EXAMINATON OF THE MECHANISM OF INTERFERENCE

BETWEEN SATELLITE NETWORKS SHARING

THE GEOSTATIONARY ORBIT

by

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FINAL REPORT

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ABSTRACT

The conventional approach to satellite orbit sharing is to determine the power level of the wanted signal, as well as the required separation between satellites on the basis of a worst case scenario of link parameters. Such an approach yields a safe but conservative system design.

An alternative design approach involves an analysis based on treatment of link parameter variations as random variables. Such variations arise because of variations in antenna pointing directions, approximations to actual antenna gain patterns, and variations in attenuation levels due to rain loss. The advantage of this alternative approach is that it avoids the practical difficulties of deciding at the outset on worst case values for the (random) variations, and avoids designs based on unlikely or even impossible events.

The report begins with the definition of a general satellite network model. The effects of small random variations of the link parameters on the carrier to interference ratio (C/I) are then determined. Actual values of C/I variations are then obtained for a worst case design approach as well as for a statistical design approach. For variations of 5% in each of 6 link parameters of the wanted and interfering signals, the two design approaches yield differences of 2.33 dB at the 98% confidence level, and 2.78 dB at the 99.9% confidence level.

The effects on the capacity of a system based on an overly conservative design are determined, and it is shown that a substantial capacity penalty can occur. Ways to better utilize existing overdesigned systems are presented.

The effects of large variations in wanted signal and interference levels resulting from rain-loss variations are examined. An algorithm is presented to determine the C/I distribution (where I also includes noise). This distribution would then be used to compare the difference in signal power determined using a worst case approach. The difference could be several dB.

The report concludes with suggestions for further work, including actual numerical calculation of the difference in C/I margins as noted above, as well as determination of improvements possible in orbit/spectrum capacity of an ensemble of satellite systems.

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I INTRODUCTION

A. Orbit Congestion and Capacity

The geostationary satellite orbit is becoming increasingly congested in some frequency bands at some orbital positions. Orbits used by fixed satellites in the 6/4 and 14/11 (or 14/12) GHz bands in the North American 70°W-140°W arc and in the Indian Ocean 0°-30°W arc are of specific interest. Congestion occurs as a result of interference between radio signals associated with different satellite networks and systems.

An increasing level of effort is being devoted to analysing the effects of such interference and to the development of design approaches to minimize its harmful effects [1-6]. For example, a recent study [1] presents some results which determine the best orbital spacing to maximize orbit capacity. It is noted that as satellite spacing decreases interference from other satellite networks increases and the capacity of a given satellite network is thereby decreased. However the capacity per unit-of-arc of orbit may actually increase, and the capacity/unit-of-arc could be optimized by properly selecting satellite spacing. The modulation method is of importance in such considerations.

B The Conventional "Worst Case" Design Approach to Orbit Sharing

The existing conventional approach to orbit sharing has grown from that used in the early days of satellite technology. Basically, degredations in carrier-to-interference ratio (C/I) for the wanted signal are determined on the basis of a worst case configuration of the various sources of C/I impairments. Reductions in C/I occur when an antenna transmitting or receiving the wanted signal is off boresight, when rainfall attenuation levels reach their maximum values or when interference levels are based on estimated worst case rather than actual antenna gains.

The worst case design approach tends to provide safe but conservative margins for C/I. However there are practical difficulties; for example how does one specify worst case values? In fact, degredations are random variables, and it is necessary to select as worst case that value which is not likely to be exceeded. What precisely is meant by not likely? Do we use 99% probability or 99.9% probability, or some other probability?

The worst case scenario may be highly unlikely. For example two independent events A and B which occur individually with 1% probability occur jointly with .01% probability. Use of worst case analysis assures performance above a desired level. However the actual performance as measured in terms of bit-error probability or signal-to-noise ratio may be well in excess of what is required. The penalty for overly-good performance is a reduction in the overall system capacity otherwise available in a given frequency range and geographic orbit.

C. A Statistical Design Approach

The purpose of the present report is to present an analysis and preliminary evaluation of the effects of system parameter variations on C/I, based on the fact that such variations are in fact random variables. We assume the variables to be statistically independent; however this assumption imposes no inherent limitation on our statistical approach.

The analysis approach developed here is particularly easy to apply when parameter variations from the nominal values are "small". Clearly, some variations including attenuation caused by variations in rainfall are not small. Large variations of this type are handled differently. In this regard we note that there is increasing interest in combatting large variations by adaptive control of transmitter power [7,9], both from ground stations on the up-link and from multi-beam satellite antennas on the downlink. Implementation of such control would tend to make all variations from nominal values small.

D. Outline of the Report

This report is organized as follows. In Section II we describe a rather general satellite network model. The model allows for interference not only from other networks but also from adjacent channels of the satellite which carries the wanted signal. The effect of small variations on C/I behaviour is determined in Section III. A simple example is presented to illustrate the power of the approach. Section IV includes further analysis of small variations in system parameters, and expresses these in terms of the product of sensitivity coefficients and changes in other system parameters. For example, the effects of the off-boresight variations are expressed in terms of the actual off-boresight angle, the broadened 3-dB angle (which broadening accounts for satellite motion) and the deviation of the actual antenna pattern from its assumed nominal value.

In Section V we use our formulation to determine actual C/I variations in terms of the number of system parameters and the standard deviations of these assuming small variations. Section VI presents an analysis of the effects on performance of overdesign based on a worst-case approach, and illustrates how a statistical approach can lead to achievable improvements in the capacity of existing systems. Both digital and analog message transmission are considered.

In Section VII we consider the effects of large variations, including attenuation due to rainfall. Concluding comments and recommendations for further work appear in Section VIII. References cited appear at the end of the report in Section IX.

II SATELLITE NETWORK MODEL

A. General Network

Fig. 1 illustrates a simplified model of a satellite network. Interfering signal I_1 is carried on the same satellite that carries the wanted signal but on a different channel, while I_2 is carried by a different satellite. As well, some fraction $\hat{I}_1$ of I_1 may also be carried by the satellite which carries I_2. Similarly, $\hat{I}_2$ may appear on the satellite which carries I_1 and the wanted signal S.

With more satellites, the interference pattern extends in an obvious way.

B. The Wanted Signal

Consider first the wanted signal on the downlink, with carrier power C (dB) received at the groundstation:

$$C = P_d + G_s(\phi) - L_m - L_f - L_{ca} - D + G_e(\theta) \tag{1}$$

where

P_d: RF carrier power of the wanted signal from the satellite transponder.

G_s: satellite antenna gain in the direction of the receiving antenna

L_m: multiplexing loss in the satellite after the power amplifier ($\approx$ 1 or 2 dB)

L_f: free space loss between satellite and earth station (including a 20 log(f) component, $\approx$ 205 dB at 12 GHz)

L_{ca}: clear air loss (fraction of a dB at SHF)

D: rain-attenuation loss (highly variable, up to $\approx$ 10 dB at 12 GHz)

