ON THE SELECTION OF AN OPTIMUM DESIGN POINT FOR PHASE-COHERENT RECEIVERS EMPLOYING BAND-PASS LIMITERS

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Summary. -In the design of phase-coherent receivers employing band-pass limiters, it is customary to specify system performance relative to its value at a fixed design point. For a given design point, it is well known that an optimum tradeoff can be found between the power allocated to the carrier and sideband signals. This paper describes an attempt to further improve the performance of such coherent carrier systems by optimizing the design point based upon a given practical optimization criterion. The single-channel system is treated in detail and a brief discussion is given on how to extend the optimization technique to a two-channel system.

Introduction. -Over the years a great deal of interest has been demonstrated relative to the problem of optimum power allocation in single-channel command, and one- and two-way channel telemetry coherent carrier systems for spacecraft applications (Refs. 1-4). One fact in common to all of the solutions that have been suggested is that the system design point (i. e., the carrier tracking loop signal-to-noise ratio at "threshold") has been arbitrarily fixed while the other system parameters are varied to achieve the optimum tradeoffs. To date, very little consideration has been given to the question of what is the optimum design point based upon a given practical optimization criterion.

To answer this question, the problem must be posed in such a way that the relation between an actual operating point and the design point is clearly placed in evidence, thereby resulting in a solution of significant importance to the practicing engineer. This paper suggests a method for doing so. The technique involved makes use of the up-to-date phase-locked loop and band-pass limiter theories recently contributed in Refs. 5 and 6.

Statement of the Problem. -There are two parts to the optimum power allocation problem associated with the design of a phase-coherent receiver employing a band-pass limiter. The first part (which is that most commonly considered in the literature) is to fix either the system error probability or total transmitter energy-to-noise ratio and minimize the other quantity by varying the system modulation indices. This approach can be applied to either single- or two-channel systems and implicitly assumes that the system design point is held fixed during the optimization.

The second part of the problem considers, in effect, the locus of these minima as the design point is varied, and this information is used to select an optimum design point. This part of the optimum power allocation problem has, to the author's knowledge, not been considered and is the principal motivation behind this paper. The details surrounding both parts of the problem are the subject of the next section.

System Model. -A functional diagram of a phase-coherent receiver preceded by a band-pass limiter (BPL) is illustrated in Fig. 1. The signal at

point ① can be characterized [with respect to the unit amplitude modulation $m(t)$] by*

$$y_{①}(t) = (2P)^{1/2} \sin(\omega_0 t + (\cos^{-1} m_1)m(t) + \theta_0) + 2^{1/2}[n_1(t) \cos(\omega_0 t + \theta_0) - n_2(t) \sin(\omega_0 t + \theta_0)] \quad (1)$$

In the above, m_1^2 is the modulation factor, ω_0 is the IF carrier radian frequency, θ_0 the random phase to be tracked, and P is the total transmitted power. Also, $n_1(t)$ and $n_2(t)$ are zero mean "white" gaussian noise processes with single-sided spectral density N_0 in W/Hz. After passing through the BPL, the signal can be expressed in the form (Fig. 1, point ③), **

$$y_{③}(t) = (2\tilde{\alpha}_1^2 P_1)^{1/2} \sin[\omega_0 t + (\cos^{-1} m_1)m(t) + \theta_0] + 2^{1/2}[n_A(t) \cos(\omega_0 t + \theta_0) - n_B(t) \sin(\omega_0 t + \theta_0)] \quad (2)$$

where, from an extension of the results in Ref. 6,

$$\left. \begin{aligned} n_A(t) &= \frac{2(2)^{1/2}}{\pi} \left\{ - \left[(1 - m_1^2) \tilde{\alpha}_1^2 \right]^{1/2} \right. \\ &\quad + \frac{\left[(1 - m_1^2) P \right]^{1/2} m(t) + n_1(t)}{\left(\left\{ \left[(1 - m_1^2) P \right]^{1/2} m(t) + n_1(t) \right\}^2 + \left[(m_1^2 P)^{1/2} - n_2(t) \right]^2 \right)^{1/2}} \left. \right\} \\ n_B(t) &= \frac{2(2)^{1/2}}{\pi} \left[(m_1^2 \tilde{\alpha}_1^2) \right. \\ &\quad \left. - \frac{(m_1^2 P)^{1/2} - n_2(t)}{\left(\left\{ \left[(1 - m_1^2) P \right]^{1/2} m(t) + n_1(t) \right\}^2 + \left[(m_1^2 P)^{1/2} - n_2(t) \right]^2 \right)^{1/2}} \right] \end{aligned} \right\} \quad (3)$$

