ENHANCED CW JAMMER MITIGATION USING PARTIALLY COHERENT CPM RECEPTION

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Abstract— **A partially coherent detector for continuous phase modulation with a cochannel CW jammer is analyzed. Improved bit error rate performance of the partially coherent detector over the noncoherent detector in a CW jamming environment is shown. The investigation identifies the dependence of performance on the modulation index variation versus the interference power-to-signal power ratio. The selection of the optimum modulation index to achieve the minimum probability of bit error while experiencing cochannel CW jamming is discussed.**

I. INTRODUCTION

Continuous Phase Modulation (CPM) coherent demodulation, which uses a phase-locked-loop (PLL) for carrier frequency and phase recovery, has been extensively studied in [1], [2], and [3]. In a hostile environment, such as a military application, a prime jamming strategy in a coherent system is to attack the center PLL frequency. With increasing jammer power the probability increases for PLL cycle slips [4]. Noncoherent systems are used to deny the jammer this advantage [5]. Noncoherent demodulation of CPM has been studied in [6] and [7].

Partially coherent (PC) demodulation attempts to take advantage of the positive attributes of both a coherent and noncoherent receiver by adapting the receiver performance based on the locked condition of the PLL. This adapting is accomplished by maintaining an estimate of quality for the carrier phase. A poor quality phase estimate results in noncoherent reception, while a good quality phase estimate results in coherent detection. The generation of this quality of the phase estimate is not addressed in this paper, but it can be obtained using parameter estimation theory [8] and knowledge of the state of the PLL (acquisition, in-lock, etc.).

PC demodulation has been less well studied since the theory was developed by Viterbi [9]. The first application of PC detection to CPM was performed by Aulin and Sundberg [10], but didn't consider the receiver's degraded performance caused by an interfering signal. Further analysis of PC demodulation applied to CPM by [12] demonstrated the amount of system degradation that occurs when an interfering binary signal is present.

The single-tone jammer is important because the jamming signal is easy to generate and is effective against direct sequence spread spectrum systems [13]. This paper analyzes the effect a cochannel CW jammer has on a binary CPM partially coherent receiver. The performance over a range of CPM modulation indices is investigated, and the optimum value to reduce the effect of the jammer is identified. The receiver will utilize sequence detection, because it provides a greater improvement in performance than bit-by-bit detection.

II. PARTIALLY COHERENT DETECTION

The full response transmitted CPM signal is a constant amplitude modulation and has the form

$$s(t,\alpha) = \sqrt{\frac{2E}{T}}\cos[\omega_0 t + \phi(t,\alpha)], \tag{1}$$

where E is the signal energy, T is the symbol period, ω_0 is the radian carrier frequency, and the data is contained in the phase function

$$\phi(t,\alpha) = 2\pi h \sum_{i=-\infty}^{\infty} \alpha_i q(t-iT). \tag{2}$$

The non-negative constant h is called the modulation index and each data symbol α_i takes on one of two binary values ± 1. The pulse shape of the phase function, $q(t)$, is the integral of the full response instantaneous rectangular frequency function 1REC [3], that when normalized yields

$$q(t) = \begin{cases} 0, & t < 0 \\ \frac{t}{2T}, & 0 \le t < T \\ \frac{1}{2}, & t \ge T. \end{cases} \tag{3}$$

The partially coherent receiver development that models the PLL carrier phase using the Tikhonov distribution is discussed in [9], and has been analyzed for CPM in [10]. A block diagram of the receiver is shown in Fig. 1 using sequence observation over the interval $[-N_1T, N_2T]$, where N_1 and N_2 are non-negative integers. The receiver complexity increases as $2^{N_1+N_2}$ due to the symbol sequence detection over a binary signal space. However, it was shown in Aulin and Sundberg [10] that observation over just three symbols ($N_1 + N_2 = 3$) approaches closely the optimum detection

