FINAL REPORT

A DETERMINATION OF TECHNICAL CRITERIA FOR THE COORDINATION OF DIGITAL AND ANALOGUE MICROWAVE SYSTEMS

VOLUME I INTERFERENCE INTO ANALOGUE SYSTEMS

FOR: THE DEPARTMENT OF COMMUNICATIONS
JOURNAL TOWER NORTH
300 SLATER STREET
OTTAWA, ONTARIO
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Prepared by
Bell-Northern Research Ltd.
Ottawa, Ontario

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ABSTRACT

This report is Volume I of a two part study related to the co-ordination of digital and analogue microwave systems. This volume deals with digital into analogue interference while the other provides results on interference into digital systems. Mathematical models for the determination of performance degradations in FM systems are presented. Interference objectives are derived for FDM/FM and video analogue systems. The results of the study are presented in the form of C/I objective curves versus frequency separation between interfering and wanted signals. The digital interference effects are shown to be less severe than analogue at small (non-zero) frequency separations, and more severe at large frequency separations, such as at adjacent channel spacings.

Table of Contents

1. INTRODUCTION

This volume of the report deals exclusively with the effects of interference from digital microwave systems into analogue microwave systems. Both FDM/FM message modulation and video modulation on the analogue wanted-channel signal have been covered in the study. Volume II deals with the effects of interference from analogue and digital microwave systems into digital microwave systems.

A not-generally-known mathematical model for interference in FM radio systems, referred to as the baseband-addition model, is developed in Appendix A and is further discussed in Sections 3 and 4 of this report. In this computationally- efficient interference model, an equivalent FM spectrum for modelling the interference effects is obtained from an equivalent baseband spectrum which is, in turn, obtained by a simple summation of the baseband spectra of the wanted and interfering signals. Because of the method used to obtain the FM spectrum from the baseband spectrum, however, the baseband-addition model can only be used for noise-like baseband signals, and so is limited to modelling interference from FDM/FM signals into FDM/FM signals.

An alternate, widely known and accepted mathematical model for interference, the convolution model, is applicable to all types of signal modulations, and is derived in Appendix B from the baseband addition model, for the FDM/FM case. For FDM/FM into FDM/FM interference, the two models are shown to be identical. In the convolution model, the FM spectra of the wanted and interfering channel signals are used as the starting point, and the equivalent FM spectrum for modelling the interference effects is obtained by convolving the two known FM spectra.

Performance objectives for interference into FDM/FM and video FM analogue radio systems are derived in Appendix F, using the Trans Canada Guidelines as a basis. Using these objectives, then, along with the interference models, digital/analogue co-ordination criteria are generated and are presented in Appendices G through K in the form of C/I objective curves plotted versus carrier frequency separation between interfering and wanted signals.

Computer runs performed with the two interference models are found to produce numerically identical results for FDM/FM into FDM/FM interference. Also, a video into video interference run made with the convolution model was found to agree within better than 0.5 dB with the co-channel results obtained from a new in-house direct FM-simulation computer program. Agreement between measurement and computed results is excellent for both FDM/FM and digital into FDM/FM interference as outlined in Section 4.

Section 2 is the statement of work for this part of the study and Section 3 discusses the method of approach. Section 4 primarily summarizes the contents of the 12 appendices of this report in which the bulk of the information generated in this study is given. Finally, Section 5 presents observations and conclusions obtained.

2. STATEMENT OF WORK

This report addresses the objectives set out in Section 4.1.1 of the statement of work provided in BNR proposal No. P157. These objectives are given below.

Digital Into Analogue Interference

i) Validate existing computer program INTFER as modified to accommodate digital modulation spectra on the interfering radio channel and video modulation spectra on the wanted radio channel.

ii) Generate C/I curves for the continuous range of frequency shift between the centers of the wanted and unwanted RF signals (up to one normal RF channel spacing) for the following combinations of wanted (C) and interfering (I) channel loadings (Table I).

Table I Wanted and Interfering Channel Parameters to be used for deriving (C/I) curves, grouped by authorized bandwidth.

Authorized Bandwidth MHz	Wanted-Channel Modulation: Min. No. of Voice channels (NOTE 1)	Wanted-Channel Modulation: Max. No. of Voice Channels (NOTE 2)	Video	Interfering Channel Maximum Modulation Bit Rates: 4PSK	8PSK	4QPRS	8FSK
3.5	0	48	No	3(DS1)	5(DS1)	---	NOTE 3
7.0	0	120	No	7(DS1)	2(DS2)	---	NOTE 3
9.75	0	300	No	2(DS2)	3(DS2)	---	NOTE 3
20	0	1260	Yes	4(DS2)	(DS3)	---	NOTE 3
29.65	0	1800	Yes	(DS3)	10(DS2)	2(DS3)	NOTE 3
40	0	2700	No	(DS3)	2(DS3)	---	NOTE 3

Note 1 - The zero voice channel case is used to force the computer program to ignore the spectral convolution and thereby use only the digital spectrum, in order to generate the best-case (C/I) curves described in Section 3.

Note 2 - For 300 and less voice channels, the rms test tone level is 200 kHz. For more than 300 voice channels, the rms test tone level is 140 kHz. For each unmodulated case, the test tone level of the corresponding maximum modulation case is used.

Note 3 - For FSK, the frequency deviation is set equal to the Baud Rate divided by the number of FSK Levels, and then the baud rate is set to the maximum value possible without exceeding the FCC Mask. An existing BNR computer program is used to generate the FSK spectrum.

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3. METHOD OF APPROACH

To determine the characteristics of digital interference into analogue radio systems, well-established techniques for studying analogue interference have been extended to consider the digital into analogue case. Available to us from earlier work is a BNR developed computer program (INTFER). An adaptation of this program enables it to accept digital modulation spectra on the interfering radio channel and video modulation spectra on the wanted, or desired, radio channel. The program directly produces complete C/I curves versus carrier separation frequency for defined interference noise criteria. The program output has been found in the past to be in excellent agreement with measured results and with established theory for FDM/FM interference.

INTFER has been built up by BNR over a number of years into a flexible and easy-to-use software tool, capable of analyzing a variety of interference situations. An earlier version of the program was used to generate information relevant to Trans Canada Guideline TG 2.010, Issue 1, July 1975. The program has the following features:

- Interactive (question and answer) operation from a remote computer terminal (i.e., user doesn't have to remember a particular data format).
- Ease of implementation of any desired pre/de-emphasis shape.
- A variety of options are available to the user for the form and content of the program output (printed and/or plotted):

a) NPR in each "slot" across the occupied baseband, plus the total-bandwidth rms S/N after de-emphasis and weighting (the latter is useful in video simulation);

b) interference power in dBm0 in each slot across the entire baseband (including both below and above the band occupied by the FDM modulation) after de-emphasis, plus the total-bandwidth rms S/N after de-emphasis and weighting;

c) NPR (or dBm0) in the computer-determined noisiest slot, versus automatically incremented carrier separation frequency;

d) total bandwidth S/N after de-emphasis and weighting, versus automatically-incremented carrier separation frequency (for video simulation).