G_e: earth station receiving gain in the direction of the satellite

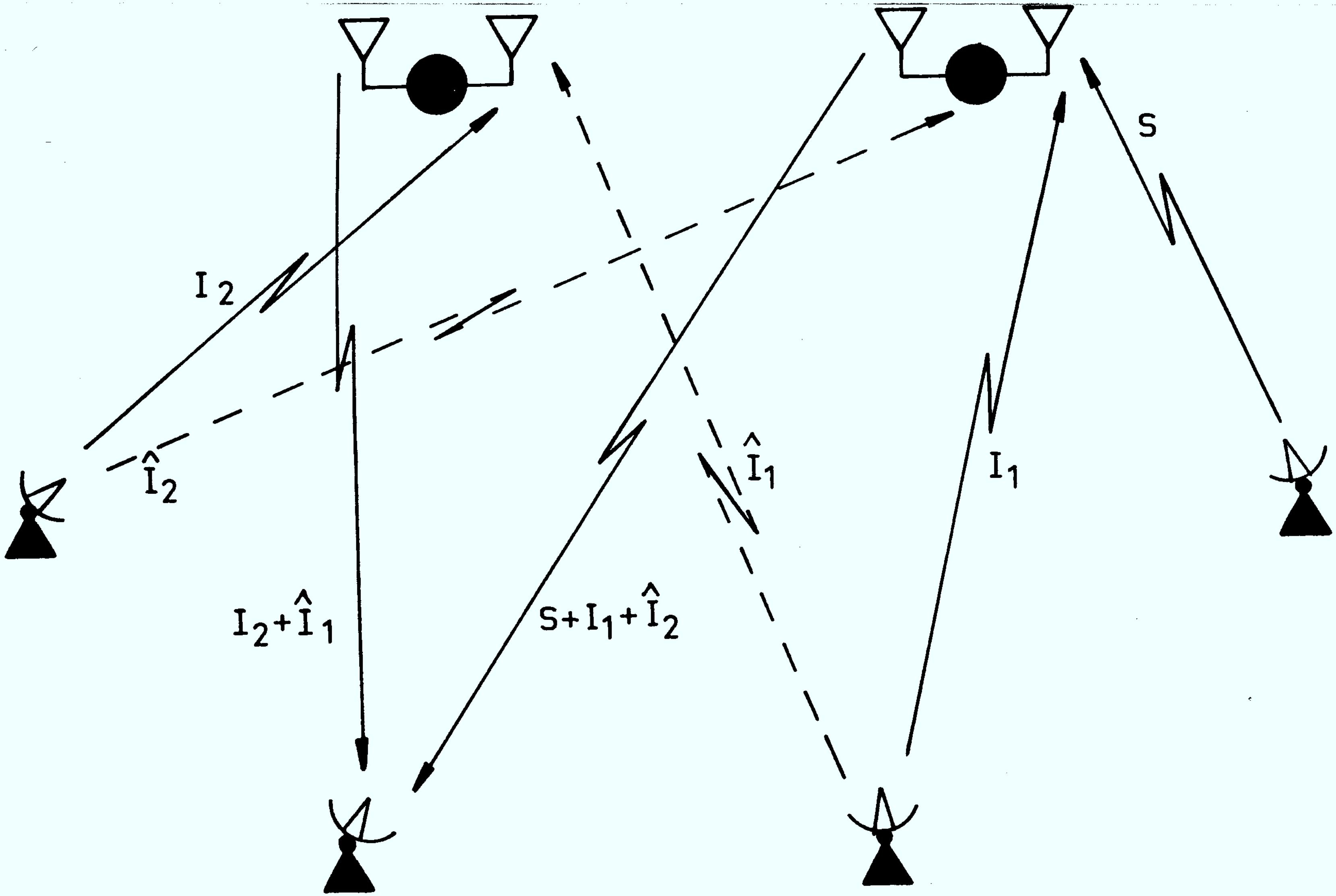

Fig. 1 Simplified satellite network model. S-wanted signal. I_k - interfering signal. $\hat{I}_k$ - secondary interference.

ϕ: angle of satellite transmitting antenna off boresight

θ: angle of earth station receiving antenna off boresight.

Combining the known factors into a single factor K_d yields

$$K_d = P - L_m - L_f - L_{ca} \quad (2)$$

$$C = K_d + G_s(\phi) - D + G_e(\theta) \quad (3)$$

A similar expression can be written for the uplink:

$$P_u = K_u + H_s(\alpha) - U + H_e(\beta) \quad (4)$$

P_u: RF carrier power of the wanted signal from the ground station transmitter

H_s: satellite receiving antenna gain in the direction of the transmitting ground station

U: rain attenuation loss on up link

H_e: earth station transmitter gain in the direction of the satellite receiving antenna

α: angle of satellite receiving antenna off boresight

β: angle of earth station transmitting antenna off boresight

Combining (3) and (4) yields

$$C = K + G_s(\phi) + G_e(\theta) - D + H_s(\alpha) + H_e(\beta) - U \quad (5)$$

where $K = K_u + K_d$.

A worst case approach is to consider the parameters in (5) as known constants, and to then consider the worst case effects of their combined variation. However we intend to consider these variations as random variables.

C. Interfering Signals

Arguments similar to those above lead to an equation similar to (5) for each received interfering carrier I_k, again expressed in dB.

$$I_k = K_k + G_{sk}(\phi_k) + G_{ek}(\phi_k) - D_k + H_{sk}(\alpha_k) + H_{ek}(\beta_k) - U_k \qquad (6)$$

The total interference in dB is obtained by first converting (6) to linear (i.e. non-dB) form in which case I_k is represented as a product of the terms in (6). The non-dB interfering carriers' powers $\{I_k\}$ are then added, and the results may then be converted again to dB form by taking $10 \log_{10} \sum_k I_k$.

In obtaining (6) we have ignored the effects of secondary interference $\hat{I}_k$. These terms could be included as well, of course, although they would be small relative to I_k.

In considering statistical variations of the various contributions, correlations between I_k and $\hat{I}_k$ would have to be included in the analysis. There is no real difficulty in doing this; however the analysis complexity increases and may obscure the major differences implied by worst-case and statistical design constraints.

III. SMALL C/I VARIATIONS

A. Linear (non-dB) Variations

To develop the C/I ratio we represent C and I as follows, in non-dB format:

$$C = \pi_k x_k \tag{7}$$

$$I = \Sigma_j I_j \tag{8}$$

$$I_j = \pi_k n_{kj} \tag{9}$$

We note here that $\{x_k\}$ and $\{n_{kj}\}$ correspond to the various variable terms in (5) and (6) respectively. Some of these variables may be functions of other variables; for example if $x_k = G_e(\phi)$ then x_k depends on ϕ.

We now consider the case where "small" variations occur in the wanted and interfering signal power levels as a result of "small" variations in the variables x_k and n_{kj}.

$$\Delta(C/I) \simeq I^{-1}\,\Delta C - (C\Delta I/I^2)$$

$$\simeq \frac{C}{I}\left[\frac{\Delta C}{C} - \frac{\Delta I}{I}\right] \tag{10}$$

$$\Delta C \simeq \sum_i (\pi_k x_k)(\Delta x_i/x_i)$$

$$\simeq C \sum_i (\Delta x_i/x_i) \tag{11}$$

$$\Delta I_j \approx I_j \sum_i (\Delta n_{ij}/n_j) \tag{12}$$

$$\Delta I = \sum_j \Delta I_j$$

$$\frac{\Delta I}{I} = \sum_j \frac{I_j}{I} \sum_i \frac{\Delta n_{ij}}{n_{ij}} \tag{13}$$

$$\frac{\Delta(C/I)}{C/I} \approx \left[\sum_i \frac{\Delta x_i}{x_i} - \sum_j \frac{I_j}{I} \sum_i \frac{\Delta n_{ij}}{n_{ij}} \right] \tag{14}$$

Eqn. (14) relates small variations $\Delta(C/I)$ relative to the nominal C/I value. One immediate consequence is that if $\frac{I_j}{I}$ is the same for all N interferers then

$$\frac{\Delta(C/I)}{C/I} \approx \sum_i \frac{\Delta x_i}{x_i} - \frac{1}{N} \left(\sum_{ij} \frac{\Delta n_{ij}}{n_{ij}} \right) \tag{15}$$

In (15), N denotes the total number of interfering signals, and $\sum_{ij}$ is over all incremental variations $\Delta n_{ij}/n_{ij}$ of all of these interfering signals.

B. <u>CIR (dB) Variations</u>

We can now determine the variations in C/I in dB. Define

$$\begin{aligned} \text{CIR} &= 10 \log_{10} (C/I) \\ &= 10 \log_{10} [(C/I)_0 + \Delta(C/I)] \\ &= 10 \log_{10} (C/I)_0 + 10 \log_{10} \left[1 + \frac{\Delta(C/I)}{(C/I)_0}\right] \end{aligned} \tag{16}$$

$$\Delta CIR = 10 \log_{10}\left[1 + \frac{\Delta(C/I)}{(C/I)_0}\right] \tag{17}$$

where $(C/I)_0$ denotes the nominal C/I (non-dB) value.

In those cases where $\Delta CIR << 1$, a further approximation is possible

$$\Delta CIR \approx (10/\ln 10) \left[\frac{\Delta\ (C/I)}{(C/I)_0}\right] \tag{18}$$

C. An Example

Since the objective of this study is to determine the feasibility of using a statistical rather than a worst case design approach, we present here a preliminary examination of the two approaches.