and the BPL suppression factor, $\tilde{\alpha}_1$, is given by

$$\tilde{\alpha}_1 = \left(\frac{\pi}{2} \right)^{1/2} \left(\frac{\rho_i}{2} \right)^{1/2} \exp \left(-\frac{\rho_i}{2} \right) \left[I_0 \left(\frac{\rho_i}{2} \right) + I_1 \left(\frac{\rho_i}{2} \right) \right] \quad (4)$$

*We shall treat only the single-channel case for simplicity of presentation.

**We assume for the moment that the bandwidth of the IF filter is sufficient to pass the modulation $m(t)$. Later on, we consider the case where $m(t)$ is outside the bandwidth of the IF filter.

Also in the above, $P_1 = 8/\pi^2$ is the total power in the first spectral zone and the parameter ρ_i denotes the input SNR existing at point ②.

If $m(t)$ is characterized by a digital data stream $d(t)$ and bi-phase modulated onto a square-wave subcarrier $\Xi(t)$, then $y_{\textcircled{3}}(t)$ can be re-written as

$$\begin{aligned} y_{\textcircled{3}}(t) = & \left(2\tilde{\alpha}_1^2 m_1^2 P_1 \right)^{1/2} \sin(\omega_0 t + \theta_0) \\ & + \left[2\tilde{\alpha}_1^2 (1 - m_1^2) P_1 \right]^{1/2} m(t) \cos(\omega_0 t + \theta) \\ & + 2^{1/2} \left[n_A(t) \cos(\omega_0 t + \theta_0) - n_B(t) \sin(\omega_0 t + \theta_0) \right] \end{aligned} \quad (5)$$

with $m(t) = d(t)\Xi(t)$.

Defining the phase-locked loop (PLL) reference signal $r(t)$ by

$$r(t) = 2^{1/2} \cos(\omega_0 t + \hat{\theta}_0) \quad (6)$$

where $\hat{\theta}_0$ is the PLL estimate of θ_0 , then the phase detector output (neglecting second-order harmonics) is given by

$$y_{\textcircled{4}}(t) = y_{\textcircled{3}}(t)r(t) = \left(\tilde{\alpha}_1^2 m_1^2 P_1 \right)^{1/2} \sin \phi + n_A(t) \cos \phi - n_B(t) \sin \phi \quad (7)$$

with $\phi = \theta_0 - \hat{\theta}_0$ denoting the phase error process. As pointed out in Ref. 6, n_A and n_B are, in general, not gaussian processes nor do they have equal mean-square values; however, they do always have zero mean. To properly apply the Fokker-Planck technique (which is necessary for finding the phase error density), certain reasonable assumptions must be made relative to the practical operating conditions of a PLL preceded by a BPL. If the loop bandwidth W_L is designed to be small relative to the equivalent noise bandwidth W_e at the phase detector output (Fig. 1, point ④), then the component noise processes n_A and n_B are approximately independent of the phase error process ϕ . Hence, from Eq. (7), the total noise power at the phase detector output is approximately

$$\sigma_e^2 = \sigma_{n_A}^2 \cos^2 \phi + \sigma_{n_B}^2 \sin^2 \phi \quad (8)$$

where

$$\begin{aligned} \sigma_{n_A}^2 &= \frac{8}{\pi^2} \left\{ (1 - m_1^2) \left[1 - \frac{1 - \exp(-\rho_i)}{2\rho_i} - \tilde{\alpha}_1^2 \right] + m_1^2 \left[\frac{1 - \exp(-\rho_i)}{2\rho_i} \right] \right\} \\ \sigma_{n_B}^2 &= \frac{8}{\pi^2} \left\{ m_1^2 \left[1 - \frac{1 - \exp(-\rho_i)}{2\rho_i} - \tilde{\alpha}_1^2 \right] + (1 - m_1^2) \left[\frac{1 - \exp(-\rho_i)}{2\rho_i} \right] \right\} \end{aligned} \quad (9)$$