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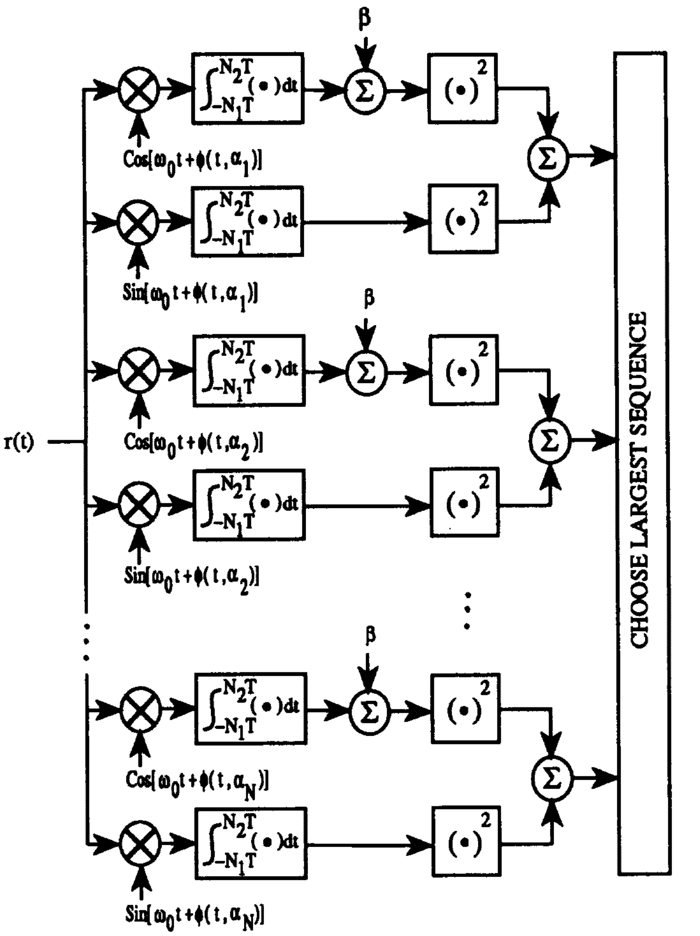

Figure 1: Binary Partially Coherent CPM Sequence Receiver with $N = 2^{N_1+N_2}$.

over an infinite sequence. Consequently, three symbol detection will be used in this analysis with the partially coherent receiver.

III. PROBABILITY OF ERROR

The probability of error is determined for the partially coherent receiver with binary symbols for detecting a desired signal while simultaneously receiving an interfering CW signal. The interfering signal has no phase information and a random carrier phase offset (Θ) uniformly distributed over $[0, 2\pi]$. It is assumed for simplicity in computation that the interfering CW signal is time synchronized with the desired signal. Although this is unrealistic in practical systems, time synchronization can be justified as a worst-case scenario. The received signals at the input to the demodulator of Fig. 1 are then

$$r(t) = s(t,\alpha) + \sqrt{\frac{2E}{T}}\sqrt{\frac{I}{K}}cos[\omega_0 t + \Theta] + n(t), \qquad (4)$$

where $s(t,\alpha)$ is the desired signal given in (1) and $n(t)$ is additive white Gaussian noise with one-sided spectral density of N_0. The second sinusoid is the interfering signal with $\sqrt{I/K}$ the interference-to-carrier power ratio that takes on real values from zero to infinity, with a value of zero corresponding to no interference.

The receiver in Fig. 1 selects the data sequence estimate, $\tilde{\alpha}$, of the true sequence, α, that maximizes the likelihood metric, given the interfering CW signal phase Θ. An error occurs when the incorrect sequence decision metric, $R(\alpha,\tilde{\alpha})$, is larger than the correct decision metric, $R(\alpha,\alpha)$. The probability of symbol error is the probability of all possible error sequences causing an incorrect decision metric to exceed the correct decision metric. Using the union bound for this probability results in an upper bound of

$$\begin{aligned} Pr(error \mid \Theta) &\leq \sum_i Pr[R(\alpha,\tilde{\alpha}_i) > R(\alpha,\alpha) \mid \Theta] \\ &= \sum_i Pr[|z_1| > |z_2| \mid \Theta], \end{aligned} \qquad (5)$$