- The effects of filtering of the FM Spectrum are not taken into account, hence only the "spillover" part, not the intelligible crosstalk part, of adjacent channel interference can be estimated (using an average value for filter skirt attenuation).
- Applicable to moderate-to-large C/I ratios only ($\geq$ about 10 dB).

INTFER produces an FDM/FM power spectrum from a set of random (i.e., noise-like) baseband tones (representing the baseband power spectrum). The original interference model used in INTFER adds the wanted and interfering channel baseband power spectra together before obtaining the effective interference FM power spectrum. A later interference model, derived by Pontano, Fuenzalida and Chitre [IEEE Transactions on Communications, June 1973, pp. 714-745] and currently widely accepted for interference calculations, including by the FCC, obtains the effective interference power spectrum (for moderate-to-large C/I ratios) by convolving the FM power spectra of the wanted and interfering channel signals. It is shown in Appendix B by a simple mathematical analysis that the INTFER model and the convolution model are identical. The adapted version of INTFER features an option whereby, rather than generating FM power spectra from random baseband tones (simulating noise loading), the interfering and/or wanted channel FM power spectra (digital and video) are calculated directly from closed-form equations, where applicable, or are read in from disc files (i.e., they are pre-stored by any desired means, including manually). The resulting spectra are then convolved to produce the required equivalent interference power spectrum. The process of frequency domain convolution is carried out by the well-known technique of Fast-Fourier-Transforming to the time domain, multiplying, and Fast-Fourier-Transforming back to the frequency domain. All of the 'software' capabilities of INTFER for automatic C/I curve generation are thus available with the convolution method. The original method is computationally more economical for FDM/FM interference and doesn't tie up additional disc file space for spectrum storage.

Regardless of the type of digital modulation used, experience and literature searches have shown that, for interference purposes, the transmitted spectrum averaged over a random bit stream of adequate length, can be considered to have a $((\text{sine } x)/x)^2$ shape, modified by the transmitter filters. The width of the $((\text{sine } x)/x)^2$ spectrum before filtering is a function of the basic baud-rate transmitter-state switching waveform, normally a square wave. There is no unique transmitter output filter shape or width established for a given type of digital modulation, the exact shape and width being a function of the detailed design of the particular radio under consideration, which varies from manufacturer to manufacturer.

Therefore for this generalized investigation we use monotonic filters of moderate complexity with whatever widths are necessary to make each given digital spectrum meet the FCC rules for out-of-band emissions in each given analogue channel bandwidth (from the DOC SRSP's).

A video colour bar spectrum is used as a maximum extreme video modulation spectrum. This spectrum is obtained from an existing BNR computer program.

As in the case of digital radios, there is no unique established criterion for determining filter bandwidths in FDM/FM analogue radio systems, the filter design varying from one manufacturer to another. For this reason, the C/I curves presently given in the Trans Canada Guidelines for FDM/FM interference do not take any filtering into account. For large carrier frequency differences (between wanted and interfering signals), such as at adjacent channel spacings, the actual interference will be reduced both by the interfering channel transmitter filtering and by the wanted channel receiver filtering. The receiver filtering will increase the effective C/I ratio, and will decrease the amount of spectrum broadening introduced by the convolution process, while the transmitter filtering will reduce the direct "spillover" energy, that is, the energy falling inside the receiver passband. The very best that could be done would be for the receiver filtering to attenuate the bulk of the interfering signal power (i.e., so that the spectrum convolution could be ignored), leaving only the direct "spillover" interference. Thus the existing Trans Canada Guideline curves represent the worst case interference, at adjacent channel spacings, and a set of best case interference curves for adjacent channel spacings could be produced by considering only the interference spectrum as transmitted (i.e., after transmitter filtering), without the convolution with the wanted channel spectrum. This is effected by running INTFER with the wanted channel unmodulated. With no transmitter filtering, and FDM/FM-into-FDM/FM interference, the difference between the worst case and best case curves (i.e., with and without spectrum convolution) at adjacent channel spacings is generally 10 dB or greater. For the digital-into-analogue interference, then, especially since some degree of transmitter filtering will be taken into account, it is possible and fruitful to produce both worst case and best case curves.

The performance objectives for interference into analogue radio systems are given in the Trans Canada Guidelines. Interference from a suitably scrambled digital radio, especially after convolution with the analogue spectrum, can be taken to be noise-like, just as analogue interference itself is taken to be. Thus, for example, the long haul radio interference objective into FDM/FM analogue radio is taken to be no more than 4 dBrnc0 of noise in the noisiest slot per interference exposure to a digital signal. A signal-to-total-noise objective for interference into a video analogue radio is derived from the FDM/FM objectives by keeping the same ratio to the thermal noise objective. INTFER measures the total rms interference noise after de-emphasis and weighting, and then translates this to a peak-to-peak-picture-to-rms-noise ratio. The performance objectives chosen for the study are given in Appendix F.

4. Summary of Results

All of the work outlined in section 2 has been completed and is presented in this report.

Validation of the convolution model additions to the computer program has been carried out by performing FDM/FM-into-FDM/FM analogue interference runs using both the original baseband-addition model and the convolution model. The resulting agreement was found to be exact. Also, a video-into-video run was found to agree within better than 0.5 dB with the co-channel results obtained from a new in-house direct FM-simulation program, FMSIM. Finally, although funding did not permit the undertaking of experimental measurements specifically for this contract, recent laboratory measurements of interference from a particular PSK transmitter into a particular analogue receiver at and around the adjacent-channel frequency spacing, have been found to agree within about 1 dB with results computed with the convolution method. Excellent agreement between measurement and computed results has been obtained on many occasions in the past for FDM/FM-into-FDM/FM interference.

When the convolution model option is used in running INTFER (as is done for digital and video signals), the effects of filtering on the signal spectra <u>can</u> be taken into account. In fact, representatiave transmitter output filtering was employed in obtaining the video FM spectrum, as well as the digital spectra. The derivation of the convolution model from the baseband-addition model is given in Appendix B.