If, in addition, the input SNR ρ_i is small (the usual case of interest for a PLL preceded by a BPL), then

$$\sigma_{n_B}^2 \approx \sigma_{n_A}^2 = \sigma_e^2 \quad (10)$$

and it follows that the effective noise in the loop is approximately gaussian-distributed with one-sided flat spectrum $2\sigma_e^2/W_L$. Thus, the effective SNR in the loop bandwidth ρ_ℓ is given by

$$\rho_\ell = \frac{2m_1^2 P}{N_0 W_L} \left(\frac{1}{\Gamma_p} \right) \quad (11)$$

where Γ_p is defined in Ref. 6 by*

$$\Gamma_p = \frac{\rho_i W_i}{\rho_e W_e} = \frac{1 - \exp(-\rho_i)}{\frac{\pi(\rho_i)}{2} \exp(-\rho_i) \left[I_0\left(\frac{\rho_i}{2}\right) + I_1\left(\frac{\rho_i}{2}\right) \right]^2 \left\{ 1 + \left(\frac{4}{\Gamma_0 \pi} - 1 \right) \exp \left[-\rho_i (1 - \pi/4) \right] \right\}} \quad (12)$$

and ρ_e , the effective SNR at the phase detector output, is defined by

$$\rho_e = \frac{\tilde{\alpha}_1^2 P_1}{\sigma_e^2} = \frac{2\tilde{\alpha}_1^2 \frac{8}{\pi^2}}{N_{0e} W_e} \quad (13)$$

Specification of Loop Parameters in Terms of the System Design Point.

-In designing phase-coherent receivers preceded by band-pass limiters, it is convenient to characterize loop performance at a given operating point relative to its value at a fixed design point. In the past, it has been customary to choose this design point to correspond to a carrier SNR in the design point loop bandwidth $\Upsilon_0 = 2P_{c0}/(N_0 W_{L0})$ of 3 db, i. e., so-called "loop threshold." More recently, it has become apparent that the threshold effect exhibited by the loop is a more complicated phenomenon than can be expressed entirely in terms of a given carrier SNR. Hence, the original motivation behind choosing $\Upsilon_0 = 2$ is, strictly speaking, no longer valid. Rather, one should allow Υ_0 to be chosen as that value that results in optimum receiver performance. Proceeding then with the assumption of arbitrary Υ_0 , the loop SNR, ρ_ℓ , can be expressed as

$$\rho_\ell = x \left(\frac{1}{\Gamma_p} \right) \left(\frac{1 + r_0}{1 + r_0/\mu} \right) \quad (14)$$

where

* Γ_0 is the value of Γ_p at $\rho_i = 0$, which, for an ideal band-pass filter, is approximately equal to 1/0.862.

$$\left. \begin{aligned} x &= \frac{2m_1^2 P}{N_0 W_{L0}} = \frac{2P_c}{N_0 W_{L0}} \\ y &= \frac{W_{L0}}{W_i} \\ r_0 &= \left(\tilde{\alpha}_{10}^2 m_1^2 P_1 \right)^{1/2} K \frac{\tau_2^2}{\tau_1} \\ \mu &= \frac{\tilde{\alpha}_{10}}{\tilde{\alpha}_1} \end{aligned} \right\} \quad (15)$$

with K denoting the PLL gain and τ_1 , τ_2 the loop filter time constants. The symbol $\tilde{\alpha}_{10}$ represents the value of $\tilde{\alpha}_1$ at the design point where by definition $x = \tau_0$.

The probability density function of the phase error $p(\phi)$ is derived via the Fokker-Planck technique in Ref. 5 with the result

$$p(\phi) = N \exp [U_0(\phi)] \int_{\phi}^{\phi+2\pi} \exp [-U_0(x)] dx \quad (16)$$

where N is a normalization constant and

$$U_0(\phi) = \beta\phi + \alpha \cos \phi \quad (17)$$

with

$$\left. \begin{aligned} \beta &= \left(\frac{r_0/\mu + 1}{r_0/\mu} \right) \frac{\rho_\ell}{2W_{L0}} \left(\frac{r_0 + 1}{r_0/\mu + 1} \right) \\ &\times \left[\Omega_0 - \left(m_1^2 P \right)^{1/2} K \left(1 - \frac{\tau_2}{\tau_1} \right) \overline{\sin \phi} \right] \\ \alpha &= \left(\frac{r_0/\mu + 1}{r_0/\mu} \right) \rho_\ell - \frac{(1 - \tau_2/\tau_1)}{(r_0/\mu) \sigma_{\sin \phi}^2} \end{aligned} \right\} \quad (18)$$

The parameter Ω_0 represents the loop detuning.