Consider (15) with N=5 interfering signals. Assume that each incremental term $\Delta x_i/x_i$ and $\Delta n_{ij}/n_{ij}$ has the same standard deviation σ_Δ and that these terms are statistically independent. Further, assume that

$$\sigma_\Delta = 1\% \ (\sigma_\Delta = 0.01)$$

Eqn. (5) indicates that there are six separate terms subject to variation in the wanted signal, and the corresponding eqn. (6) indicates six variables for each interference term. Thus, the standard deviation σ of the (non-dB) carrier-to-interference ratio is given by

$$\sigma^2 = 6\sigma_\Delta^{\ 2} + \left(\frac{30}{5}\right) \sigma_\Delta^{\ 2}$$

$$\sigma = \sqrt{12}\ \sigma_\Delta$$

$$= 3.46\ \sigma_\Delta$$

$$\sigma = 3.46\ \%$$

Since there are a large number of independent random variables, the central limit theorem implies that $\Delta(C/I)$ (non-dB format) is Gaussian with zero mean and variance σ^2. Thus, with 98% probability $\Delta(C/I)/(C/I)_0 < 2.33\ \sigma$ [10] which implies

$$\Delta(C/I)/(C/I)_0 < 2.33(0.0346)$$

$$< 8.06\%$$

Now consider a "worst case" scenario where all incremental parameters are at their maximum magnitudes and of such sign to maximize $\Delta(C/I)$. Immediately a problem arises; what should be assumed for worst case variation? We assume a worst case value of $2.33\ \sigma_\Delta$, since 98% of the time the parameter remains below its worst case value if the incremental variation is Gaussian (which it probably is not).

Thus, with

$$|\Delta x_i/x_i| = |\Delta n_{ij}/n_{ij}| = 2.33\%$$

$$\left|\frac{\Delta(C/I)}{(C/I)_0}\right| = 6\left|\frac{\Delta x_i}{x_i}\right| + \frac{30}{5}\left|\frac{\Delta n_{ij}}{n_{ij}}\right|$$

$$= 12\left|\frac{\Delta x_i}{x_i}\right|$$

$$= 12\ (2.33)\%$$

$$= 28.0\%$$

Thus, the two approaches give very different estimates of variability in $\Delta(C/I)$. Actually, the probability of the worst case scenario here is

$$P_{wc} = [0.01]^{36}$$

$$= 10^{-72}$$

In terms of ΔCIR we see that the worst case scenario predicts

$$|\Delta CIR|_{wc} = 10 \log_{10} [1 + 0.280]$$

$$= 1.07 \text{ dB} \qquad (\sigma_\Delta = 1.0\%)$$

If we use the statistical approach we find that with 98% probability

$$\Delta CIR_{stat} < 10 \log_{10} [1 + 0.0806]$$

$$< 0.34 \text{ dB} \qquad (\sigma_\Delta = 1.0\%)$$

Thus, if used for design purposes the two approaches would show a 0.73 dB difference in ΔCIR.

If we repeat the above analysis with a larger value of σ_Δ, say $\sigma_\Delta = 3\%$ then

$$|\Delta CIR|_{wc} = 10 \log_{10} [1 + 3(0.280)]$$

$$= 2.65 \text{ dB} \qquad (\sigma_\Delta = 3\%)$$

$$|\Delta CIR|_{stat} < 10 \log_{10} [1 + 3(0.0806)]$$

$$< 0.94 \text{ dB} \qquad (\sigma = 3\%)$$

Here the difference is approximately 1.71 dB. Under the worst case scenario $\Delta(C/I)$ is no longer "small" and the analysis is subject to some error.

This rather simple example clearly illustrates how a worst case analysis may lead to overly conservative system designs.

IV SENSITIVITY COEFFICIENTS

In order to calculate ΔCIR in a specific situation it is necessary to calculate $\Delta x_i/x_i$ and $\Delta n_{ij}/n_j$. These incremental parameters may themselves be expressed in terms of other parameters, as illustrated in the examples below.

A. Antenna Off-Boresight Variations

Consider the antenna gain factor x_i for the wanted signal, as follows:

$$x_i = f_i \left(\frac{\phi}{\phi_0}\right) G_i \left(\frac{\phi}{\phi_0}, \phi_0\right) \tag{19}$$

Eqn. (19) is explained with the help of Fig. 2 which shows both an actual antenna pattern as well as a "standard template" pattern, both plotted vs. ϕ/ϕ_0 where ϕ denotes the actual angle off boresight, and ϕ_0 is the 3-dB beamwidth of the antenna. Actually, ϕ_0 for the standard template is broadened artificially beyond the necessary ground coverage area to account for satellite movement (pitch, roll, and yaw in the case of the satellite antennae, or station keeping errors for groundstation antennae). Standard templates are normally developed in such a way that all of the peaks, or sometimes 90% of the peaks lie below the template (envelope). Such template envelopes have been developed by CCIR.

At an angle ϕ off boresight the actual gain x_i will differ from the design gain x_i at $\phi = 0$ for 3 reasons:

1. Angle $\phi \neq 0$.
2. The actual gain at angle ϕ differs from that predicted by the

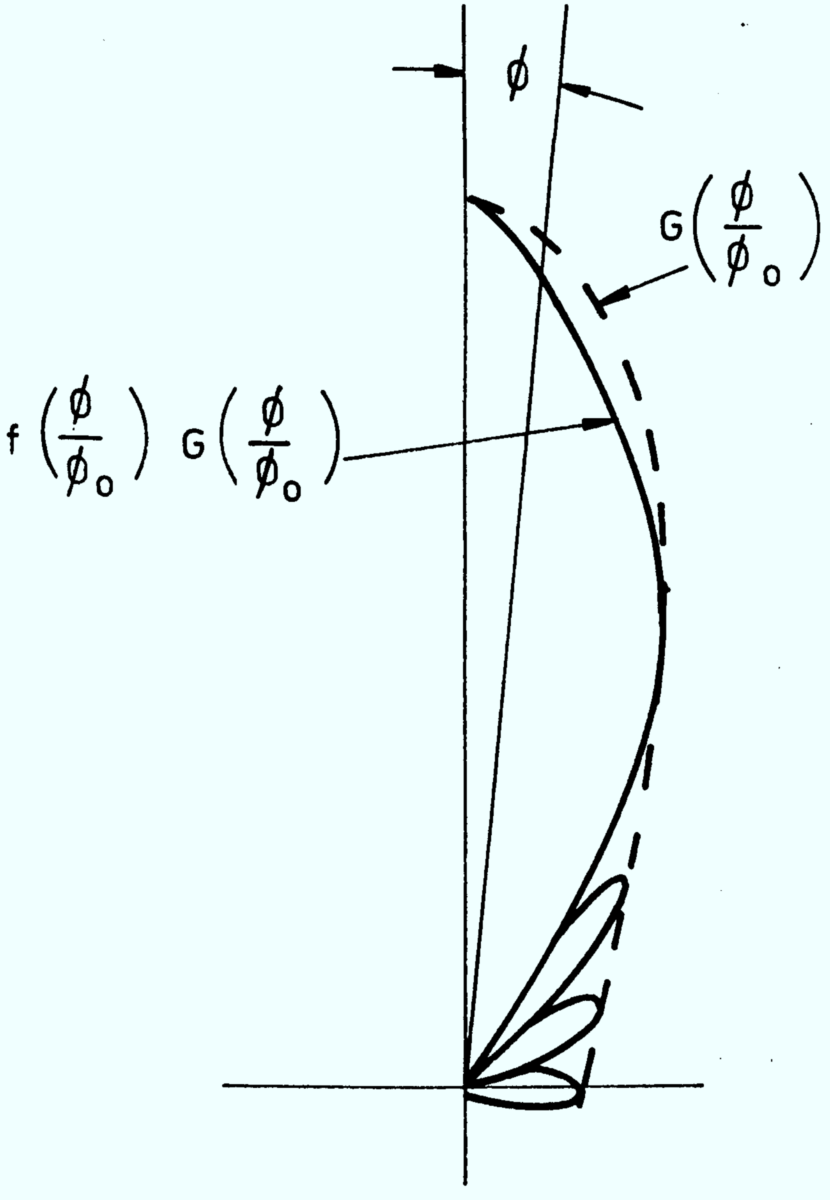

Fig. 2 Antenna Off-Boresight Variations.

standard template by the factor $f_i(\phi/\phi_0)$ in (19). (See also Fig. 2.)