where z_1 and z_2 are complex correlated Gaussian random variables defined from the decision metrics in [12] as

$$\begin{aligned} z_1 &\triangleq c(\alpha,\tilde{\alpha}) + \sqrt{\tfrac{I}{K}}c_I(\tilde{\alpha},\Theta) + \sqrt{\tfrac{2T}{E}}n_c(\tilde{\alpha}) + \tfrac{1}{2}\beta \\ &+ j\left[s(\alpha,\tilde{\alpha}) + \sqrt{\tfrac{I}{K}}s_I(\tilde{\alpha},\Theta) + \sqrt{\tfrac{2T}{E}}n_s(\tilde{\alpha})\right] \end{aligned} \qquad (6)$$

and

$$\begin{aligned} z_2 &\triangleq c(\alpha,\alpha) + \sqrt{\tfrac{I}{K}}c_I(\alpha,\Theta) + \sqrt{\tfrac{2T}{E}}n_c(\alpha) + \tfrac{1}{2}\beta \\ &+ j\left[s(\alpha,\alpha) + \sqrt{\tfrac{I}{K}}s_I(\alpha,\Theta) + \sqrt{\tfrac{2T}{E}}n_s(\alpha)\right], \end{aligned} \qquad (7)$$

and where we have introduced the interference effects with the following definitions:

$$c(\alpha,\tilde{\alpha}) \triangleq \frac{1}{T}\int_{-N_1T}^{N_2T} cos\,[\phi(t,\tilde{\alpha}) - \phi(t,\alpha)]\,dt, \qquad (8)$$

$$s(\alpha,\tilde{\alpha}) \triangleq \frac{1}{T}\int_{-N_1T}^{N_2T} sin\,[\phi(t,\tilde{\alpha}) - \phi(t,\alpha)]\,dt, \qquad (9)$$

$$\begin{aligned} c_I(\alpha,\Theta) &\triangleq \sum_{j=-N_1}^{N_2-1} \frac{-2\sin(0.5\pi h\alpha_j)}{\pi h\alpha_j} \\ &\cdot \cos\left(\pi h\left\{\sum_{i=-N_1}^{j-1}\alpha_i + (\tfrac{1}{2}+j)\alpha_j\right\} - \Theta\right), \end{aligned} \qquad (10)$$

$$\begin{aligned} c_I(\tilde{\alpha},\Theta) &\triangleq \sum_{j=-N_1}^{N_2-1} \frac{-2\sin(0.5\pi h\tilde{\alpha}_j)}{\pi h\tilde{\alpha}_j} \\ &\cdot \cos\left(\pi h\left\{\sum_{i=-N_1}^{j-1}\tilde{\alpha}_i + (\tfrac{1}{2}+j)\tilde{\alpha}_j\right\} - \Theta\right), \end{aligned} \qquad (11)$$

$$s_I(\alpha,\Theta) \triangleq \sum_{j=-N_1}^{N_2-1} \frac{-2\sin(0.5\pi h\alpha_j)}{\pi h\alpha_j} \cdot \sin\left(\pi h\left\{\sum_{i=-N_1}^{j-1}\alpha_i + (\tfrac{1}{2}+j)\alpha_j\right\} - \Theta\right), \quad (12)$$

$$s_I(\tilde{\alpha},\Theta) \triangleq \sum_{j=-N_1}^{N_2-1} \frac{-2\sin(0.5\pi h\tilde{\alpha}_j)}{\pi h\tilde{\alpha}_j} \cdot \sin\left(\pi h\left\{\sum_{i=-N_1}^{j-1}\tilde{\alpha}_i + (\tfrac{1}{2}+j)\tilde{\alpha}_j\right\} - \Theta\right), \quad (13)$$

$$n_c(\tilde{\alpha}) \triangleq \frac{1}{T}\int_{-N_1T}^{N_2T} n(t)\cos\left[\omega_0 t + \phi(t,\tilde{\alpha})\right] dt, \quad (14)$$

$$n_s(\tilde{\alpha}) \triangleq \frac{1}{T}\int_{-N_1T}^{N_2T} n(t)\sin\left[\omega_0 t + \phi(t,\tilde{\alpha})\right] dt. \quad (15)$$

The parameter β is the quality of the phase estimate coinciding with the ratio of the PLL SNR to the received SNR, and can take on non-negative real values that correspond to partially coherent detection ($0 \le \beta \le \infty$). Partially coherent detection reduces to coherent and noncoherent detection for values of $\beta = \infty$ and $\beta = 0$, respectively.

