PERFORMANCE OF DUOBINARY DIGITAL FM IN AN INTERFERENCE ENVIRONMENT

ALI EL Moghazy*, S.El habiby*, E.M.Rateb**, M.Fouad ***

* Ass.Professors - Comm.Department- M.T.C. Cairo, Egypt.
** Proffessor -M.T.C.
*** Eng.(MSc.). Signal Corp.

RESUME

On expose ici l'évaluation de la performance d'un système duobinaire numerique utilizant une modulation en frequence. La detection est basée sur un ecrateur-discriminateur en considerant un bruit "blanc gaussien" additive.
Or , un modéle de la simulation et un progarm sont fournis permettant d'avoir la r[illegible]ion entre le taux d'erreur(Pe)et la ra[illegible]t signal-sur-bruit(C/N),pour des valeurs differentes du raport signal-sur-brouillage (C/I).Le signal duobinaire est supposé limité en bande ($\pm f_b/2$, autour de la frequence porteuse).Le resultant de la simulation permit de comparer la performance du systèmeconsideré avec d'autre systèmes telle que (OOK,NCFSK , DQPSK,8PSK). La comparison est faite en considerant un brouillage à onde continue et le rapport C/I = 10 et 15 dB.

SUMMARY

This paper is concerned with the performance evaluation of duobinary digital FM modulation system, with limiter-discriminator detection in the presence of additive white gaussian noise (AWGN), CW and modulated cochannel interference. To allow for the performance evaluation, a simulation model and a computer program are build up. The relation between the error probability (p_e) and the carrier-to-noise ration(C/N) is given for different values of carrier-to-interference ratio (C/I). The duobinary FM modulated signal is assumed to be band-limited to $\pm f_b/2$ with respect to the carrier frequency. This corresponds to 99% of power at modulation index d= 0 .375. Comparison with published results concerning other digital modulation techniques is performed . The results show that,at $p_e=10^{-4}$ the required E_b/N_o for the concerned modulation technique is largerthan that of BPSK, DPSK, QPSK and smaller than that of OOK,NCFSK, DQPSK ,8BSK. However ,it is found that at $p_e=10^{-4}$ the degradation (in terms of the db increment of E_b/N_o to be added to get performance without interference) of the concerned modulation is smaller than the values published for the above mentioned modulation techniques. The comparison is performed for CW interference at C/I= 10 , 15 db.

I- INTRODUCTION

In microwave communication systems, interference is considered as one of the major transmission impairements that causes according to its nature, a specific amount of performance degradation.
Interference can be classified as cochannel or adjacent channel,according to whether the interfering signal carrier cooincides or not with victim signal. It can also be either unmodulated carrier wave (CW) or modulated,with a type of modulation, usually, of the same type as the victim one. Most of the previous work concerning performance evaluation of digital modulation system in the presence of interference is devoted to phase shift keying(PSK)and related variants(M-ry PSK , DPSK,....).Oetting[1], in his excellent state of-the-art article gives table comparing several digital modulation techniques in CW interference environments for C/I = 10 db and 15 db.
Duobinary digital FM modulation system[2], found special interest in practical microwave line of sight radios[3]. This is because of its spectral effeciency.The analytical study of this type of modulation is difficult due to the inherent correlation between the adjacent symbols. In this paper the performance evaluation is presented using simulation technique. In section II, the system simulation model is described. In section III, the results of siumulation are presented with comparison to published data for other modulation techniques, and section IV is the conclusion.

II. SYSTEM SIMULATION MODEL

Figure(1)shows the duobinary digital FM system simulation model.The input data sequence d(t) is represented by:

$$d(t)=\sum_{n=-\infty}^{\infty} a_n \, \delta(t-nT) \qquad (1)$$

where a_n is a random variable assuming equiprobable values of ± 1, $\delta(t)$is the Dirac impulse and T is the bit duration.This sequence is generated using a proper transformation of uniformly distributed random variable. The precoder performs the operation $c_n=a_n+c_{n-1}$ to prevent error propagation in the detection process[2]. Noting that,while the precoding is performed the input data (a_n)are stored for the purpose of error probability evaluation.Formally,the precoder output is giving by:

$$p(t)=\sum_{n=-\infty}^{\infty} c_n \, \delta(t-nT) \qquad (2)$$

The duobinary filter is represented by its impulse response h(t) as [2]

$$h(t)=\frac{4\cos(\pi t/T)}{\pi[1-4(t/T)^2]} \qquad (3)$$

and its output x (t) is given by :

$$x(t)=p(t) * h(t)=\sum_{n=-\infty}^{\infty} a_n\, h(t-nT),$$

where $*$ stands for the convolution integral the signal $x(t)$ is used as a modulating signal for the frequency modulator whose output $S(t)$ is given by:

$$S(t)=S\cos\left(2\pi f_c t+2\pi f_d\int_{-\infty}^{t} x(\tau)\,d\tau\right) \quad (5)$$