The methods of obtaining the filtered PSK, FSK and QPRS digital spectra are given in Appendix D. It was found that transmitter filtering was required for the FSK signals, as with the PSK signals, to meet the FCC mask requirements. As explained in Appendix E, both the worst-case (no receiver filtering) and the best-case (perfect receiver filtering) adjacent channel curves are presented.

The overall results of the study are presented in Appendices A through L. A summary of the contents of each appendix follows.

Appendix A

Section 2 of Appendix A develops the mathematical model used for FDM/FM-into-FDM/FM interference. This model has been implemented for some years now in a computer program referred to as "INTFER". The initial interference model used in INTFER is referred to as the "Baseband Addition" model. A more recent model, referred to as the "Convolution" model, and which is currently very popular, is shown in Appendix B to be identical to the baseband addition model for FDM/FM interference. Both models transform the incoming interference spectrum into an equivalent interference spectrum which appears to the wanted-channel demodulator to be doubly-modulated, i.e., modulated by both the wanted-signal and the interfering signal modulations. Both models are now implemented in INTFER.

Section 3 of Appendix A then shows the method by which the resulting wanted-channel baseband interference noise is calculated from the above-mentioned equivalent interference spectrum. Since the two interference models produce the same equivalent interference spectra, the method of baseband noise calculation is the same for both models.

Section 4 of Appendix A describes a simple method of calculating the FM power spectrum from the baseband power spectrum for noise-like baseband modulating signals. This is the method initially implemented in INTFER, and is the reason why INTFER was initially restricted to FDM/FM-into-FDM/FM interference, only. With the more-general convolution method, the starting point is the FM power spectrum, rather than the baseband power spectrum, and thus the convolution model can be used for any interference signal whose FM spectrum is known, as well as for video modulation on the wanted signal. The convolution model can also include the effects of signal spectrum filtering.

Section 5 of Appendix A describes the make-up of the initial FDM/FM version of INTFER and the options available to the user in running the program. Section 7 explains how the baseband noise values provided by INTFER are interpreted and scaled for desired interference C/I objectives, and Section 8 lists the relevant references for Appendix A.

Appendix B

As its title implies, Appendix B presents a simple derivation of the convolution model of FM interference from the baseband spectrum addition model initially implemented in INTFER, showing that the two models are one and the same thing for FDM/FM-into-FDM/FM interference (noise-like baseband signals).

Appendix C

This appendix presents some general notes on how the convolution model of FM interference was implemented in INTFER. The size of the INTFER program precludes giving a more detailed description.

Appendix D

As stated in our proposal, this study was limited to specific classes of digital signals, namely, PSK signals which can be represented by a (sine (x)/x)-squared spectrum before transmitter output filtering; 8-level FSK signals whose modulation index (k) is the reciprocal of the number of levels (i.e., 0.125); and a particular 4QPRS signal (the Northern Telecom Ltd. DRS-8 digital radio system). Appendix D describes how these digital spectra were calculated for use with the convolution model option in INTFER. Appendix D lists some of the restrictions on the PSK signals and provides a starting selection of references for digital signal spectra.

Because of a lack of any other known general guidelines, and because funding and time did not permit the derivation of specific new guidelines for each case studied, the FCC Rules and Regulations out-of-band emission "mask" was used to determine the amount (presumably minimum) of filtering to apply to the digital transmitted signals. To cut down on computation costs, theoretical, loss-less (infinite Q), Tchebychev filtering was used in the study. The Tchebychev filter equations and the FCC mask equations are also presented.

Appendix E

Tables 1 through 3 list the signal parameters used (except for the video and 4QPRS cases) for the wanted and interfering channels when running INTFER to obtain the various C/I curves given in Appendices G through K. The appendix also presents the methodology behind, and the derivation of, the signal parameters chosen for the tables. The video and the 4QPRS characteristics are given in the relevant sections of the descriptive part of the appendix.

Appendix F

This appendix presents a discussion of interference objectives and how they are implemented in INTFER.

The first section of the appendix reviews the commonly-accepted "Long-Haul" FDM/FM objective of 4 dBrnC0 worst-slot-noise (see also Appendix A, Section 7), and how this is implemented in INTFER by means of feeding in a value of -84 dB for the C/I ratio. This section also clears up a common mis-interpretation of the meaning of the expression "C/I ratio". section of appendi

The second section of the appendix demonstrates how the commonly-accepted "Short-Haul" FDM/FM interference objective of a 1 dB degradation of the fully-faded noise can be related to the above-mentioned long-haul objective for a given radio system, thus permitting the long-haul C/I objective curves to be used for short-haul systems as well.

The third section of this appendix takes three different approaches to developing an interference objective for the cases where there is video modulation on the wanted channel. The appendix then chooses a compromise objective from these, and shows how this objective is implemented in INTFER. The compromise objective chosen is a weighted peak-to-peak picture to rms noise (from the interference) ratio of 90 dB.

Appendices G Through K: $\frac{C}{I}$ Curves

The end-result of the study is the set of C/I curves found in Appendices G through K. The appendices and the curves are self-explanatory so no more will be said about them here (see Table of Contents, and the introduction to each of the appendices).

Appendix L: Power Spectra Curves

The 20 MHz Licensed Bandwidth maximum-loading post-filtering power spectra for FDM-FM, PCM, FSK and Video transmission signals are given in Appendix L.

5. OBSERVATIONS AND CONCLUSIONS

General

The two mathematical interference models used in this report are equivalent and produce results which agree with available measurements. The digital interference spectrum, regardless of the type of modulation used (within the constraints of Appendix D), can be successfully modelled by mathematically deriving a (sine x/x) -squared spectrum shape and then suitably mathematically bandpass filtering this shape. Using these models, C/I curves have been generated for digital (PSK, FSK, QPRS) interference into analogue (FDM/FM, video) radio systems. These curves are presented in Appendices G through K of this report. Examination of these curves provides the following observations and conclusions.