The absolute value of a complex Gaussian statistic is conditionally Rician, and the probability of one Rician exceeding another Rician, conditioned on the interference random variables, has been solved in [11] as

$$Pr\left[R(\alpha,\tilde{\alpha}_i) > R(\alpha,\alpha) \mid \Theta\right] = \tfrac{1}{2}\left[1 - Q_M\left(\sqrt{b},\sqrt{a}\right) + Q_M\left(\sqrt{a},\sqrt{b}\right)\right], \quad (16)$$

where $Q_M(\cdot,\cdot)$ is Marcum's Q-function

$$Q_M(x,y) = \int_y^\infty r \cdot e^{-(r^2+x^2)/2} \cdot I_0(x\cdot r)\, dr, \quad (17)$$

and the other terms are defined as

$$\left\{\begin{array}{c} a \\ b \end{array}\right\} = \frac{1}{2\sigma^2}\left[\frac{|M_1|^2+|M_2|^2-2\sigma_1\sigma_2 Re\{M_2 M_1^* \rho\}}{1-|\rho|^2} \mp \frac{|M_2|^2-|M_1|^2}{\sqrt{1-|\rho|^2}}\right], \quad (18)$$

with the minus sign corresponding to the parameter "a" and the plus sign corresponding to the parameter "b". The terms comprising the parameters a and b are

$$\begin{aligned} M_i &= \mathbf{E}\{z_i\},\ i=1,2, \\ \sigma^2 &= var\{z_i\} = \tfrac{1}{2}\mathbf{E}\{(z_i-M_i)^*(z_i-M_i)\},\ i=1,2, \\ \rho &= \mathbf{E}\{(z_2-M_2)^*(z_1-M_1)\}/2\sigma^2, \end{aligned} \quad (19)$$

which are the mean, variance and the normalized cross correlation coefficient, respectively; $\mathbf{E}$ is the expectation operation. For $b >> 1$, $a >> 1$, and $\sqrt{b} >> \sqrt{b}-\sqrt{a} > 0$, it has been shown in [10] that the probability of one Rician exceeding another Rician can be approximated as

$$\begin{aligned} Pr\left[R(\alpha,\tilde{\alpha}) > R(\alpha,\alpha) \mid \Theta\right] &\approx Q\left(\sqrt{b}-\sqrt{a}\right) \\ &= Q\left(\sqrt{d_e^2(\Theta)\tfrac{E}{N_0}}\right), \end{aligned} \quad (20)$$

where $d_e(\Theta) \triangleq (\sqrt{b}-\sqrt{a})/(E/N_0)$ is the normalized squared Euclidean distance, and Q(x) is the normal Q-function, $Q(x) = 1/\sqrt{2\pi}\int_x^\infty e^{-v^2/2}dv$.

Calculating the normalized cross-covariance, variance, and mean, and using suitable simplifying terms, defined below, (18) is rewritten for "a" and "b" as

$$\left\{\begin{array}{c} a \\ b \end{array}\right\} = \frac{E}{2N_0}\left[\frac{\zeta_S+\zeta_I}{\zeta_D} \mp \frac{\eta_S+\eta_I}{\sqrt{\zeta_D}}\right]. \quad (21)$$