Computation of Error Probability Performance. -In computing the error probability performance of the data detector, several possibilities exist depending on how the data signal is demodulated off the carrier. Specifically, the following three cases are of interest.

- (1) The input IF filter passes the total modulation which is subsequently transferred to the data detector directly from the PLL phase detector output (point ④ in Fig. 1). This situation is typical of present and past command transponders used in deep-space applications.
- (2) The input IF filter passes the total modulation; however, the modulated carrier is tapped off prior to entering the band-pass filter and is demodulated by the PLL reference in a separate phase detector (point ⑤). The output of this phase detector represents

the modulation to be recovered by the data detector. This case is typical of present-day methods for recovery of ranging data and is of interest in future command applications.

- (3) The input IF filter completely filters out the modulation, with the modulation being recovered as in case (2). This situation would correspond to the case where the data modulation is placed on a subcarrier whose frequency is outside the IF bandwidth. This case is typical of phase-coherent reception of telemetry data.

In cases (1) and (2), the BPL suppression factor, and hence the effective loop SNR, is a function of the total power-to-noise ratio in the input bandwidth. In case (3), the BPL loss is increased (relative to case (2)) because it is now only a function of the carrier component of the total power-to-noise ratio in the input bandwidth. Hence, with all other parameters unchanged, the effective loop SNR is reduced and the noisy reference loss increased. This effect produces a minor degradation in case (3) relative to case (2).

Regardless of which case is applicable, an important parameter to consider is the loop bandwidth-symbol time product. Ordinarily, for command applications $\delta_0 = 1/(W_{L0}T)$ is sufficiently small (e.g., < 0.5) such that the phase error can be assumed to vary rapidly over a symbol interval T . In many low-to-medium rate telemetry systems, this assumption still remains valid. Hence, to a good approximation, the data detector error probability, P_E , is for all three cases described by

$$P_E = \frac{1}{2} \operatorname{erfc} \left(R_d^{1/2} \overline{\cos \phi} \right) \quad (19)$$

where

$$R_d = \frac{S_d T}{N_{0d}} \quad (20)$$

$$\overline{\cos \phi} \triangleq E \{ \cos \phi \} = \frac{I_{1-j\beta}(\alpha)}{I_{-j\beta}(\alpha)}$$

The parameters S_d and N_{0d} are the data signal and noise powers, respectively, which both depend upon which of the three cases is being considered. The validity of the gaussian noise model [which is implicit in the form of Eq. (19)] is justified as follows. In cases (2) and (3), the noise affecting the data detection process is clearly gaussian because it has not been transmitted through the BPL. In case (1), the noise entering the matched filter detector (i.e., the phase detector output at point ④ in Fig. 1) is not gaussian, but, for small ρ_i , is approximately zero mean. However, because the effective bandwidth of this noise is much wider than $1/T$, the integrate and dump action of the matched filter acts as a narrow-band filter and the output statistic is approximately gaussian.

a. Case (1). If the data signal applied to the matched filter detector is taken directly from the phase detector output (i.e., point ④ in Fig. 1), then

$$S_d = (1 - m_1^2) \tilde{\alpha}_1^2 P_1 \quad (21)$$

$$N_{0d} = N_{0e} = \frac{2\sigma_e^2}{W_e}$$

An expression for N_{0e} in terms of the system design parameters can be obtained using the definition of Γ_p as in Eq. (12) with the result

$$N_{0e} = \frac{2m_1^2}{W_{L0}} \left(\frac{\tilde{\alpha}_1^2 P_1 \Gamma_p}{x} \right) \quad (22)$$

Combining Eqs. (21) and (22),

$$R_d = \left(\frac{1 - m_1^2}{m_1^2} \right) \frac{x}{2\Gamma_p \delta_0} \quad (23)$$