PSK into FDM/FM Curves

These curves are given in Appendix G, and three main conclusions can be made from the sets of three figures for each licensed bandwidth (Note that the curves of Appendix H may be used to clarify the results).

a) For a given digital signal and licensed bandwidth, the co-channel and small-frequency-offset interference effects are 5 to 10 dB more severe at maximum capacity wanted signal loadings than at minimum-capacity wanted signal loadings (See Figure H-1). This is because, at higher FDM/FM loadings (higher number of voice channels), the wanted channel baseband is being scanned out to a higher baseband frequency when searching for interference noise. The digital interfering signal spectrum being essentially flat over the wanted signal first-order sideband region, causes so-called triangular noise in the wanted channel baseband. This triangular noise increases in power with the square of the baseband frequency, thus making the maximum-capacity interference more severe than the minimum-capacity interference.

b) For a given digital signal spectrum width, the convolution process results in a broader effective digital interference spectrum width when the analogue wanted signal spectrum itself is broad (large number of FDM/FM channels) than when the analogue wanted signal spectrum is narrow (small number of FDM/FM channels). At adjacent channel spacings, for the range of modulations chosen in the study, the resulting interference severity increased 10 to 30 dB with increase in FDM/FM loading (See Figure H-1). Part of this increase is due to the fact that, as in (a) above, the wanted channel baseband with the higher FDM/FM loadings is being scanned to a higher baseband frequency (i.e. closer to the centre of the interference spectrum) than in the case of the wanted channel baseband with the lower FDM/FM loading.

c) The particular case of zero wanted channel loading has been discussed in detail in Appendix G. The curves of Appendices G and H show that the difference between the worst-case, no-receiver-filtering curves (both signals modulated) and the best-case, perfect-receiver-filtering curves (wanted channel unmodulated) at adjacent channel spacings ranges to over 20 dB. The magnitude of the difference depends on the locations of the digital spectral nulls relative to the wanted-signal first-order sidebands. Thus, accurate C/I objectives at adjacent channel spacings cannot be obtained without including specific analogue channel receiver filter characteristics, which will vary widely from manufacturer to manufacturer and with class of service.

FSK Into FDM/FM Curves

Comparing the curves in Appendix H with the maximum Baud-rate curves in Appendix G reveals little difference between similar Baud-rate PSK and FSK interference, for the FSK mod. index studied (equal to the reciprocal of the number of FSK levels). The FSK curves in Appendix H support the three observations given above for the PSK curves.

QPRS Into FDM/FM Curves

The single QPRS case studied supports the observations made above for PSK interference. Because the QPRS spectral bandwidth, for the case studied, was very broad compared with the FDM/FM bandwidth being co-ordinated with it, there was less difference between the worst-case and best-case interference curves (ie., with and without spectral convolution) than in the PSK and FSK cases studied.

Digital Into Video

These curves are given in Appendix J. Only the worst-case (both signals modulated) interference is given. The resulting curves are seen to be similar in shape to the digital-into-FDM/FM curves, but are considerably more severe for the video performance objectives used in the study.

Comparison Of PSK and FDM/FM Into FDM/FM

Filtering determined solely by FCC out-of-band emission constraints results in digital signals with interference effects at adjacent channel spacings that are in the order of 20 Fig K2 dB more severe than the corresponding-capacity unfiltered FDM/FM analogue signal interference effects. The broad digital spectrum produces co-channel interference that is 5 to 10 dB more severe than FDM/FM analogue interference. Due to the strong carrier component of FDM/FM analogue signals, however, the FDM/FM interference effects in the range between adjacent and co-channel separation are in the order of 20 dB more severe than the digital interference effects (although only over a narrow region of an FDM/FM wanted channel baseband). These relationships are shown graphically in Figure K-1.

Power Spectra

A comparison of the RF Power Spectra of FDM/FM, PSK, FSK and video signals is given in Figure L-1 of Appendix L, for the 20 MHz licensed bandwidth case. It should be noted that the FDM/FM spectrum is not filtered at all, whereas the PSK and FSK spectra are severely filtered in order to meet the FCC out-of-band emission mask. The video colour bar spectrum reflects normal analogue transmitter output filtering. 4GHz RA-3 T4 ? 5 or 6 pc

It is seen that the PSK, FSK and video spectra are quite comparable over the 20 MHz bandwidth displayed in Figure L-1 and that the FDM/FM spectrum is comparatively less severe except in the region of the strong FDM/FM carrier component.

APPENDIX A

COMPUTER SIMULATION OF INTERFERENCE EFFECTS

IN FDM-FM SYSTEMS

TABLE OF CONTENTS

1.0 INTRODUCTION

This appendix gives a description of the mechanism responsible for co-channel and off-channel interference in FDM/FM analog message systems. A computer program, INTFER, is outlined which simulates this interference phenomenon so that theoretical C/I objectives as a function of frequency separation can be calculated for various wanted and interfering channel loadings. The effects of emphasis are included in the program. The appendix documents the background material used in developing the interference model, and shows how the interference model is implemented in the computer program. The required user inputs are described as well as the options available when simulating a given situation.

The interference model derived by Godier[1], and described in Section 2 of this report, was chosen for this computer simulation because of its ease of implementation (along with the FM Spectrum calculation method of Stewart[2]) with the availability of the Fast Fourier Transform, and because of the clear insight it afforded into the general FM interference problem. It allowed the bulk of the programming effort to be directed towards automating the calculations and in general simplifying the task of (interactively) running the program for a variety of interference situations (see options in Section 6). The computed results agree well with both measurement and theory, and are economical to obtain. The model imposes no restrictions on modulation index, and any desired pre-emphasis shape is easily implemented, as the program calculates the FM spectrum shape 'on-the-fly' from the baseband information. Interference power lying both inside and outside the occupied regions of the baseband (eg, in a so-called "noise-slot") is calculated by the program, the results being both printed and plotted.

The model is, however, limited to moderate-to-large C/I ratios, and also to random-noise basebands, capable of being simulated by equally-spaced, random-phase, sinusoidal baseband tones. Moreover, the effects of filtering of the FM Spectrum are not taken into account, and hence only the "spillover" part, not the "intelligible cross-talk" part, of adjacent channel interference can be calculated (by increasing the C/I by the estimated average value of filter-skirt attenuation).

2. FDM-FM INTERFERENCE MODEL

If the interfering carrier is unmodulated the overall expression for the combined carriers (wanted and interfering) is given by:

$$U(t) = A(t)\ \mathrm{COS}\ (W_c t + U_w(t) + U_d(t)) \qquad (1)$$

where Wc – is the wanted carrier frequency in radians/sec,

$U_w(t)$ – is the wanted phase modulation,

$U_d(t)$ – is the instantaneous phase error caused by the interfering carrier,

and $$A(t) = (1 + r^2 + 2r\ \mathrm{COS}\ (W_d t - U_w(t)))^{1/2} \qquad (2)$$

Godier [1] has shown that:

$$U_d(t) = \mathrm{TAN}^{-1}\ \left(\frac{r\ \mathrm{SIN}\ (W_d t - U_w(t))}{1 + r\ \mathrm{COS}\ (W_d t - U_w(t))}\right) \qquad (3)$$

where r – is the voltage ratio of the interfering and wanted signals,

W_d – is the frequency difference between the wanted and interfering carriers.