3. The actual value of ϕ_0 is different from the artificially broadened value used in the standard template.

For the signal, $\phi = 0$ ideally, and we assume $\Delta\phi$ to be small. Then with $\psi = \phi/\phi_0$:

$$\frac{\Delta x_i}{x_i} \approx \frac{\Delta f(\psi)}{f(\psi)} + \frac{\Delta G(\psi, \phi_0)}{G(\psi, \phi_0)} \tag{20}$$

$$\frac{\Delta G}{G} = \left(\frac{\partial G/\partial\psi}{G/\psi}\right) \frac{\Delta\psi}{\psi} + \left(\frac{\partial G/\partial\phi_0}{G/\phi_0}\right) \frac{\Delta\phi_0}{\phi_0} \tag{21}$$

The coefficients in (20) of the form $[\partial G/\partial y)/[G/y]$ are sensitivity coefficients which multiply the incremental parameter variations $\Delta\psi/\psi$ and $\Delta\phi/\phi_0$. These sensitivity coefficients are important, since they determine the effect of the normalized incremental variations which they multiply. If the instantaneous rate of change of G is larger than the average rate of change G/y then the effect of $\Delta\psi/\psi$ or $\Delta\phi_0/\phi_0$ is amplified, and conversely.

Substitution of (20) into (15) shows that the simple linear combination of signal terms $\sum_i \Delta x_i/x_i$ is replaced by a weighted sum of the form $\sum_i \alpha_i(\Delta y_i/y_i)$ where $\{\alpha_i\}$ denote the sensitivity coefficients and $\{y_i\}$ the variables in terms of which $\{x_i\}$ have been expressed.

A similar approach can be used for the interference terms $\Delta n_{ij}/n_{ij}$, with some modifications. First, the nominal angle (ϕ/ϕ_0) would have a non-zero value. Second, the value $\Delta f/f$ may not be small, since as ϕ increases the difference between the template antenna pattern and actual pattern would tend to increase. Thus, the small variation assumption for $\Delta f/f$ may not always be accurate particularly for interference well off boresight. Having said this, we note that the effect of interfering signals at large ϕ values would be small relative to that of other interfering signals with small ϕ values. Therefore, the use of (19) and (20) would not lead to serious overall errors in $\Delta C/I$ calculations.

In those cases where $\Delta f/f$, $\Delta\psi/\psi$ and $\Delta\phi_0/\phi_0$ are small, the interference terms in (15) are of the form $\sum_i \beta_i \, \Delta z_i/z_i$ where $\{\beta_i\}$ denote the interference sensitivity coefficients and $\{\Delta z_i/z_i\}$ the incremental variables.

B. Adjacent-Channel Interference

The separation ν between adjacent-channel carriers is one of the factors which determines the adjacent-channel interference. Other factors include the modulation format and receiver filter (window) characteristic [11-14].

We denote the dependence of adjacent-channel interference on separation ν as $I_j = I_j(\nu)$. Then for variations in ν from the nominal value

$$\frac{\Delta I_j}{I_j} = \left[\frac{\delta I_j/\delta\nu}{I_j/\nu} \right] \frac{\Delta\nu}{\nu} \tag{22}$$

Given the modulation format and receiver window $I_j(\nu)$ can be calculated [10-13]. For binary PSK or OQPSK, for example, $\delta I_j/\delta\nu \sim (\Delta\nu)^{-2}$. For binary MSK $\delta I_j/\delta\nu \sim (\Delta\nu)^{-4}$ and for sinusoidal PSK $\delta I_j/\delta\nu \sim (\Delta\nu)^{-8}$. These asymptotic behaviours apply for $\Delta\nu$ above some threshold value, where the sensitivity to changes in $\Delta\nu$ is clearly greater for MSK than for PSK.

The purpose of this immediate discussion is to show that our approach to determining ΔCIR variations is quite general, and is not limited only to interference from other satellite networks.

V. SOME EXAMPLES

At this point we generalize the example presented earlier. We consider the following generalized version of (14), where we have included the effects of sensitivity coefficients α_i and β_{ij}:

$$\frac{\Delta(C/I)}{C/I} \approx \left[\sum_i \alpha_i \frac{\Delta x_i}{x_i} - \sum_j \frac{I_j}{I} \sum_i \beta_{ij} \frac{\Delta n_{ij}}{n_{ij}} \right] \tag{23}$$

When I_j/I is the same for all interfering signals, the (15) also generalizes as follows:

$$\frac{\Delta(C/I)}{C/I} \approx \left[\sum_i \alpha_i \frac{\Delta x_i}{x_i} - \frac{1}{N} \sum_{ij} \beta_{ij} \frac{\Delta n_{ij}}{n_{ij}} \right] \tag{24}$$

A. Equal Weighted Parameter Variances

We consider first the case of an arbitrary number of signal and interference variables. Let m and r denote the number of variables for the wanted signal and each interfering signal, respectively. We assume that the variance of $\Delta x_i/x_i$ and $\Delta n_{ij}/n_{ij}$ in (15) is such that the effective variances are equal to σ_Δ. Thus $\alpha_i \Delta x_i/x_i$ and $\beta_{ij}\Delta n_{ij}/n_{ij}$ all have variance σ_Δ, in which case the variance σ of $\frac{\Delta(C/I)}{C/I}$ is as follows:

$$\sigma^2 = m\sigma_\Delta{}^2 + \frac{1}{N}(Nr)\sigma_\Delta{}^2$$

$$\sigma = \sqrt{m+r}\ \sigma_\Delta \tag{25}$$

If we use a worst-case design aproach, and assume the worst case values

$$|\alpha_i \Delta x_i/x_i| = \Delta \tag{26a}$$

$$|\beta_{ij}\Delta n_{ij}/n_{ij}| = \Delta \tag{26b}$$

then

$$\left.\frac{\Delta C/I}{C/I}\right]_{wc} = m \left| \frac{\Delta x_i}{x_i} \right| + \frac{1}{N} (Nr) \left| \frac{\Delta n_{ij}}{n_{ij}} \right|$$
$$= (m+r)\Delta \tag{27}$$

Further, if we assume that the worst-case values $\Delta = 2.33\ \sigma_\Delta$, then [10]

$$\left.\frac{\Delta C/I}{C/I}\right]_{wc} = 2.33\ (m+r)\ \sigma_\Delta \tag{28}$$

We can now take the ratio of the worst-case value in (28) to 2.33 σ in (25) to obtain

$$R_{wc/stat} = (m+r)\Delta/2.33\sigma$$
$$= \frac{(m+r)(2.33)\sigma_\Delta}{2.33\sqrt{m+r}\ \sigma_\Delta}$$
$$= \sqrt{m+r}$$

We can also take the dB difference between $\Delta CIR]_{wc}$ and $\Delta CIR]_{stat}$ to obtain the following:

$$\Delta CIR]_{wc} = 10 \log_{10} (1 + (m+r)\Delta) \tag{30}$$

$$\Delta CIR]_{stat} = 10 \log_{10} (1 + \sqrt{m+r}\ 2.33\sigma_\Delta) \tag{31}$$

$$\Delta CIR]_{wc} - \Delta CIR]_{stat} = 10 \log_{10} \left[\frac{1 + (m+r)(2.33\sigma_\Delta)}{1 + \sqrt{m+r}\ (2.33\sigma_\Delta)} \right] \tag{32}$$

In those cases where ΔCIR is small in both cases, (32) simplifies to

$$\Delta CIR]_{wc} - \Delta CIR]_{stat} \simeq (10/\ln 10)[(m+r)2.33\sigma_\Delta - \sqrt{m+r}\ (2.33\sigma_\Delta)]$$
$$\simeq (10/\ln 10)(2.33\sigma_\Delta)\sqrt{m+r}(\sqrt{m+r} - 1) \tag{33}$$

Figs. 3 and 4 show (32) plotted vs m for various values of σ_Δ. In Fig. 3 r = m while in Fig. 4 r = 1.5 m. These graphs enable quick determination of the effect of using a worst case approach rather than a statistical approach when the incremental parameter variaitons are small.