Also, in this case, the input SNR ρ_i , which affects the computation of $p(\phi)$, is given by

$$\rho_i = \frac{xy}{m_1^2} \quad (24)$$

b. Case (2). When the data signal is obtained by demodulating the BPL input with the PLL reference signal (i. e., point ⑤ in Fig. 1), then

$$\left. \begin{aligned} S_d &= (1 - m_1^2)P \\ N_{0d} &= N_0 \end{aligned} \right\} \quad (25)$$

and, hence,

$$R_d = \left(\frac{1 - m_1^2}{m_1^2} \right) \frac{x}{2\delta_0} \quad (26)$$

The parameter ρ_i is given by Eq. (24).

c. Case (3). For the case where the data signal is obtained as in case (2), but the input filter to the BPL completely filters out the modulation, then R_d is still given as in Eq. (26), but ρ_i now becomes

$$\rho_i = xy \quad (27)$$

Optimization of Performance as a Function of Design Point. -Our first step is to divide the power between carrier and sideband to minimize the total energy-to-noise ratio, PT/N_0 , at the transmitter for a fixed error probability, P_E , in the data channel and a fixed ratio of data rate, $1/T$, to design point loop bandwidth, W_{L0} . This procedure enables the design engineer to constrain, a priori, the minimum acceptable signal-to-noise ratio in the design point bandwidth, $x_{\min}$; this minimum is based upon tracking performance considerations and, once established, allows the engineer to seek the optimum power split subject to that constraint. If the minimum total transmitted energy-to-noise ratio occurs at a value of $x = 2P_c/N_0 W_{L0}$ less than $x_{\min}$, the system designer selects the value of PT/N_0 at $x_{\min}$. As we shall see shortly, this additional constraint determines an optimum design point.

From Eqs. (14) to (27) for a given P_E and specified values of τ_2/τ_1 , r_0 , T_0 , W_i , W_{L0} , and T , one can find an infinite set of paired coordinate

values (x , R_d). For each pair of values, the ratio of the total energy per bit to noise density may be found from*

$$\frac{PT}{N_0} = R_d' + \frac{x}{2\delta_0} \quad (28)$$

For any of the three cases being considered, a plot of PT/N_0 vs. x exhibits a minimum. As an example, consider a command application wherein the command data is obtained by demodulating the BPL input with the PLL reference signal [case (2)]. Plotted in Fig. 2 is PT/N_0 in db vs. x in db for $P_E = 10^{-5}$ and parameter values $\tau_2/\tau_1 = 0.002$, $r_0 = 2$, $\Upsilon_0 = 2$, $W_i = 9$ kHz, $W_{L0} = 18$ Hz, $1/T = 4$ bps, and $\Omega_0 = 0$. We observe that the minimum value of PT/N_0 occurs at $x = 5.8$ db which is greater than Υ_0 . If Υ_0 is now increased in value, all other parameters held fixed, one finds that the minimum PT/N_0 together with the value of x at which it occurs both continue to decrease. Thus, it appears, at least at first glance, that continued improvement in performance (in the sense of minimum PT/N_0) can be had simply by raising the design point indefinitely. Of course this cannot be true in practice, and hence some additional constraint must be placed on the problem to counteract this anomaly.

One practical consideration is that the value of x chosen for the final system design must be greater than some minimum value $x_{\min}$. Where $x_{\min}$ is related to the design point by $x_{\min} = K_0 \Upsilon_0$, and where Υ_0 increases, a point is eventually reached at which the value of x at the minimum of PT/N_0 becomes equal to $x_{\min}$. From that point on, the value of PT/N_0 at $x = x_{\min}$ is selected as the best operating condition. Figure 3 illustrates a plot of $(PT/N_0)_{\min}$ vs. Υ_0 for case (2) and the same parameters as those used in arriving at Fig. 2.** Also assumed in Fig. 3 is that $x_{\min}$ is chosen at the design point itself (i. e., $K_0 = 1$). The physical significance of such a choice is that the operating point loop bandwidth can never fall below its value at the design point, i. e., W_{L0} . One notes from Fig. 3 that an improvement of 0.41 db can be obtained by allowing the design point, Υ_0 , to be increased from its previously chosen value of 3 db (i. e., 0 db in $2B_{L0}$) to 5.8 db, the point at which the minimum value of $(PT/N_0)_{\min}$ is achieved.