If we assume that $r << 1$ and that a perfect limiter is used, then $A(t)$ will become a constant, A, and $Ud(t)$ can be rewritten as:

$$U_d(t) = \mathrm{TAN}^{-1}\ (r\ \mathrm{SIN}\ (W_d t - U_w(t))) \qquad (4)$$

Since $\mathrm{TAN}^{-1}\ x = x$ for $x << 1$, (first order approximation) then (4) can be written as:

$$U_d(t) = r\ \mathrm{SIN}\ (W_d t - U_w(t)) \qquad (5)$$

Therefore by taking into account the above assumptions, equation (1) can be rewritten as:

$$U(t) = A\ \mathrm{COS}\ (W_d t + U_w(t) + r\ \mathrm{SIN}\ (W_d t - U_w(t))) \qquad (6)$$

where $U_w(t) + r\ \mathrm{SIN}\ (W_d t - U_w(t))$ is the total phase modulation on the carrier.

It is important to note that the expressions we have derived apply to the output of a perfect phase demodulator, and would have to be differentiated to apply to the output of a frequency demodulator. This process of differentiation, however, does not change the relative amplitudes of signal and distortion components at a given baseband frequency. Thus, one can write the distortion to signal ratio over narrow frequency intervals in the baseband as:

$$D/S = \frac{r \text{ SIN } (W_d t - U_W(t))}{U_W(t)} \tag{7}$$

Equation (7) can be expressed in dB's as follows:

$$D/S \text{ (dB)} = 20 \text{ LOG10 } \left(\frac{r \text{ SIN } (W_d t - U_W(t))}{U_W(t)}\right) \text{ dB} \tag{8}$$

One can see that the distortion term will increase directly with r (the voltage ratio of the interfering and wanted carriers).

This distortion term, r SIN $(W_d t - U_w(t))$, adds a carrier in the desired baseband at a frequency of W_d (the difference between the wanted carrier frequency and the interfering carrier frequency), modulated by the desired modulation $U_w(t)$. Therefore the distortion in the desired baseband equals the contributions of this modulated carrier that lie within the desired baseband frequency limits. Figure 1 shows the relationship at the output of a phase demodulator (for convenience, the wanted channel baseband spectrum is shown as being flat after pre-emphasis and integration).

Godier [1] shows that, at any point in the baseband, the distortion is made up of two terms. If f_p is the frequency at which one wishes to calculate the total distortion due to the interference, there is a distortion contribution at both f_p and $-f_p$ because of the zero frequency folding which is shown as a dashed line in Fig. 1. By adding the two distortion terms across the baseband one can obtain the overall baseband spectrum.

One finds that the shape of an FM power spectrum is independent of the carrier frequency. Therefore the spectrum of the unwanted carrier modulated by $U_w(t)$ will be identical to the wanted modulated carrier spectrum. Once this spectrum is calculated one can add it, centered at the carrier difference frequency f_d, to the wanted baseband spectrum, (at the point of a phase demodulator).

The above section has dealt with the effect of an interfering unmodulated carrier on the baseband spectrum. Godier [1] also shows that the interfering carrier is modulated, equation (5) becomes:

$$U_d(t) = r \text{ SIN } (W_d t + U_{uw}(t) - U_w(t)) \tag{9}$$

where $U_{uw}(t)$ is the modulation on the unwanted carrier,

and $U_w(t)$ is the modulation on the wanted carrier

The spectrum of the modulated (unwanted) carrier at the output of the wanted signal phase modulator is no longer the same as that of the wanted modulated signal. Now one has to calculate the spectrum of an interfering carrier modulated by both the wanted modulation ($U_w(t)$) and the unwanted modulation ($U_{uw}(t)$) (See Section 4). Once this is accomplished the distortion can be calculated as outlined above.

No reference has been made to the effects of practical pre-emphasis networks on the distortion calculations. The pre-emphasis will weight the baseband spectrum according to an easily calculated frequency response. Thus once the baseband spectrum has been so modified, the calculations can be carried out in the prescribed manner.

3. PROCEDURE FOR CALCULATING NOISE

Figure 2 is a general block diagram of an FM system, as implemented in the INTFER program[5].

The baseband is simulated in the program by a set of constant amplitude tones, which are equally spaced across the desired baseband and have random phase. Therefore the input to the system approximates a white noise source. The pre-emphasis weights the baseband spectral density according to the type of emphasis networks used (CCIR, TD-2 etc). The pre-emphasis network is followed by an amplifier with gain 'A', such that the desired RMS frequency deviation is obtained after frequency modulation. The next two blocks form an FM modulator. First the signal is integrated, and then this integrated version is applied to a phase modulator. The FM modulated signal is then fed to the transmitter section. The input to the receiver is made up of two components, the desired signal (C) and the interfering signal (I), where the interfering carrier may or may not be modulated. The FM receiver demodulates the composite signal, amplifies it by a factor 1/A and weights it using the corresponding de-emphasis network. Ideally, if one assumes there is no interference in the system the signal at point M will equal that at point A. When interference is introduced into the model it will result in unwanted noise in the wanted baseband as described in the previous section. If a 0 dBm0 test tone signal at a frequency f_b is applied to point A, the signal levels will be given by:

$$A = M = 0 \text{ dBm0}$$

$$B = L = 0 \text{ dBm0} + 20 \text{ LOG}_{10} P(f_b)$$

$$D = K = 20 \text{ LOG}_{10} (P(f_b) \times A)$$

$$E = J = 20 \text{ LOG}_{10}\left(\frac{P(f_b) \times A}{f_b}\right)$$

$$F = H = \frac{P(f_b) \times A}{f_b} = \text{RMS deviation in radians.}$$

The following steps outline the procedures for calculating the D/S ratios across the wanted baseband[5].

Calculation of the distortion power (D)

A) Calculate the FM spectrum of the carrier modulated by the combined basebands (wanted and interfering).

B) Express the above power spectral density in dB below the total power in the spectrum, using the desired slot width. For this analysis the slot width is 3.1 KHz (voice channel).

C) Now sum the spectral components at the frequency of interest ($f_d \pm f_{slot}$) as shown in Fig. 1. The two values are added on a power basis.

D) The distortion in the slot is then equal to:

$$D(dBm) = D' - C/I - 3 \text{ dBm}$$

Where D' is the sum of the two components in C. Since the spectral density is calculated with respect to 1 watt ($\sqrt{2}$ SIN ($W_d t + U_w t$)), rather than 1 volt (SIN ($W_d t + U_w t$)), the calculated distortion value D must be reduced by 3 dB.

Calculation of the wanted signal power (C)

A) Calculate the wanted signal loading power in the slot under consideration.