With these terms substituted into (20) the normalized Euclidean distance expression becomes

$$d_e^2(\Theta) = \frac{(\eta_S+\eta_I)^2}{\zeta_S+\zeta_I+\sqrt{(\zeta_S+\zeta_I)^2-\zeta_D(\eta_S+\eta_I)^2}}, \quad (22)$$

where the subscript "S" corresponds to the effects of the desired signal, and the subscript "I" corresponds to the effects of the interfering signal. The definitions for the terms in (21) and (22) are

$$\begin{aligned} \zeta_I \triangleq\ & \left(\tfrac{I}{K}\right)\Big[(N_1+N_2)\left\{\Delta_I^2(\tilde{\alpha},\Theta)+\Delta_I^2(\alpha,\Theta)\right\} \\ & - 2c(\alpha,\tilde{\alpha})\left\{c_I(\tilde{\alpha},\Theta)c_I(\alpha,\Theta)+s_I(\tilde{\alpha},\Theta)s_I(\alpha,\Theta)\right\} \\ & - 2s(\alpha,\tilde{\alpha})\left\{-c_I(\tilde{\alpha},\Theta)s_I(\alpha,\Theta)+s_I(\tilde{\alpha},\Theta)s_I(\alpha,\Theta)\right\}\Big] \\ & + 2\sqrt{\tfrac{I}{K}}\Big[(N_1+N_2)\left\{c(\alpha,\tilde{\alpha})c_I(\tilde{\alpha},\Theta)+s(\alpha,\tilde{\alpha})s_I(\alpha,\Theta)\right. \\ & \left.+(N_1+N_2)c_I(\alpha,\Theta)\right\} \\ & - c(\alpha,\tilde{\alpha})\left\{c(\alpha,\tilde{\alpha})c_I(\alpha,\Theta)+(N_1+N_2)c_I(\tilde{\alpha},\Theta)\right. \\ & \left.+s(\alpha,\tilde{\alpha})s_I(\tilde{\alpha},\alpha,\Theta)\right\} \\ & - s(\alpha,\tilde{\alpha})\left\{-c(\alpha,\tilde{\alpha})s_I(\alpha,\Theta)+s(\alpha,\tilde{\alpha})c_I(\alpha,\Theta)\right. \\ & \left.+(N_1+N_2)s_I(\tilde{\alpha},\Theta)\right\}\Big] \\ & + \beta\sqrt{\tfrac{I}{K}}\Big[c_I(\tilde{\alpha},\Theta)+c_I(\alpha,\Theta) \\ & - c(\alpha,\tilde{\alpha})\left\{c_I(\tilde{\alpha},\Theta)+c_I(\alpha,\Theta)\right\} \\ & - s(\alpha,\tilde{\alpha})\left\{-s_I(\alpha,\Theta)+s_I(\tilde{\alpha},\Theta)\right\}\Big], \end{aligned} \quad (23)$$

$$\begin{aligned} \zeta_S &\triangleq (N_1+N_2+\beta)\left[(N_1+N_2)^2-\Delta^2(\alpha,\tilde{\alpha})\right] \\ &+ \tfrac{1}{2}\beta^2\left[N_1+N_2-c(\alpha,\tilde{\alpha})\right], \end{aligned} \quad (24)$$

$$\zeta_D \triangleq (N_1+N_2)^2-\Delta^2(\alpha,\tilde{\alpha}), \quad (25)$$

$$\eta_I \triangleq \left(\tfrac{I}{K}\right)\left\{\Delta_I^2(\alpha,\Theta)-\Delta_I^2(\widetilde{\alpha},\Theta)\right\}$$
$$+\ 2\sqrt{\tfrac{I}{K}}\,\{(N_1+N_2)\,c_I(\alpha,\Theta)-c(\alpha,\widetilde{\alpha})c_I(\widetilde{\alpha},\Theta)$$
$$-\ s(\alpha,\widetilde{\alpha})s_I(\widetilde{\alpha},\Theta)\}$$
$$+\ \beta\sqrt{\tfrac{I}{K}}\,\{c_I(\alpha,\Theta)-c_I(\widetilde{\alpha},\Theta)\}\,, \tag{26}$$

and

$$\eta_S \triangleq (N_1+N_2)^2-\Delta^2(\alpha,\widetilde{\alpha})+\beta\,[N_1+N_2-c(\alpha,\widetilde{\alpha})]\,. \tag{27}$$