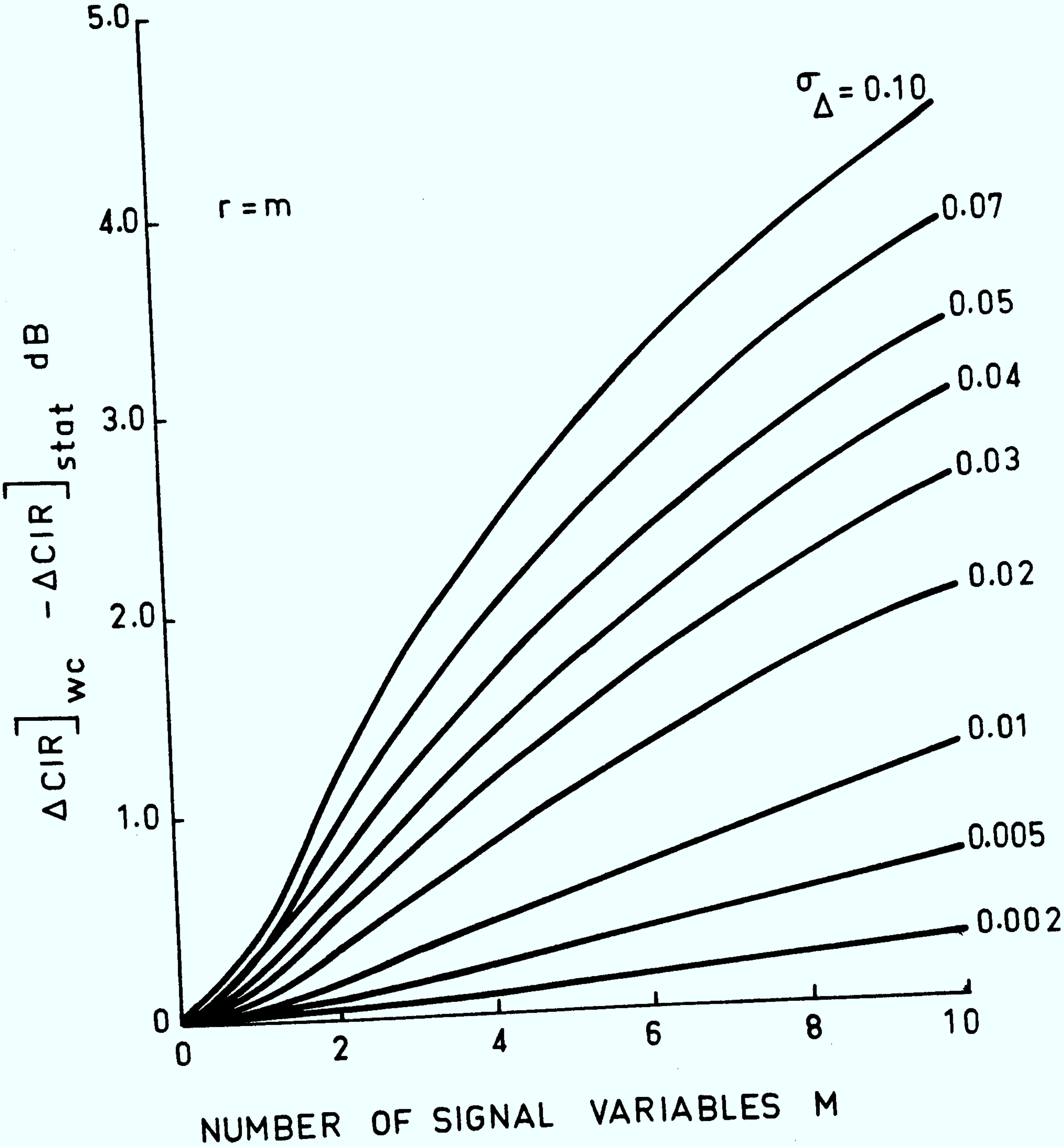

Fig. 3 ΔCIR vs σ_{Δ} and m; r = m (98% level)

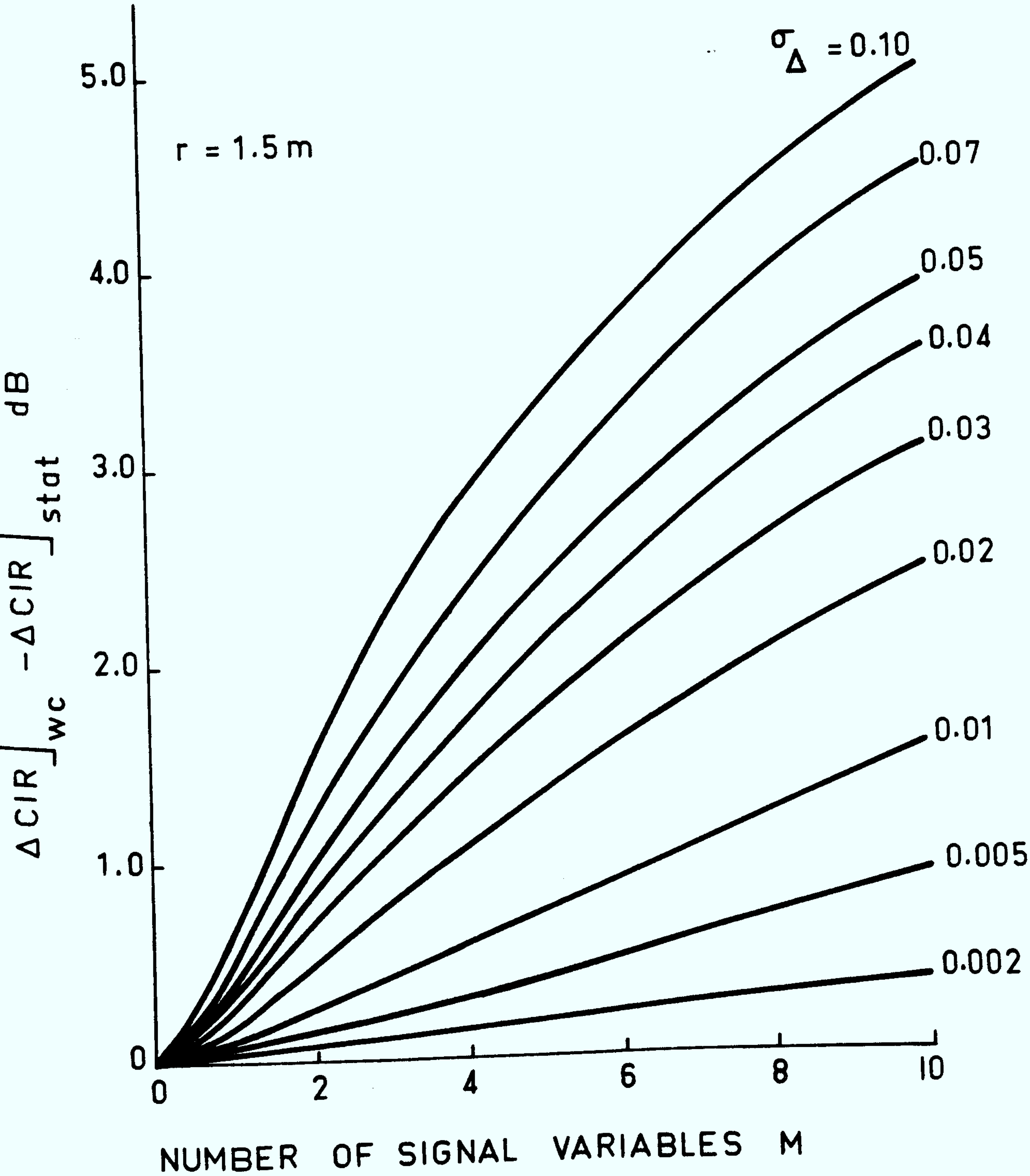

Fig. 4 ΔCIR vs σ_Δ and m; r = 1.5 m (98% level)

For example with $m = 6$, $r = 1.5(m) = 9$ and $\sigma_\Delta = 0.05$ (5%) the two approaches show a difference in ΔCIR of 2.77 dB. This is a rather substantial difference. With $\sigma_\Delta = 0.10$ (10%) the difference is even larger at 3.73 dB.

B. Linearly Tapered Variances

As a second example consider the case of linearly tapered variances of the signal and interference variations. Thus, let the variances of $\alpha_i \Delta x_i / x_i$ and $\beta_{ij} \Delta n_{ij} / n_{ij}$ be as follows:

$$\sigma_i = i\sigma_x \tag{34a}$$

$$\sigma_{ij} = i\sigma_n \qquad \text{(all j)} \tag{34b}$$

Then

$$\sum_{i=1}^{m} i\sigma_x = \sigma_x \sum_{i=1}^{m} i$$
$$= \frac{m(m+1)}{2}\sigma_x \tag{35}$$

$$\sum_{i=1}^{r} i\sigma_n = \frac{r(r+1)}{2}\sigma_n \qquad \text{(all j)} \tag{36}$$

These values can be substituted into (15) to obtain $[\Delta C/I]/[C/I]$ without difficulty.

