx$$

and P_I is evaluated using a Gaussian Quadrature Rule as

$$P_I \simeq \sum_{u=1}^{N} \exp\left(-\frac{\beta^2}{2} \frac{1}{1+G.\rho x_u}\right).W_u$$

where x_u and W_u, called the abscissas and the weights of the formula respectively, are obtained from the algorithm introduced in (6) by using the 2N+1 moments of the random variable V:

$$m_n = E\{V^n\} \quad n=0, \; ... \; , \; 2N$$

Due to that V is formed by the addition of IL+IR+1 statistically independent terms, m_n is easily obtained from the moments of each term:

$$m_{cn} = E\{[a_c \text{sinc}^2(c+b_c)]^n\} = \frac{M}{2^K} \int_{-\frac{1}{2}}^{\frac{1}{2}} \text{sinc}^{2n}(c+b_c) \; db_c$$

Analogously, as we did with P_I, the miss probability is formulated as

$$P_{MISS} = \text{Prob}\{|\tau R \exp(-j\theta)+r.\tau+n_f(\tau)|<C_0\}$$

Bearing in mind that $\tau R \cos\theta$ and $\tau R \sin\theta$ are independent Gaussian random variables, we can include them in $n_x(\tau)$ and $n_y(\tau)$. Therefore

$$P_{MISS} = 1-\int_0^{\infty} \exp\left[-\frac{\beta^2}{2} \frac{1}{1+\rho.(1+Gx)}\right] f_V(x) \; dx$$

the above integral is computed by the GQR already used to calculate P_I.

NUMERICAL RESULTS

The design variables K=8, L=19 and W=20MHz were chosen as in (3) in order to get analytical results. Moreover, two reasonable assumptions were made.

a.- We did not take into consideration the coherence bandwidth of the mobile channel. This approach is justified by that, on the average, the tones emitted by a station are separated by $\frac{2^K}{L\tau} \simeq$ 1MHz, a value not lower than the usual coherence bandwidth encountered, which ranges from 200 KHz a 1 MHz.

b.- When we compute the BER from P_I and P_{MISS}, we assume that both P_I and P_{MISS} are independent from the chip and frequency channels positions. In reality, one isolated tone interfering would cause L chip positions to be equally distributed; however, the great number of tone interferers involved and its random mixing with the adress code in the receiver enforces the above assumption of independence.

On the other hand, one major difficulty is that setting the optimum threshold value that minimizes the BER, requires knowing the signal to noise ratio and the interference activity. This can be avoided by counting the fraction of threshold crossings in the receiver and over a reasonable number of chips; then, as suggested in (7), setting the insertion rate to a fixed value. If this fraction exceeds the nominal insertion rate, the threshold value, C_0, is increased by an amount of ΔC_0 and conversely it is decreased if the fraction is less than the nominal P_I.

In the numerical results of this section the convergence of the GQR was obtained, at least, up to the first two significants digits. Moreover, IL=IR=4 were considered, since by increasing this number we do not obtain a significant variation of the BER.

Fig. 5 shows BER in relation to M with a signal to interference ratio (CIR) of 0dB. Fixed P_I values ranging from 0.2 to 0.5 were considered. The signal to noise ratio (SNR) chosen was 25dB because it is a typical value in a mobile environment. Values of P_I lower than 0.2 cause an inadequate performance if the number of tone interferers is high. Values of P_I greater than 0.5 cause an inadequate performance in any case. Fig. 6 and 7 show the behaviour of the FH-MFSK system for different CIR values. Two families of curves are plotted for P_I=0.3 and P_I=0.4 and SNR=25dB. The number of tone interferers is used as a parameter. By comparing both families of curves we can observe that the optimum P_I value depends on CIR and M. However by observing the above figures, $P_I \simeq 0.3$ seems to be a good value in most cases. If M<40 and CIR>0dB, then lower values of P_I could be

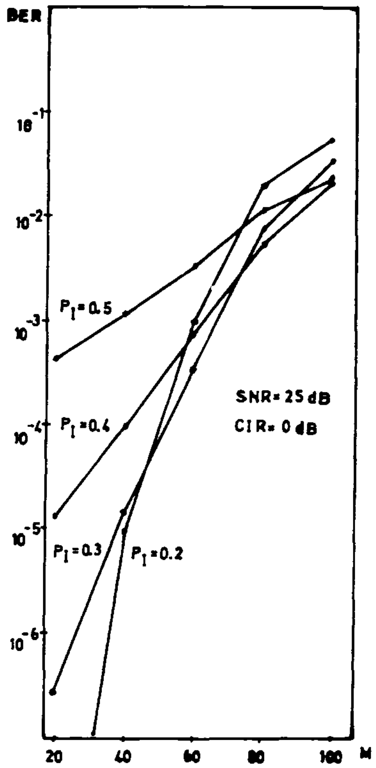

Figure 5. Bit Error Rate for several number of interferers.