The expressions above have been simplified by using the definitions:

$$\begin{aligned}\Delta^2(\alpha,\widetilde{\alpha}) &\triangleq c^2(\alpha,\widetilde{\alpha})+s^2(\alpha,\widetilde{\alpha}),\\ \Delta_I^2(\widetilde{\alpha},\Theta) &\triangleq c_I^2(\widetilde{\alpha},\Theta)+s_I^2(\alpha,\widetilde{\alpha},\Theta),\\ \Delta_I^2(\alpha,\Theta) &\triangleq c_I^2(\alpha,\Theta)+s_I^2(\alpha,\alpha,\Theta).\end{aligned} \tag{28}$$

In these expressions the sequence estimate is $\widetilde{\alpha}$ and the true transmitted data sequence is α.

For large SNR powers one error sequence, called a Type 1 by Aulin and Sundberg [10], dominates the probability of symbol error. Given the correct sequence

$$\alpha=\{\alpha_{-N_1},\alpha_{-N_1+1},\cdots,\alpha_{N_2-1},\alpha_{N_2}\} \tag{29}$$

and the incorrect sequence

$$\widetilde{\alpha}=\{\widetilde{\alpha}_{-N_1},\widetilde{\alpha}_{-N_1+1},\cdots,\widetilde{\alpha}_{N_2-1},\widetilde{\alpha}_{N_2}\}\,, \tag{30}$$

the Type 1 error sequence has the property that α and $\widetilde{\alpha}$ only differ in the i=0 and i=1 positions, and additionally that

$$\widetilde{\alpha}_0-\alpha_0=-\left(\widetilde{\alpha}_1-\alpha_1\right). \tag{31}$$

Other error sequences result in larger normalized square Euclidean distances and hence yield smaller error probabilities.

The resultant probability of error is summed over all possible data sequences in the union bound and averaged over the random interference phase. This can be written as

$$P_e=\sum_{i=1}^{2^{N_1+N_2}}\int_{-\pi}^{\pi}Q\left(\sqrt{d_e^2(\alpha_i,\Theta)\frac{E}{N_0}}\right)\frac{d\Theta}{2\pi}. \tag{32}$$

IV. NUMERICAL RESULTS AND DISCUSSION

The probability of error, P_e, vs. the modulation index, h, for noncoherent binary detection ($\beta=0$) is plotted in Fig. 2. It was obtained by using (32) with a 1REC phase function observed over the three symbol sequence interval [-T,2T]. The curves in this figure are indexed by the various interference-to-signal power ratios (ISR). Immediately evident is that a CW jammer has the most devastating effect on values of the modulation index, h, less than about 1.2. The larger the modulation index, the less performance loss

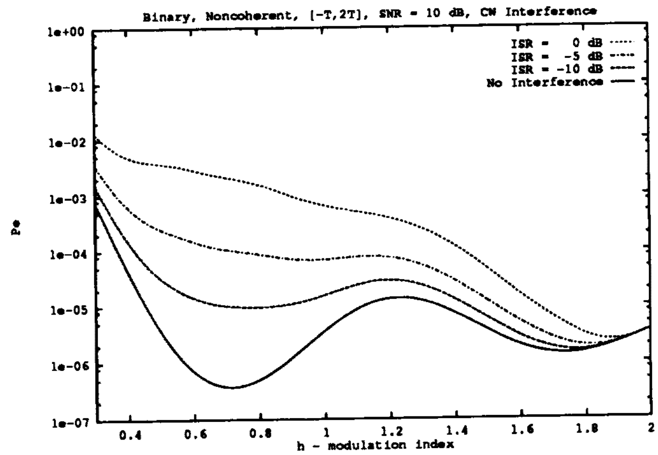

Figure 2: Probability of bit error for noncoherent detection using a 1REC phase function with different CW jamming powers.