Assuming case (3), i. e., the single data channel is on a subcarrier that is outside the bandwidth of the input IF amplifier, Figs. 4 and 5 illustrate the choice of an optimum design point for two sets of parameters typical of telemetry systems for $P_E = 10^{-3}$. In both Figs. 4 and 5, $\tau_1/\tau_2 = 0.002$, $r_0 = 2$, $W_i = 4$ kHz, $1/T = 8-1/3$ bps, and $\Omega_0 = 0$; in Fig. 4, $W_{L0} = 12$ Hz; in Fig. 5, $W_{L0} = 48$ Hz.

Indicated by dashed lines on Figs. 3, 4, and 5 is the asymptotic behavior of Eq. (28) as x becomes large corresponding to the case where the noisy reference due to the RF carrier becomes negligible. This asymptote satisfies the equation

$$\frac{PT}{N_0} = \left[\text{erfc}^{-1}(2P_E) \right]^2 + \frac{K_0 \Upsilon_0}{2\delta_0} \quad (29)$$

*For case (1), $R_d' = R_d \Gamma_p$; whereas for cases (2) and (3), $R_d' = R_d$. Also, we have neglected any symbol or subcarrier synchronization and demodulation losses that might be present.

**The results for case (1) where the data signal is derived directly from the phase detector output are negligibly different from those of case (2). The difference is strictly due to the effect of the BPL loss on R_d because the noisy reference loss (approximately 0.2 db) is the same for both. Hence, Fig. 3 can also be thought to apply to case (1).

Application to Two-Channel Systems. - We conclude this paper with a brief discussion on how the optimization technique presented here might be applied in a two-channel system. Of particular interest in deep-space applications is the situation where both channels are used for the transmission of telemetry information, e. g. the science and engineering channels in the Mariner Mars 1969 mission. Assuming that the two-channel telemetry model is classified as case (3), the optimization equation analogous to Eq. (28) is given by

$$\frac{PT_1}{N_0} = R_{d1} + \left(\frac{T_1}{T_2}\right)R_{d2} + \frac{x}{2\delta_{01}} + \frac{R_{d1}R_{d2}\left(\frac{T_1}{T_2}\right)}{\frac{x}{2\delta_{01}}} \quad (30)$$

where the subscripts 1 and 2 now refer to channels 1 and 2 respectively. For given error probabilities in the two channels, i. e., P_{E1} and P_{E2} , and fixed parameter values for τ_2/τ_1 , r_0 , T_1 , T_2 , W_i , W_{L0} , Ω_0 , and T_0 , a plot of PT_1/N_0 vs. x once again exhibits a minimum. If as before T_0 alone is now allowed to vary subject to the constraint $x \geq x_{\min}$, then, as in the single-channel case, there exists an optimum design point in the sense of minimum PT_1/N_0 . The practical significance of this optimum design point depends upon the specific parameter values that characterize the system. The asymptotic behavior of Eq. (30) for large x satisfies an equation similar to Eq. (29), namely,

$$\begin{aligned} \frac{PT_1}{N_0} = & \left[\operatorname{erfc}^{-1}(2P_{E1}) \right]^2 + \left(\frac{T_1}{T_2} \right) \left[\operatorname{erfc}^{-1}(2P_{E2}) \right]^2 + \frac{K_0 T_0}{2\delta_{01}} \\ & + \frac{\left[\operatorname{erfc}^{-1}(2P_{E1}) \right]^2 \left[\operatorname{erfc}^{-1}(2P_{E2}) \right]^2 \left(\frac{T_1}{T_2} \right)}{\frac{K_0 T_0}{2\delta_{01}}} \end{aligned} \quad (31)$$

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Fig. 1. A Phase-Coherent Receiver
Employing a Band-Pass Limiter

Fig. 2. Total Power-to-Noise vs. Carrier Signal-to-Noise Ratio in the Design Point Loop Bandwidth for a Fixed Design Point - Parameters are Typical of Command Application

Fig. 3. Minimum Total Power-to-Noise Ratio as a Function of Design Point - Parameters are the Same as Those in Fig. 2

Fig. 4. Minimum Total Power-to-Noise Ratio as a Function of Design Point - Parameters are Typical of Telemetry Application; Case No. 1

Fig. 5. Minimum Total Power-to-Noise Ratio as a Function of Design Point - Parameters are the Same as Those in Fig. 4; Case No. 2