This is equal to $BBLP - 10 \text{ LOG}_{10} \left(\frac{Fmax - Fmin}{3100}\right)$ signal loading per channel (10)

Where Fmax and Fmin are the upper and lower frequencies, respectively, of the desired baseband in Hz, 3100 is the slot bandwidth in hertz and BBLP is the total wanted signal baseband loading power, given by:

$$\begin{aligned} BBLP &= -15 + 10 \text{ LOG}_{10} N && \text{for } N \geq 240 \\ BBLP &= -1 + 4 \text{ LOG}_{10} N && \text{for } 60 \leq N < 240 \\ BBLP &= 2.6 + 2 \text{ LOG}_{10} N && \text{for } 12 \leq N < 60 \end{aligned} \qquad (11)$$

where N is the number of loading (message) channels.

B) Now the wanted signal power is calculated at point J in Fig. 2 by adding the term,

$$20\ \mathrm{LOG}_{10}\left(\frac{P(f_{slot}) \times A}{f_{slot}}\right)$$

to the value in part A, above.

C) Therefore the overall wanted signal power, S, will be given by:

$$S = \mathrm{BBLP} - 10\ \mathrm{LOG}_{10}\left(\frac{\mathrm{Fmax} - \mathrm{Fmin}}{3100}\right) + 20\ \mathrm{LOG}_{10}\left(\frac{P(\mathrm{slot}) \times A}{\mathrm{Fslot}}\right) \qquad (12)$$

Now the D/S ratio (in dB) will be equal to the distortion power (D) minus the signal power (S). The noise power ratio (NPR (dB)) in the slot is simply - D/S (dB).

In order to calculate the absolute interference power in the FM baseband slot one must add the term:

$$20\ \mathrm{LOG}_{10}\left(\frac{f_{slot} \times d\ (f_{slot})}{A}\right) \qquad (13)$$

This will give the normalized interfering power in dBm0 at point M in Figure 2, at a $\frac{C}{I}$ of 0 dB.

4. FM SPECTRUM CALCULATION

In order to estimate the FM spectrum one may first calculate the autocorrelation function of the modulated carrier, $R(\tau)$. Then using the Wiener Khintchine Theorem, the power spectrum can be calculated by taking the fourier transform of the autocorrelation function, as follows,

$$G(f) = [R(\tau)] = \int_{\infty}^{\infty} R(\tau)\ e^{-jw\tau} d\tau \qquad (14)$$

where $G(f)$ is the desired power spectrum and $R(\tau)$ can be defined as follows:

$$R(\tau) = \lim_{T\to\infty} \frac{1}{T} \int_{-T}^{T} x(t)\ x\ (t + \tau)\ dt \qquad (15)$$

Let an FM or PM signal be given by:

$$V(t) = A\ e^{j\ (Wct + \Psi(t))} \qquad (16)$$

where A is the peak amplitude, W_c is the carrier frequency in radians/sec and $\Psi(t)$ is the phase, which is a function of the modulating signal. For PM, $\Psi(t)$ is proportional to the modulating voltage while for FM, (t) is proportional to the integral of the modulating signal, therefore,

$$\begin{aligned} &\Psi(t)\ \alpha\ Vm(t) && PM \\ \text{and}\quad &\Psi(t)\ \alpha \int_{T} Vm(t)\ dt && FM \end{aligned} \qquad (17)$$

where $V_m(t)$ is the modulating signal, itself.

The autocorrelation function of the modulated carrier, $R(\tau)$, is given by Stewart [2] as:

$$Rc(\tau) = (A^2/2\ COS\ Wc\tau\ EXP\ (\overline{\Psi(t)\quad \Psi(t+\tau}) - \overline{\Psi(t)^2}) \qquad (18)$$

where $\overline{\Psi(t)\ \Psi(t+\tau)} = R_B(\tau)$ = Autocorrelation of the baseband signal

and $\overline{\Psi(t)}^2 = R_B(0)$

Therefore equation (18) can be rewritten as:

$$Rc(\tau) = (A^2/2 \cos W c\tau \quad EXP\ (R_B(\tau) - R_B(0)) \tag{19}$$

Equation (19) applies for both FM and PM signals when the carrier frequency is large compared with the baseband frequency. ?

Equation (19) can be further simplified if we let the carrier frequency, Wc = 0. This can be done since the spectral density shape is independent of the carrier frequency. One can also normalize the constant A2/2 to unity (1 watt) without changing the meaning of equation (19). In lieu of the above simplifications equation (19) can be expressed as:

$$Rc(\tau) = EXP\ (R_B(\tau) - R_B(0)) \tag{20}$$

Therefore $Rc(\tau)$ can be evaluated once the autocorrelation function of the baseband ($R_B(\tau)$) is determined.

The baseband can be represented by white noise across the band. One can simulate this white noise by a set of equally spaced sinusoids over the desired frequency band. The magnitudes of these tones should be equal across the baseband, before any pre-emphasis. Therefore in the time domain the baseband signal can be written as:

$$V_B(t) = \sum_{n=1}^{N} B(n) \cos(nwt + \emptyset(n)) + B(0)/2 \tag{21}$$

where N is the total number of tones, the B(N)'s are the tone amplitudes, W is the frequency spacing of the tones (radians/sec), B(0)/2 is the DC component, for generality (=0) and $\emptyset(n)$ are the random phases of the tones. $V_B(t)$ can be written in exponential form, giving:

$$V_B(t) = \sum_{n=N}^{N} B(n)/2\ e^{j\phi(n)} e^{jnwt} \tag{22}$$

which is the inverse discrete fourier transform of the baseband signal. The magnitudes are given by B(n)/2 and the powers are equal to $(B(n)/2)^2 = B^2(n)/4$. Therefore the baseband power spectral density will be given by:

$$S_{BB}(n) = B^2(n)/4 \qquad n = 1,2,3, \ldots,N \tag{23}$$

Now $R_B(\tau)$ can be calculated by taking the inverse transform of $B^2(n)/4$, since the autocorrelation and the spectral density are fourier transform pairs, (Wiener-Khintchine). Note that the random tone phases, $\emptyset(n)$, do not appear in equation (23) and therefore play no part in the calculation of $S_{BB}(n)$, and hence of $R_B(\tau)$. Thus the $\emptyset(n)$'s are set to zero in the simulation program.

Once $R_B(\tau)$ is known, one can proceed to calculate $R_c(\tau)$ using equation (20). From here the spectral density of the modulated carrier can be calculated taking the discrete fourier transform of the autocorrelation function, $R_c(\tau)$.