If

$$(m+1)\sigma_x/2 = (r + 1)\sigma_n/2 = \sigma_\Delta \tag{37}$$

then (33) again results, and Figs. 3 and 4 are applicable with σ_Δ given by (37).

C. Higher Confidence Levels

Figs. 3 and 4 are applicable when a 98% confidence level is used. Fig. 5 shows the effects of using the higher confidence level of 99.9%. In this case the factor 2.33 in (32) is replaced by 3.3 [10].

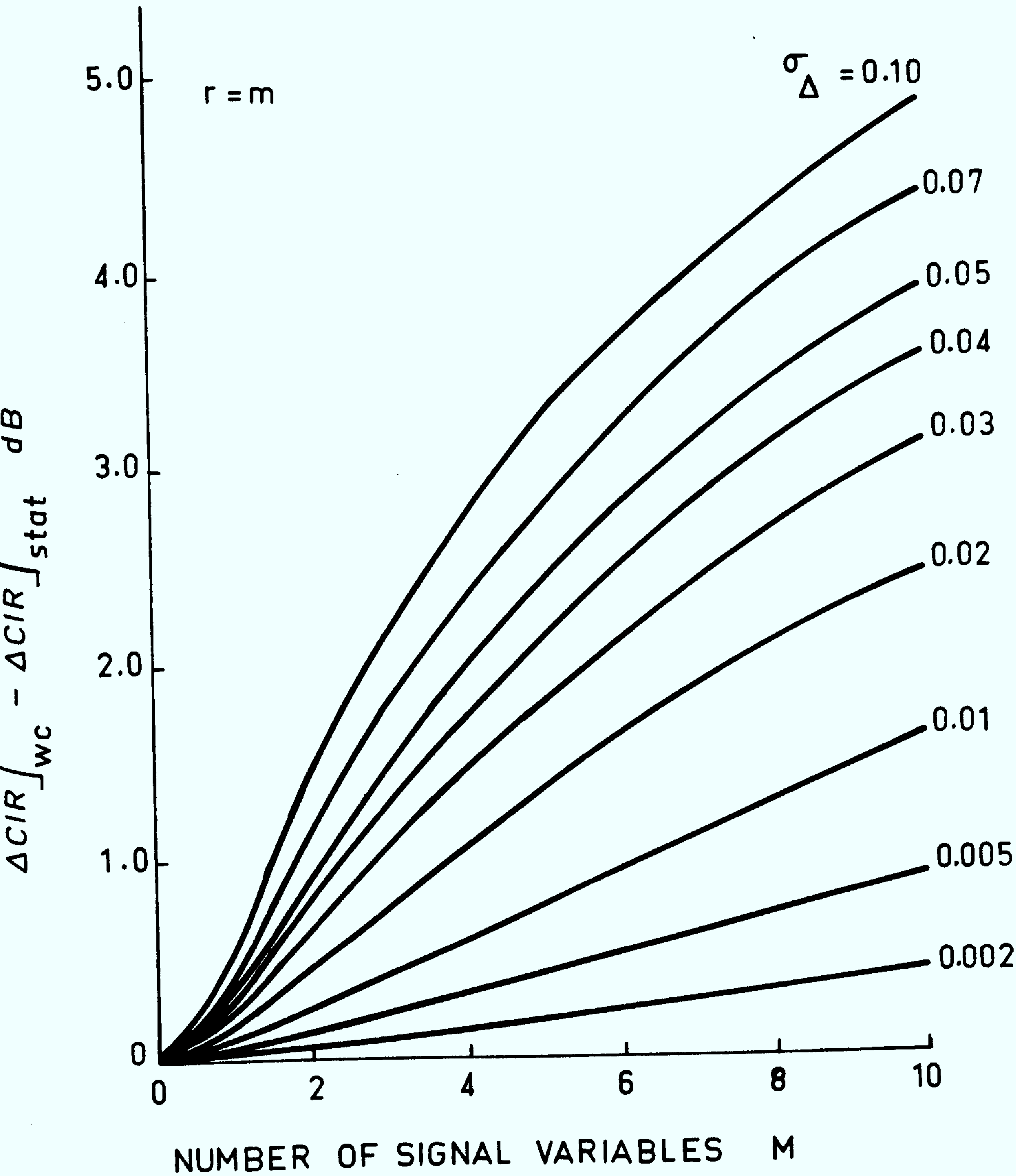

Fig. 5 ΔCIR vs σ_Δ and m; r = m (99.9% level)

Clearly the higher confidence level results in an increased difference between the two design approaches. For m = r = 6, the 98% level shows that the difference $\Delta CIR]_{wc} - \Delta CIR]_{stat} = 2.33$ dB (see Fig. 3). At the 99% level $\Delta CIR]_{wc} - \Delta CIR]_{stat} = 2.78$ dB with $\sigma_{\Delta} = 5\%$.

Similarly a lower confidence level would reduce the difference between the two ΔCIR's. However, a level below 98% would probably be unacceptable for many applications.

These examples illustrate what might be expected in terms of differences between a worst case and statistical design approach for small parameter variations. In an actual situation all variations would not be equal, and actual values of these variations would be required for substitution into (23).

VI. EFFECTS OF ΔCIR ON PERFORMANCE

We now examine the effects of an overly conservative design on system performance, and determine ways to effectively utilize the results of improved throughput implied by our statistical design approach. We do not have the flexibility here to actually change satellit orbit spacing although space shuttle technology may eventually enable such a change. Instead, we are restricted to changing modulation parameters, data transmission rates, code rates or other signal transmission variables.

To see what is possible we consider a specific example involving digital data transmission using binary PSK modulation and forward error correction (FEC) coding using a linear block code [15,16]. We assume that the channel bit rate (chip rate) is to remain constant, and that any flexibility in data throughput is via control of the FEC code rate k/n. We require a decoded message bit error probability $p_m \simeq 10^{-7}$. We assume that the system was designed using a worst case approach to yield a channel bit error probability $p = 10^{-4}$. We assume further that a statistical analysis indicates that the worst case approach is too conservative by 2.5 dB, thus

$$\Delta CIR]_{wc} - \Delta CIR]_{stat} \simeq 2.5 \text{ dB}$$

It is reasonable to assume a random error channel, and optimum matched filter detection of the PSK signal. Assuming that the totality of interference can be approximated by additive white Gaussian noise then the actual chip error rate is $p = 10^{-5}$ (see Fig. 6).

Consider FEC coding using a (n,k) block code on a random error channel to yield $p_m = 10^{-7}$. For such a code

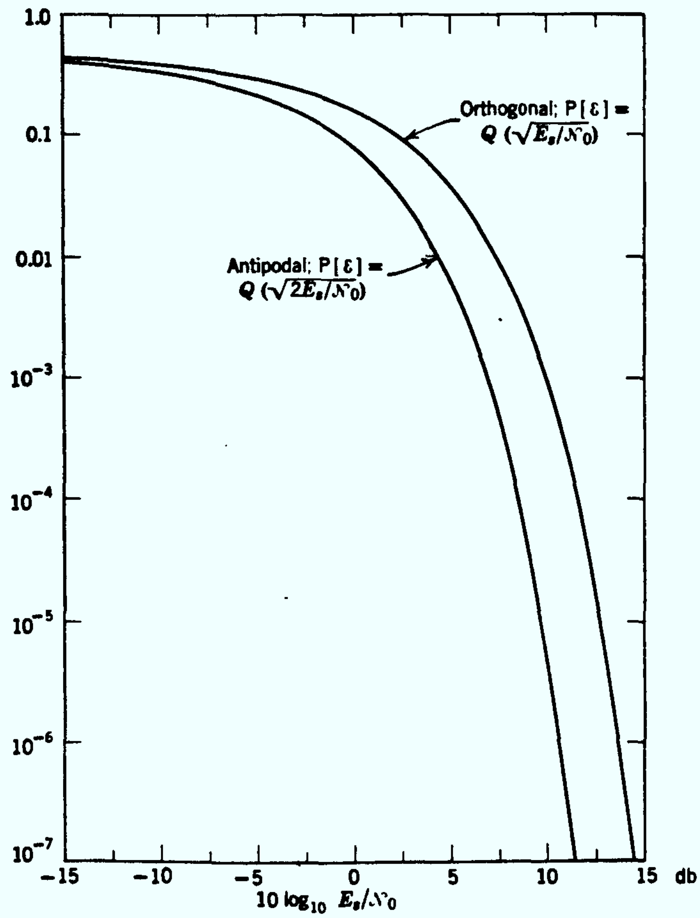

Fig. 6 Bit-error probability vs received energy per bit E_s over noise spectral density N_0 (from Ref. [17])