S, f_c being the unmodulated carrier amplitude and frequency respectively and f_d is the maximum frequency deviation. For our purpose the quantity of interest is $\Phi(t)$ given by:

$$\Phi(t)=2\pi f_d\int_{-\infty}^{t} x(\tau)\,d\tau = 8 f_d\int_{-\infty}^{t} y(\tau)\,d\tau \quad (6)$$

where $y(t)=\sum_{n=-\infty}^{\infty} c_n\, w(t-nT)$

and $w(t)=\dfrac{\cos(\pi t/T)}{1-4(t/T)^2}$

$W(t)$ could be considered zero outside the time interval $|t|<3.5T$. This approximation is mandatory for the numerical integration.

Since, at the discriminator side, the values of $\Phi(t)$ are needed at successive instants t_m and $t_m-\Delta t$, where t_m is equal to $mT-T/2$, $m=1, 2, \ldots\ldots\ldots$

Then :

$$\Phi(t_m)=\Phi(t_m-T)+8 f_d\int_{t_m-T}^{t_m} y(t)\,dt \quad (7.a)$$

$$\Phi(t_m-\Delta t)=\Phi(t_m-T)+8 f_d\int_{t_m-T}^{t_m-\Delta t} y(t)\,dt \quad (7.b)$$

The integration over $y(t)$ $[w(t)]$ is performed numerically using the generalized trapezoidal method with integration step size equal to $T/100$, a value that accounts for both accuracy and speed of calculation. Equations (7.a) and (7.b) leads to the followings [4]:

$$\Phi(t_m)=\Phi(t_m-T)+0.04\, d\,[(c_{m-4}+C_{m+2})\,w1+(C_{m-3}+C_{m+1})\,w2+(c_{m-2}+C_m)\,w3+C_{m-1}\,w4] \quad (8.a)$$

and

$$\Phi(t_m-\Delta t)=\Phi(t_m-T)+0{,}04\, d\,(C_{m-4}v1+C_{m-3}v2+C_{m-2}v3+C_{m-1}v4+C_m v5+C_{m+1}v6+C_{m+2}v7) \quad (8.b)$$

where $d=f_d T$ is the modulation index, and the values of $w1$, $w2, \ldots\, w4$ and $v1, v2, \ldots\, v7$ are given by:

w1 = 3.700155	w3 = 70.90823
w2 = -8.85789	w4 = 185.1911
v1 = 3.699497	v2 = 8,85657
v3 = 70.90425	v4 = 183.2125
v5 = 69.34534	v6 = -8.85402
v7 = 3.698857	

The narrow band gaussian noise $n(t)$ given by:

$$n(t)=n_c(t)\cos w_c t-n_s(t)\sin W_c t \quad (9)$$

where $n_c(t)$ and $n_s(t)$ are statistically independent gaussian processes. Again, the values of interest at the receiver side are the values of $n_c(t)$ and $n_s(t)$ at two successive instants t_m, $t_m-\Delta t$. This yeild zero mean gaussian random variables $n_c(t_m)$, $n_c(t_m-\Delta t)$, $n_s(t_m)$ and $n_s(t_m-\Delta t)$.

The random varaible $n_c(t_m)$ $[n_s(t_m)]$ and $n_c(t_m-\Delta t)$ $[n_s(t_m-\Delta t)]$ are correlated. Their correlation coefficient ρ is calculated from the autocorrelation function $R(\tau)$ of the baseband process $n_c(t)$ that given by:

$$R(\tau)=N_o f_b \frac{\sin \pi f_b \tau}{\pi f_b \tau} \quad (10)$$

where N_o is the one sided noise power spectral density, and f_b is the IF filter bandwididth and is numerifcally equal to the rate $1/T$. The values of ρ is equal to $R(\Delta t)/\sigma^2$, $\sigma^2=N_o f_b$ being the noise variance. For $\Delta t=T/100$, $\rho=0.9998355$.

In simulating these correlations between the random variables we use the following relations [5]:

$$E\,[n_c(t_m-\Delta t)\,/\,n_c(t_m)=n_c]=\rho\, n_c \quad (11.a)$$

and

$$\sigma^2[n_c(t_m-\Delta t)/\,n_c(t_m)=n_c]=\sigma^2(1-\rho^2) \quad (11.b)$$

Where n_c accounts for the current value of $n_c(t_m)$.