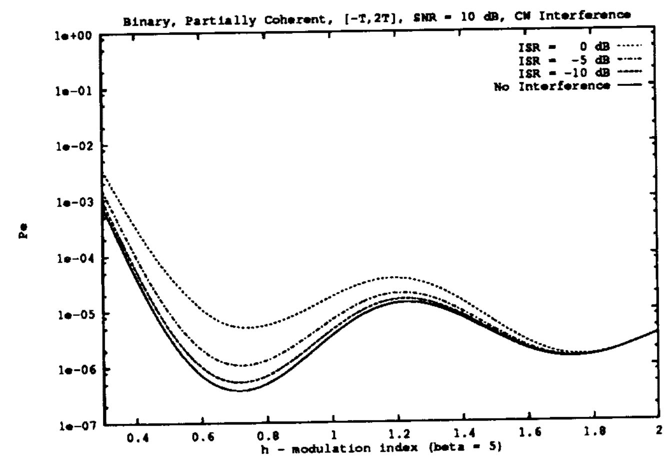

Figure 3: Probability of bit error for partially coherent detection using a 1REC phase function with different CW jamming powers.

the jammer inflicts on the noncoherent reception. The price for a larger modulation index is a larger bandwidth.

The probability of error for coherent detection is identical to the noncoherent curve without interference, for all values of ISR less than 0 dB. This says the coherent detector is unaffected by the CW jammer, provided the PLL maintains lock and has no cycle slips. The probability of error for partially coherent detection is shown in Fig. 3, with a suitably chosen coherent quality parameter of $\beta=5$, and is observed to be an improvement over the noncoherent detector performance of Fig. 2. The minimum values, or troughs, of the curves in these figures correspond to the best upper bound on the probability of error performance.

The partially coherent reception performs significantly better than noncoherent detection. While the partially coherent reception does not achieve the coherent receiver's excellent performance, it does approach it with a reasonable value of

ISR (dB)	Minimum Pe h	Pe (min.)	Degrade (dB)
Noncoherent:			
-∞	0.715	3.781×10^{-7}	0
-10	0.783	9.810×10^{-6}	1.457
-5	0.962	6.920×10^{-5}	2.943
0	1.880	2.407×10^{-6}	6.029
Part. Coh. ($\beta = 5$)			
-∞	0.715	3.781×10^{-7}	0
-10	0.717	5.369×10^{-7}	0.125
-5	0.722	1.077×10^{-6}	0.398
0	0.739	5.160×10^{-6}	1.111
Coherent:			
ISR $\leq$ 0	0.715	3.781×10^{-7}	0

Table 1: Optimum modulation index for minimum probability of error performance and degradation with SNR = 10 dB, 1REC, and observation over three symbol intervals [-T,2T], binary data.

the coherent parameter β. Notice in these figures the distinctive minimum troughs for some modulation index values. These minima correspond to the minimum probability of error for the given ISR and signal-to-noise power ratio (SNR) values. The minimum value of these troughs is the desired operating point, but some of these minima also correspond to the highest system degradation. The system degradation is defined as the amount the SNR may be reduced when the interference is removed for the same system probability of error performance. The minima for modulation index values greater than 1.0 correspond to a larger spectral bandwidth occupancy.

A summary of the modulation indices, h, that give the minimum probability of errors, P_e, and the degradation for detection over three symbols is shown in Table 1. The modulation indices for minimum Pe increase as the interference power increases. The variation of the minimum modulation index over ISR values is greatest for noncoherent detection and nonexistent for coherent detection. The variation of the minimum h for partially coherent reception is much smaller than the noncoherent detector, even with a β as small as five. The P_e performance of a partially coherent receiver is likewise close to the coherent receiver. Consequently, even better performance is expected for larger values of the coherence quality parameter β. The partially coherent receiver is then seen as a way to alleviate the effects of a CW jammer and obtain performance close to ideal coherent reception in the absence of jamming.

V. CONCLUSION

A partially coherent receiver with a binary signal constellation and an intentional CW cochannel jammer was analyzed. Using a small value of the phase coherence quality parameter, β, significant improvements over the noncoherent receiver are obtained that approach those obtained by coherent detectors with perfect carrier recovery. When the jammer is present, the optimum modulation index for the best probability of bit error performance increases as a function of the interference strength. The variation with ISR of the optimum modulation index, h, and probability of error, P_e, is much less for the partially coherent receiver than for the noncoherent receiver.

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