$$\frac{d}{n} P_e < p_m < \frac{1}{2} P_e \tag{38}$$

where d is the minimum Hamming distance between codewords, n is the block length including both information and check bits and P_e is the block error probability. In this case [15,16] with t = (d+1)/2

$$P_e = \sum_{t}^{n} \binom{n}{i} p^i (1-p)^{n-1} \tag{39}$$

$$\simeq \frac{n!}{(t+1)!\ (n-t+1)!} p^{t+1} \qquad (np \ll 1) \tag{40}$$

For np << 1 the lower bound in (38) is applicable in determining p_m.

Consider a (15,11) single-error-correcting Hamming code (d=3) for use with $p = 10^{-4}$. Then

$$p_m \simeq \frac{d}{n} \frac{[n(n-1)]}{2} p^2$$

$$\simeq 1.5\ (14)(10^{-4})^2$$

$$\simeq 2 \times 10^{-7}$$

However, if the actual value for p is 10^{-5} then $p_m \simeq 2 \times 10^{-9}$ which is two orders of magnitude better than required. With $p = 10^{-5}$ we could use a (1024, 1014) single-error correcting Hamming (d=3) code with

$$p_m \simeq 1.5\ (1024)(10^{-5})^2$$

$$\simeq 1.5 \times 10^{-7}$$

The (15,11) code has an efficiency of k/n = 11/15 = 73%, while the (1024,1014) code efficiency is 99%. The statistical system design indicates a potential efficiency increase of 36% over that implied by the worst case design. Thus, data originally sent at rate R bits/sec could actually be sent at rate 1.36 R bits/sec and still meet the error probability requirements.

In many cases there is virtually no advantage in having a decoded bit error probability much better than required for a given application. Consider the transmission of high-quality graphics or imagery, uniformly quantized to L = 10 bits per picture element. The output signal-to-noise ratio (SNR) for reconstructed images following transmission over a noisy channel and decoded with bit error probability p_m is [18]

$$\mathrm{SNR} \simeq \left[\left(\frac{1}{2^L}\right)^2 + 4p_m\right]^{-1} \qquad (41)$$

With $p_m \simeq 10^{-7}$

$$\begin{aligned}\mathrm{SNR} &\simeq (10^{-6} + 4 \times 10^{-7})^{-1} \\ &\simeq 10^6/1.4 \\ &\simeq 58.5 \text{ dB}\end{aligned}$$

Reducing p_m to 10^{-9} will cause a SNR improvement of only 1.5 dB which does not justify the code rate efficiency reduction from 99% to 73%. The only advantage of using the (15,11) code is to reduce the probability that p will rise above 10^{-4}, during some very unfavourable and very unlikely coincidence of satellite link parameter variations. At the outset, the designer must specify the acceptable probability of such an event. Once such a specification is available, a statistical design approach is appropriate.

A similar type of analysis can be undertaken for digital transmission systems where an ARQ error control strategy is used, or when analog FM transmission is employed. In the case of ARQ with an overly good channel bit error probability p, fewer check bits relative to information bits would be needed to maintain an acceptable level of retransmissions and an acceptable decoded message bit error probability. In the case of analog FM, an overly good C/I ratio would allow a reduction in the FM bandwidth expansion ratio.

In this case either more analog messages could be multiplexed on the FM carrier, or the FM bandwidth could be reduced, thereby enabling additional FM carriers to be transmitted.

It may not always be possible to increase the utliziation of a satellite network which has been overdesigned using a worst case approach. For example, consider direct digital transmission of differentially encoded speech with 4 bits of quantization. In this case [18]

$$\mathrm{SNR} \simeq \left[(1-\rho^2)2^{-2L} + 4p\right]^{-1} \tag{42}$$

With $\rho \simeq 0.9$ and $L = 4$ the quantization noise term is

$$[1 - (0.9)^2]/256 \simeq 7.5 \times 10^{-4}$$

To this is added $4p$. Clearly there is little advantage in reducing p below $\sim 10^{-4}$, since such reduction does not significantly reduce SNR in (42). If overdesign yields $p \simeq 10^{-6}$ for example, there is no easy way to increase the system efficiency. Faster transmission of the bits is one potential solution, however, this may not be feasible. For example in a single-channel-per-carrier system a higher transmission rate may result in an expanded transmission bandwidth which would lead to increased adjacent-channel interference at the very least.

Finally we note that there is an increased tendency toward spread spectrum transmission [19-22]. Here it is possible to change the data rate and hence the channel bit error probability, without altering the transmitted signal spectrum. Thus, spread spectrum systems allow for adjustment of the data rate in response to a varying channel bit error probability. In fact, variable data rate SSMA is one alternative to adaptive satellite link power control [7-9] discussed earlier.

VII LARGE VARIATIONS IN SIGNAL AND INTERFERENCE LEVELS

A. The Conventional Approach

Large variations in the level of the wanted signal and the interference can occur as result of variations in the rainfall level [7-9,23,24]. At 12 GHz attenuation levels up to 10 dB can occur. The conventional approach is to determine the rain-loss margin in dB such that the carrier-to-noise ratio C/N_t for the wanted signal is at an acceptable level for a specified percentage of the time. For voice, data and network televion this percentage is usually 99.99% for the worst month. For other services such as direct broadcasting or thin-route voice, this figure may reduce to 99.9% for the worst month. To obtain the required rain-loss margin, rain-loss data is obtained at various geographic locations at various times. Curves showing rain-loss probabilities are obtained, and transmitted power levels necessary to overcome such losses 99.9% or 99.99% of the time during the worst month in the worst service areas are thereby determined. For most of the time in most service areas, the level of the wanted signal is well above what is required for adequate service.

B. Adaptive Power Control

The potential wasting of signal power to overcome rain-loss attentuation under worst case conditions has been recognized. One approach for combatting deep fades is to control the transmitted signal power, both from the up-link groundstation and the down-link satellite. A brief but lucid explanation of the approach appears in [7] and is reproduced below:

"At SATCOM frequencies above 1 GHz, rain causes signal fades. At the frequencies most extensively used in today's systems, 4/6 and 7/8 GHz, the rain fading is moderate and can be handled by allocating a margin in transmit power of a few (4-6) dB above what is required in clear weather. Even if the downlink transmitter is the limiting resource in the system, most existing systems use a fixed power margin and accept the resulting reduction in satellite capacity. As the congestion of the frequency bands dictates the use of the higher frequencies such as the 20/30 bands, fades as deep as 15-30 dB have to be overcome. Then, it is no longer realistic to support a fixed power margin, and alternative methods to reduce communication outage due to rain fading have to be found. The use of multiple satellite ground terminals (SGT) [1], [2] has been studied. This method, called site (or path) diversity, reduces the probability of outage for two (or more) ground terminals. Reasonable separations (35 km) cause the most severe rain fades to be uncorrelated, and terrain features between the sites [2] may further diminish the harmful effects of intense rain cells. The site-diversity method applies only to static users and, because of the extra investment in ground terminals, the method seems also to be restricted to high-traffic trunks between major communication centers.

The outage can also be reduced by adaptively sharing the satellite EIRP between several users, depending on the actual fading at the various sites. In particular, frequency diversity multiple-access (FDMA) systems lend themselves to this method because the satellite effective isotropically radiated power (EIRP) for a carrier can be increased (according to some strategy to be discussed) simply by transmit power control for the earth terminals. Transmit power control has been studied [3], [4] for satellite communication systems with fixed downlink antennas. The introduction of multiple beam transmit antennas for satellites [5] offers a new tool which can be used to transfer the satellite resources between the various users according to their needs [11]. It can be used as the only means of adapting the link margins [6] or, as will be studied in this paper, in combination with control of the transmit power of the ground terminals."