The same applies for $n_s(t_m)$ and $n_s(t_m-\Delta t)$

The noise variance σ^2 is calculated for a given carrier-to-noise ratio (C/N) by:

$$C/N=10\log_{10}\frac{S^2}{2\sigma^2} \quad (12)$$

The interference considered is of cochannel type whose expression is given by:

$$i(t)=I\cos(2\pi f_c t+\theta) \quad (13)$$

where I is the interference, and θ is considered to be uniformly distributed random variable. For (CW) case, and duobinary FM modulated with the same conditions as the victim signal for "digitally modulated" case. The carrier-to-interfernce ratio C/I is defined by:

$$C/I=10\log_{10}\frac{S^2}{I^2} \quad (14)$$

The limiter action is implicitly taken into consideration by the "ideal" discriminator whose output depends solely on the phase of its input. The phase of the input signal to the discriminator is given by:

$$\gamma(t)=\tan^{-1} Y(t)/X(t) \quad (15)$$

where

$$X(t)=S\cos\Phi(t)+n_c(t)+I\cos\theta$$

$$Y(t)=S\sin\Phi(t)+n_s(t)+I\sin\theta$$

the discriminator output $D(t)$ at instant t_m is given by $k_d\,\dot{\gamma}(t)$, $t=t_m$ with $k_d=1/2\pi f_d$ being the discriminator gain, then:

$$D(t_m)=k_d\,\frac{X(t_m)\,\dot{Y}(t_m)-Y(t_m)\,\dot{X}(t_m)}{X^2(t_m)+Y^2(t_m)} \quad (16)$$

The differentiation considered in (16) are related to random processes $\Phi(t)$, $\theta(t)$ (in case of digitally modulated), $n_c(t)$, $n_s(t)$.

These derivatives exist (in the mean square sence) if corresponding autocorrelation function are differenctiable to the order 2 [5]. This codition is obvious for $n_c(t)$, $n_s(t)$ and is supposed to be satisfied for the others.

The output of the sampler is compared to 2 thresholds at ± 1 and the decision rule is to assign -1 to $\hat{a}$ if $-1 > D(t_n) > 1$ and to assign $+1$ to $\hat{a}$ otherwise. The recovered $\hat{a}_n$ is compared to the corresponding transmitted bit a_n and an error is counted if they are different. The probability of errors and the total number of observed bits. The number of observed bits are choosen equal to $10/p_e$ to allow for acceptable level of accuracy [6]

III. RESULTS AND COMMENTS

Figure (2) shows the dependence of the probability of error p_e on the C/N for different values of C/I ratio. To allow for comparison with other digital modulation techniques, table (1) presents the required E_b/N_o (which in our case equal to C/N) for error rate $P_e = 10^{-4}$. Moreover the values of degradation for CW interfernce, at C/I = 10 db and 15 db are given. The reference for the values in the table (other than the duobinary FM) is [1]. The comparison shows that, at $p_e = 10^{-4}$, the required E_b/N_o for the concerned modulation technique is larger than that of BPSK, DPSK, QPSK and smaller than that of OOK, NCFSK, DQPSK, 8PSK, However, it is found that at $p_e = 10^{-4}$ the degradation of the concerned modulation is always smaller than the values puplished for the above mentioned modulation techniques.

IV. CONCLUSION

Using simulation technique, the performance of duobinary digital FM is evaluated in the presence of AWGN, CW interference and digital modulated interference. The results of the simulation shows the resistive property of this type of modulatuon to interference.
This conclusion is supported by field measurments of microwave radio links using different equipments of different modulation techniques,

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Table (1)

Typs	Modulation Scheme	E_b/N_o* (db)	E_b/N_o* Required for CW interference		Degradation at	
			C/I = 10 db	C/I = 15	C/I=10 db	C/I=15 db
PM	OOK-erivelop detection	11.9	20.0	14.5	8.1	2.6
FM	FSK-noncherent (d=1)	12.5	14.7	13.4	2.2	0.8
	Duobinary FM	11.3	13.0	12.0	1.7	0.7
PM	BPSK (COHERENT)	8.4	10.5	9.2	2.1	0.8
	OPSK	9.3	12.5	10.3	3.2	1.0
	OPSK	8.4	12.2	9.8	3.8	1.4
	DQPSK	10.7	> 20.0	14.2	>9.3	3.5
	SPSK (coherent)	11.8	20.0	15.8	8.2	4.0