5. COMPUTER PROGRAM DESCRIPTION

A computer program has been developed which simulates the FDM-FM interference system for analog message into analog message channel loadings. The overall program consists of a main program and four subroutines. Each subroutine will now be described along with a discussion on how it is used in the overall simulation.

A) Subroutine FFREAD (N, ISP, IDP, RSP, RDP)

This subroutine allows the user to read in a single variable from a remote keyboard. The main feature of this routine is that any one of the four different types of variables can be read in (single-precision integer or real variables and double-precision integer or real variables). Once the user has entered the variable the routine can return the received variable back to the users' computer terminal in it's specified format so that the user is sure of what has been read by the system.

B) Subroutine BASBND (NTONE, AMP, PHASE, DF, L, WEIGHT, NCYCLE, POWER, ICALL, MONTE, ISHAPE, FREQ, NCHAN, IPRINT, NO, NORD, NTONE1, NCT, TTDEUN, ITERM, IASK)

This subroutine is used to set up the baseband for both the wanted and interfering channels (if modulated). The baseband is simulated with a set of equally spaced tones. The user must feed in the appropriate baseband upper and lower frequencies along with the number of tones, the channel loading and the RMS test tone deviation. The baseband is now simulated as follows:

F_{MAX_W} = maximum frequency of the wanted baseband (input by user)

F_{MIN_W} = minimum frequency of the wanted baseband (input by user)

$NTONE_w$ = number of baseband tones (input by user, for wanted channel only)

$DF = FMAX_w/FLOAT(NTONE)$ = tone spacing (calculated by the program). Therefore $FMAX_w$ will be one of the tone values while $FMIN_w$ will generally lie somewhere between two of the tone values. For simulation purposes a new $FMIN_w$ is calculated within the subroutine so that it is represented by a tone value. This bottom tone value is calculated as follows:

$$FMIN_w' = FMIN_w - DF/10$$

Now the bottom frequency = N* DF $(N \geq 1)$ (24)

where $FMIN_w' + DF > N^* DF \geq FMIN_w'$.

To establish the interfering signal baseband the frequency parameters must be input by the user as above. A value for NTONE is not fed in since it is calculated using the previously calculated frequency spacing (DF). Thus the interference baseband is simulated as follows:

$FMAX_i$ = maximum frequency of the interfering baseband (input by user)

$FMIN_i$ = minimum frequency of the interfering baseband (input by user)

DF remains the same as that calculated for the wanted baseband.

$$NTONE: = NTONE_w * \frac{FMAX_i + DF/2}{FMAX_w}$$

$$FMIN_i' = FMIN_i - DF/10$$

Now the bottom tone frequency = N* DF

Where $FMIN_i' + DF > N^* DF \geq FMIN_i$

Once the flat baseband tone amplitudes are determined, they are pre-emphasized according to the options available in the program. One can choose either a flat characteristic, CCIR emphasis, pure phase modulation or TD-2 emphasis. The first three shapes are generated within the subroutine

BASBND, while the TD-2 emphasis is performed in a separate subroutine PREMPH, which can be modified to calculate any pre-emphasis shape and will be described later. The CCIR and TD-2 pre-emphasis used in the program have power transfer characteristics as follows:

$$\text{CCIR [6]} \qquad L(f) = \left(\frac{V_{IN}}{V_{OUT}}\right)^2 = 1 + \frac{6.90}{\dfrac{5.25}{1+\left(\dfrac{fr}{f} - \dfrac{f}{fr}\right)^2}} \tag{25}$$

where $fr = 1.25\ F_{MAX}$

and F_{MAX} = Upper Baseband Frequency.

$$\text{TD-2 [7]} \qquad L(f) = \left(\frac{V_{IN}}{V_{OUT}}\right)^2 = (\alpha + 1)^2 \left(\frac{1 + \left(\dfrac{f}{f_2}\right)^2}{1 + \left(\dfrac{f}{f_1}\right)^2}\right) \tag{26}$$

where f_2 is the pre-emphasis corner frequency (f_2 is normally not equal to the F_{MAX} of the baseband since one pre-emphasis network is designed for a variety of channel loadings (baseband bandwidths)) and f_1 is given by:

$$f_1 = \frac{f_2}{\alpha + 1} .$$

where α is a parameter used in the design of the emphasis networks. The emphasis networks used in this sumulation were calculated with $\alpha = 4$. The de-emphasis networks have an inverse characteristic to that of the pre-emphasis so that (ideally) if a signal is passed through both a pre-emphasis and a de-emphasis network (back to back), the output signal spectrum should be unchanged. Also the total baseband power should be unaltered after either pre-emphasis or de-emphasis. In practice this is normally not the case since one particular emphasis network will be used for a variety of channel bandwidths (ie., a 1200 channel pre-emphasis network may be used for a 1320 channel system). Therefore the crossover frequency has to be adjusted for different channel loadings so that excess

CCIR used for curves of C/I

gain or attenuation is not added into the system. The pre-emphasis response can be adjusted by adding in a gain or attenuation term so that the total baseband power is unchanged (insertion loss = 0). For this simulation, the adjustment factor has not been included in the emphasis networks since the baseband power is scaled to the desired level after the pre-emphasis operation.

As mentioned above, the baseband power is scaled after pre-emphasis so that the correct level is established. The mean power level (MPL) for a single voice channel has been estimated by the CCITT as -15 dBm0 for high capacity systems (over 239 voice channels). This value was determined as follows,

$$MPL = 10 \ LOG_{10} \ \left(\frac{MST + MS \ . \ MAF}{1mW}\right) \tag{27}$$

where MST - Mean power level of signalling and tones = $10\mu W$

MS - Mean power level of speech = $88\mu W$

MAF - Mean activity factor = .25

$\therefore$ MPL - 10 LOG_{10} .032 $\simeq$ -15 dBm0.

Therefore the mean speech power (MSP) in a baseband carrying speech is equal to,

$$MSP = -15 + 10 \ LOG_{10} \ N \quad dBm0 \tag{28}$$

where N is the number of voice channels. This expression is considered valid for all channel loadings greater than or equal to 240. For medium capacity systems ($60 \leq N < 240$), the MSP works out to be:

$$MSP = -1 + 4 \ LOG_{10} \ N \quad dBm0 \tag{29}$$

and

$$MSP = +2.6 + 2 \ LOG_{10} \ N \quad dBm0 \tag{30}$$

for low capacity systems ($12 \leq N < 60$). For systems with less than 12 channels capacity, there is as yet no generally accepted expression for the MSP.