Much analysis remains to be done regarding the effects on capacity of a satellite network under adaptive power control. It appears that virtually nothing has been done regarding the optimization of orbit/spectrum capacity under adaptive power control. However it is clear that power control reduces the variability of received signal and interference levels due to rainfall and perhaps other causes. In such case the small signal analysis in the previous sections may be appropriate in estimating the variability of CIR, particularly if this analysis is extended to incorporate variations which result from any limitations of power control algorithms.

C. Stastical Design Approach

We now consider a statistical design approach when large rain-loss variations accompany the other small variations considered earlier. An initial evaluation of the effecs of a worst case design approach of an interference-limited system is first illustrated below.

Consider the requirement that the C/I ratio remain above a certain level with probability $P=10^{-3}$ even when rain attenuates the wanted signal. Define probability Q as the probability that rain causes significant fading. Consider a worst case scenario with N interfering signals of equal power. Worst case conditions occur when maximum rain loss occurs for the wanted signal, with no rain loss for any of the interfering signals. The probability of this unlikely event during signal rain-loss conditions is:

$$P_1 = P(1-Q)^N \tag{30}$$

For Q = 10% and N = 10

$$P_1 = 3.48 \times 10^{-4}$$

For Q = 20% and N = 5

$$P_1 = 3.28 \times 10^{-4}$$

For Q = 20% and N = 10

$$P_1 = 1.07 \times 10^{-4}$$

Thus, with 20% rain-loss and 10 interfering signals, $P = 10^{-2}$ would yield $P_1 = 10^{-3}$. The required power of the wanted signal would be less for $P = 10^{-2}$ than for $P = 10^{-3}$. The actual difference in power would depend on the background noise level at the receiving groundstation, relative to the signal and interference power. The results in [7] suggest that the difference might be 5 dB. The actual probability can be calculated, using the method described below.

D. Statistical Analysis Algorithm for Large Variations

Of interest is $C/(N_t+I)$, where we now include noise power N_t explicitly as separate from interference. If $N_t = 0$, then (non-adaptive) scaling of the wanted and interfering signals does not change C/I. If $N_t \neq 0$, then C and I are scaled until $C/(N_t+I)$ is at an acceptable level.

We represent the wanted signal as follows, with a similar expression for each interfering signal:

$$C = xud \tag{31}$$

In (31), x includes all terms except those due to up-link (u) and down-link(d) rain loss.

Thus,

$$C/(N_t+I) = \frac{xud}{N_t + \sum_i x_i\, u_i\, d} \tag{32}$$

$$= \frac{xu}{(N_t/d) + \sum_i x_i\, u_i} \tag{33}$$

In (32) it is assumed that the wanted and interfering signals all pass through the same rain cell and all undergo the same down-link attenuation d. This assumption corresponds to the case where the groundstations of the wanted and interfering signals are close together [5]. Other cases are handled with obvious changes.

To determine the probability density of $C/(N_t+I)$, densities of sums and products of random variables must be obtained, as follows [17]:

For $z = x + y$

$$f_z(\alpha) = \int_{-\infty}^{\infty} f_{xy}(\beta,\ \alpha-\beta)\,d\beta \tag{34}$$

For $z = xy$

$$f_z(\alpha) = \int_{-\infty}^{\infty} |\beta|^{-1} f_{xy}(\alpha, \alpha/\beta)d\beta \tag{35}$$

We also require the densities for N_t, x, u, and d. The noise density would be Gaussian with zero-mean; however the variance would increase with rain loss [7] because of rain noise. An appropriate density for the small variations term x would be Gaussian, with mean equal to the mean attenuation and variance obtained using the small variation approach described earlier. The central limit theorem supports this Gaussian assumption. The rain-loss attenuation density is obtained from the log-normal approximation for the dB loss [7], and is as follows:

$$f_u(\alpha) = \left[\sqrt{2\pi}\ \sigma\ \alpha\ \ell n\alpha\right]^{-1} \exp\left[-\frac{1}{2}\ \{\ell n(10\ \log_{10}\alpha/\alpha_m)/\sigma\}^2\right]Q$$
$$+\ \delta(\alpha - 1)(1 - Q) \tag{36}$$

where α_m and σ denote the median attenuation and standard deviation in dB.

To determine the $C/(N_t+I)$ density:

1. Determine the density of xu, $x_i u_i$ and N_t/d using (35). It is first necessary to determine the density of d^{-1}, which is given by (36) with a minus sign multiplying 10 ℓn 10 in (36) [7].

2. Using (34) determine the density of $(N_t/d) + \sum_i x_i u_i$. In many situations up-link rain loss between all signals including the wanted signal and interference would be uncorrelated.

3. Determine the density of $C/(N_t+I)$ or that of $(N_t+I)/C$. The latter is probably the easier of the two to obtain.

The resulting $C/(N_t+I)$ density would depend on the variance of x and $\{x_i\}$, C/N_t and C/I in the absence of rain loss, and Q, σ and α_m in (36). The calculations would have to be done numerically, in the absence of some simplifying assumptions, and good judgement would be needed to make meaningful selection of the above parameters. The results could be plotted showing the cumulative distribution of $C/(N_t+I)$, which could then be compared directly with a similar type of distribution based on worst case analysis. The dB differences between the two approaches would indicate reasonably well the differences between a statistical and worst case design approach. Based on the earlier results for small variations and those for the "fixed transmission power" case in [7], these differences could be several dB.

VIII. CONCLUSIONS AND RECOMMENDATIONS

As a result of the work reported here we state the following conclusions and make recommendations as listed below.

A. Conclusions

1. Variations of wanted signal and interference parameters are random variables. In a worst case design approach it is necessary to assign worst case values to such variations. This assignment must eventually be based on the probability P that these parameters do not exceed some range of variability. Choosing P poses difficulties which are avoided by recognizing at the outset the random nature of these parameter variations.

2. When all parameter variations from nominal values are small, it is relavitely easy to make meaningful comparisons of worst case and statistical approaches to system design. This statement is true even if some parameters are rather complicated functions of other parameters.

3. Typical examples indicate that with parameter standard deviations of 5%, the two approaches yield C/I variations which differ by 2.33 dB. For 10% variances the difference is typically 3.22 dB. These results apply when 98% probability confidence levels are used. Higher confidence levels result in larger differences between the two design approaches.

4. The differences of ≈ 2.5 dB in C/I referred to in 2 above imply substantial differences in performance obtainable for existing satellite networks. For example, digital transmissions which use forward error correction (FEC) for error control may have an unnecessarily low

throughput for a design based on worst case parameter variations. The specific example considered earlier indicates that a throughput increase from 73% to 99% is possible for a system originally designed to operate at a bit-error probability of $p = 10^{-4}$ on the basis of worst case system parameter values. The price of a worst case design approach is an overly conservative performance at the expense of severely reduced system capacity.

5. The effects of large variations in parameter values on system design approaches have received initial consideration. Signal and interference attenuation changes resulting from variations in rainfall intensity can cause large variations in C/I. Here again it is clear that a worst case design approach can lead to an overly conservative system specification at the expense of reduced system capacity. Some recent efforts have been made to analyse the effects of using adaptive power control of either the satellite multibeam transmitter or the groundstation transmitter or both. These efforts are likely to continue and would further enhance the utility of our "small" variation analysis and results.

B. Recommendations for Future Work

1. The detailed effects of large variations in attenuation caused by changes in rainfall intensity requires additional study. Reasonably accurate probability densities of rainfall attenuation are needed for careful use in the analytical approach outline in the Section VII. It would then be possible to compare the estimates of CIR variations obtained using worst case and statistical design approaches, when

rainfall effects are included.

2. The ultimate objective is to determine the spectrum/orbit capacity obtainable, based on a statistical design approach. The resulting capacity would then be compared with that determined using a worst case design approach. The general problem of spectrum/orbit capacity determination is difficult; however some recent work has been published [1] which would assist in this effort.

3. It would be useful to examine satellite antenna characteristics, to determine the sensitivity coefficients particularly for the interference contributions where the nominal value for the off-boresight angle is non-zero.

These three recommendations all involve a substantial effort. However, the potential payoff is large, and could result in real increases in spectrum/orbit capacity of satellite systems.

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