* $P = 10^{-4}$

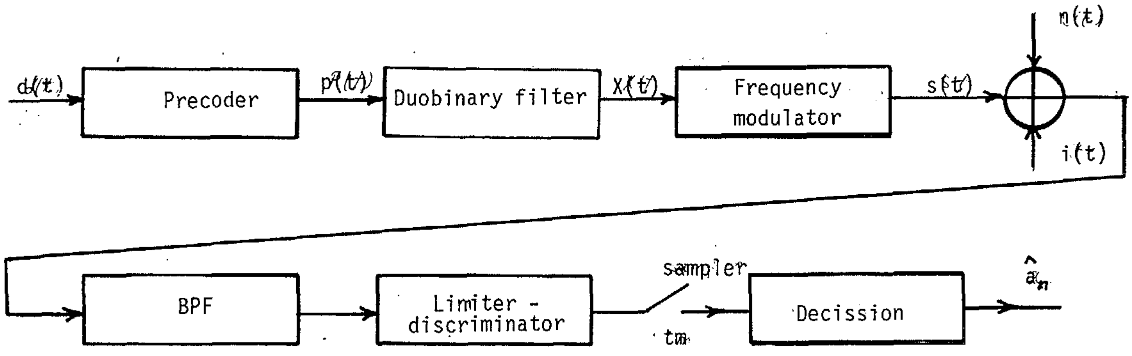

Fig (1) Duobinary FM system simulation model

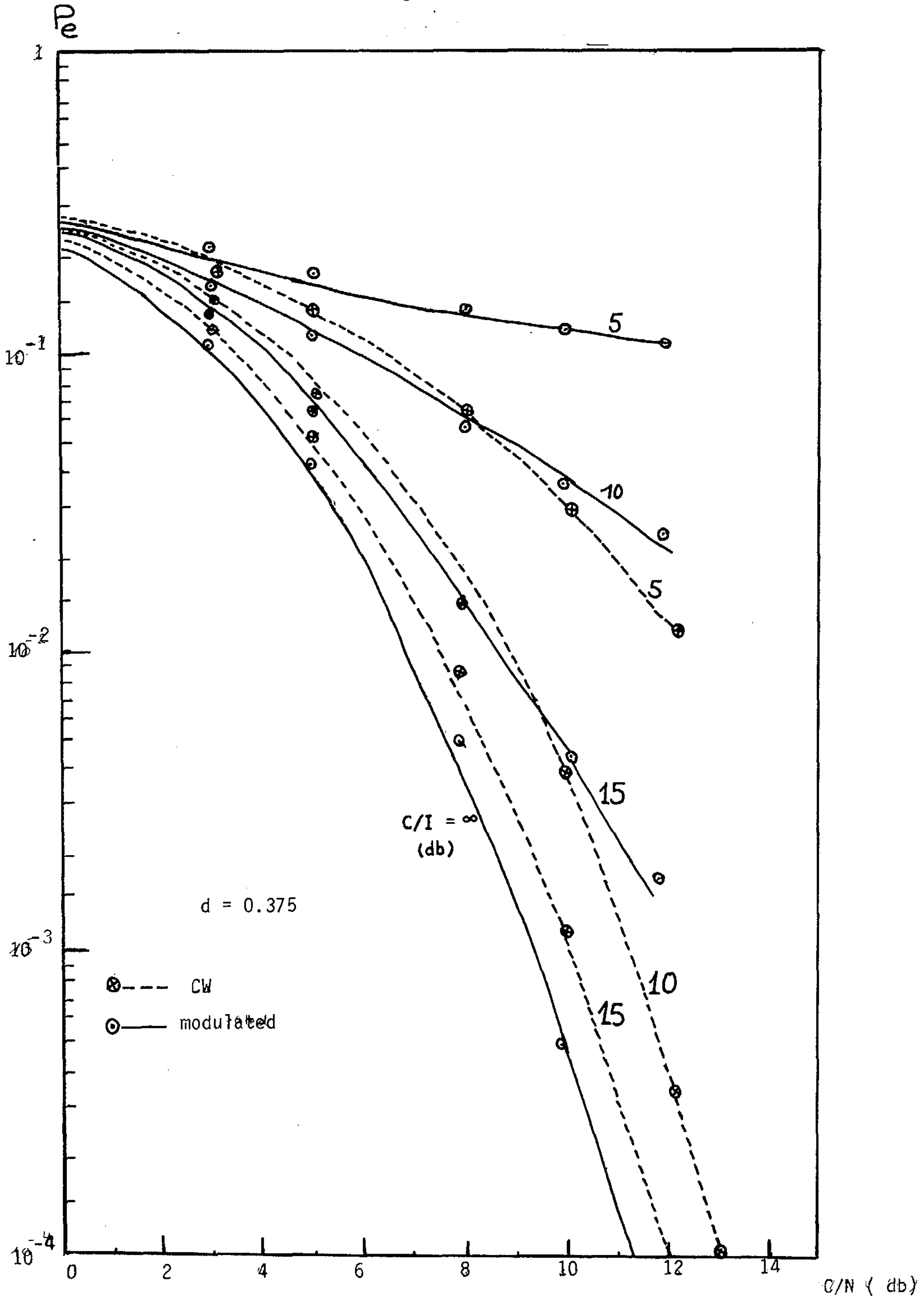

Fig (2) Performance in the presence of cochannel interference.