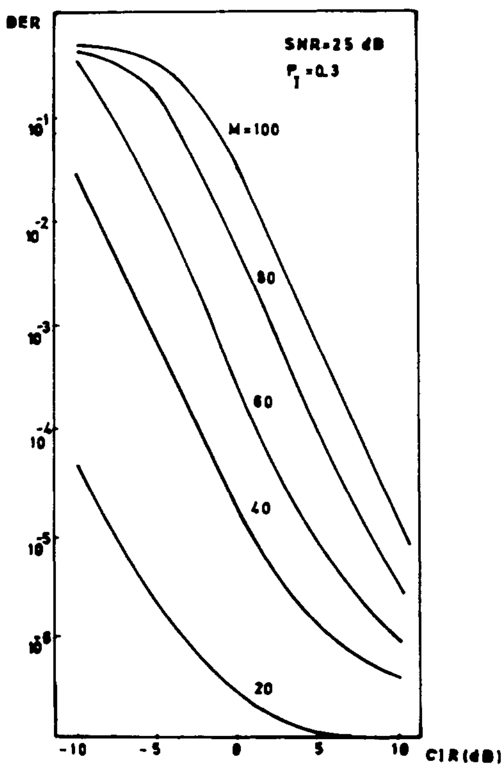

Figure 6. Bit Error Rate versus CIR for P_I=0.3.

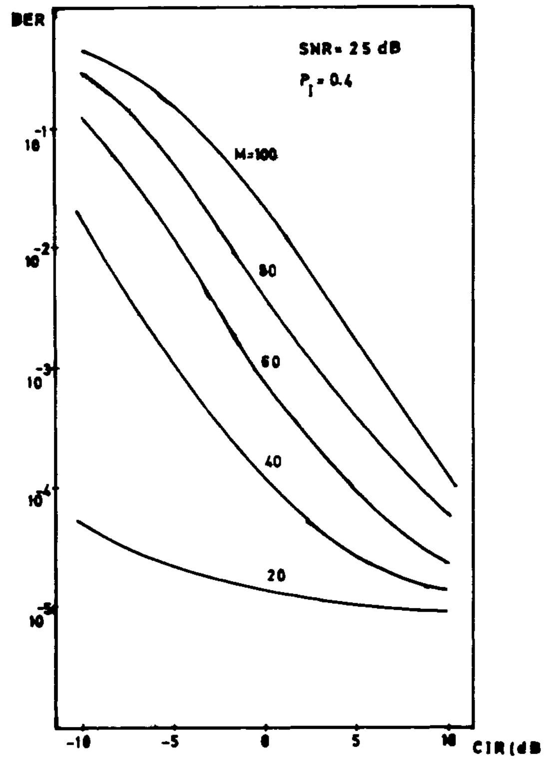

Figure 7. Bit Error Rate versus CIR for P_I=0.4.

considered.

Comparing the behaviour of the FH-MFSK system studied with the FH-DPSK analyzed in (2), we do note the better performance of the first system. In particular, FH-MFSK can maintain BER$<10^{-3}$ with up to about 40 tone interferers and CIR$\simeq$-5dB. Even about 40 tone interferers would not cause the performance to be worse than BER=10^{-2} if CIR=-10dB.

CONCLUSION

In this paper we have discussed the performance of an FH-MFSK spread spectrum communication system contaminated by tone interferers in a Rayleigh channel. One finality of this work has been to compare the mentioned system with another proposed system, FH-DPSK, in the same interference environment. The results obtained have shown, even allowing for a margin of error due to the assumptions made, a better performance of the FH-MFSK system when compared with the results obtained in (2). In particular, by using a constant insertion rate, $P_I\simeq 0.3$, up to aproximately 40 tone interferers can be supported with a CIR=-10dB and BER$\simeq 10^{-2}$ in the FH-MFSK system.

However, when CIR<0dB, the FH-DPSK system performance degrades enormously.

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