C. Subroutine PREMPH (NT, WT, NCT)

The PREMPH subroutine is used to calculate the weighting factors for any other pre-emphasis networks, including the TD-2 networks. The TD-2 weighting factors are calculated using the expression given previously. The PREMPH subroutine has been used to simulate actual white noise test loading with CCIR emphasis, by extending the tones beyond F_{MAX}, tapered off in amplitude to simulate an actual low pass filter characteristic. The PREMPH subroutine has also been used to perform a crude simulation of video and digital loading.

D. Subroutine DPHARM (A, S, M, IFS, IFERR)

This is a fast fourier transform (FFT) subroutine used to calculate a discrete fourier transform of a complex time signal, or calculate an inverse discrete fourier transform from the frequency domain information (magnitude and phase) of a signal. An FFT is an algorithm which calculates a discrete fourier transform or inverse discrete transform in a computationally efficient manner.

E. Main Program INTFER

INTFER makes use of the above mentioned subroutines to carry out the required calculations for determining the noise introduced in to the baseband of a wanted channel by an interfering channel (modulated or unmodulated). The steps involved in calculating the D/S ratio have been outlined in the previous section. The D/S ratio is the distortion power in a voice slot relative to the normal signal power, expressed in dB. The D/S ratio is then inverted in the program and becomes the more familiar NPR (Noise Power Ratio).

6. OPTIONS

The INTFER program has a variety of options available to the user, so that the simulation can be executed for different conditions. The order of the simulation is manually set by the user to a value between 2 and 11 (=N). Then 2^N complex points are used in calculating the RF spectrum. Therefore, as an extreme, if 2^N just equals 2 times the number of tones used in simulating the baseband, then only one sideband is used to estimate the RF spectrum. As a rule of thumb, N should be chosen such that 2^N is $\geq$ 6 times the number of tones used in the baseband so that aliasing effects are not introduced into the RF spectrum. As mentioned earlier, the emphasis can be any desired shape, such as a flat response, CCIR, triangular (phase modulation), or TD-2. There is an option as to whether the corresponding de-emphasis is included, or whether a flat response is used after demodulation. The number of tones (NTONE) used in simulating the baseband is optional, with a maximum set at 500. The number of tones used can affect the simulation results, so the choice is not completely arbitrary. If the value of NTONE is too small, then the calculated interfering powers will be inaccurate. This inaccuracy, though partly due to the non-Gaussian-ness of a small number of tones, is primarily due to errors in calculating the RF carrier level, which is a strong function of the actual lowest baseband frequency present. It has been shown that the lowest baseband frequency is made equal to a tone frequency. Thus if the baseband resolution is low, the estimated lowest frequency may be in error and may cause a significant error in the RF carrier component, which in turn may affect the interfering power levels in the region of the carrier component.

Once the spectrum of the carrier modulated by the combined baseband is determined, the user has the option of plotting this spectrum. Then the user has the option of outputting any or all of the following information:-

a) NPR in each "slot" across the occupied baseband, plus the total-bandwidth rms S/N,

b) interference power in dBm0 in each slot across the entire baseband (including both below and above the band occupied by the FDM modulation) after de-emphasis, plus the total-bandwidth rms S/N after de-emphasis and weighting (the latter used in video simulation),

c) NPR in the computer-determined noisiest slot, versus automatically-incremented carrier separation frequency,

d) as in c), except that the output is in dBm0,

e) Total-bandwidth S/N after de-emphasis and weighting, versus automatically-incremented carrier separation frequency (for the video simulation).

Note that the S/N calculation above, was added primarily for video simulation, and is the ratio of total wanted signal loading power to total interference power, integrated across the full wanted signal baseband (to a user-defined F_{MAX}) after user-defined noise weighting (FLAT; 525-line CCIR weighting; 525-line CBC weighting; user-supplied subroutine "weigh") and de-emphasis have been applied.

7. DERIVATION OF C/I CURVES

The output using the d) option, as described previously, is in the form of plots of the interference power in dBm0 in the noisiest baseband slot versus the carrier frequency separation, normally calculated at 0 dB C/I. Curves giving C/I objectives for 4 dBrnc0 of noise can be derived from these output plots as follows: -

$$0 \text{ dBrn} = -90 \text{ dBm} \tag{31}$$

where dBrn is referred to as dB above reference noise, where reference noise is 1 pw (10^{-9} watts). When C-message weighting is used the relationship becomes: 1pw = 10^{-9} W

$$0 \text{ dBrnc0} = 2 \text{ dBrn} = -88 \text{ dBm} \tag{32}$$

The factor of 2 dB is due to the attenuation of flat noise in the voice baseband (300 - 3400 Hz) caused by a C-message weighting network (used to simulate subjective hearing effects).

Thus,

$$\begin{aligned} 4 \text{ dBrnc0} &= (4-88) \text{ dBm0} \\ &= -84 \text{ dBm0} \end{aligned} \tag{33}$$

Thus the C/I objectives are calculated as follows: -

$$\begin{aligned} \text{C/I Obj.} &= X \text{ dBm0} - (-84 \text{ dBm0}) \\ &= (X + 84) \text{ dBm0} \end{aligned} \tag{34}$$

where X dBm0 is the value computed by the program.

Note that the interference power, for this "ordinary" interference, varies dB-for-dB with the C/I ratio. Thus it is a simple matter to scale the C/I objective curves for other than 4 dBrnc0 of interference. In particular, one can do the 84 dB scaling, required by equation (34), automatically by simply using a value of -84 dB for the C/I. The dBm0 values output by the program are then numerically equal to the required C/I values for 4 dBrnC0 noise.

When determining the C/I objectives for carrier separation frequencies which fall in the wanted channel baseband one has to include a "burble"[3] factor in the calculations. This is done because the D/S levels calculated by the program assume that the two carriers beat and cause pure tone noise in the baseband. Generally, the noise is spread throughout several voice channels so that the true noise in the baseband slot is less than that calculated by the program. This noise is due to the fact that one or both of the carriers are low frequency modulated, due to noise and equipment instability, thus the results of beating are not stable tones. The burble factor for the unmodulated interfering carrier is taken as 5 dB and for a modulated interfering carrier is taken as 10 dB. Then the resultant C/I objective can be calculated as follows,

$$\text{C/I Obj.} = 84 \text{ dBm0} + X \text{ dBm0} - \text{BF (dB)} \tag{35}$$

for carrier separations within the wanted baseband, where BF = 5 dB for the unmodulated carrier interference and BF = 10 dB for the modulated carrier interference, and

$$\text{C/I Obj.} = 84 \text{ dBm0} + X \text{ dBm0} \tag{36}$$

for carrier separations outside the wanted baseband.

8. REFERENCES

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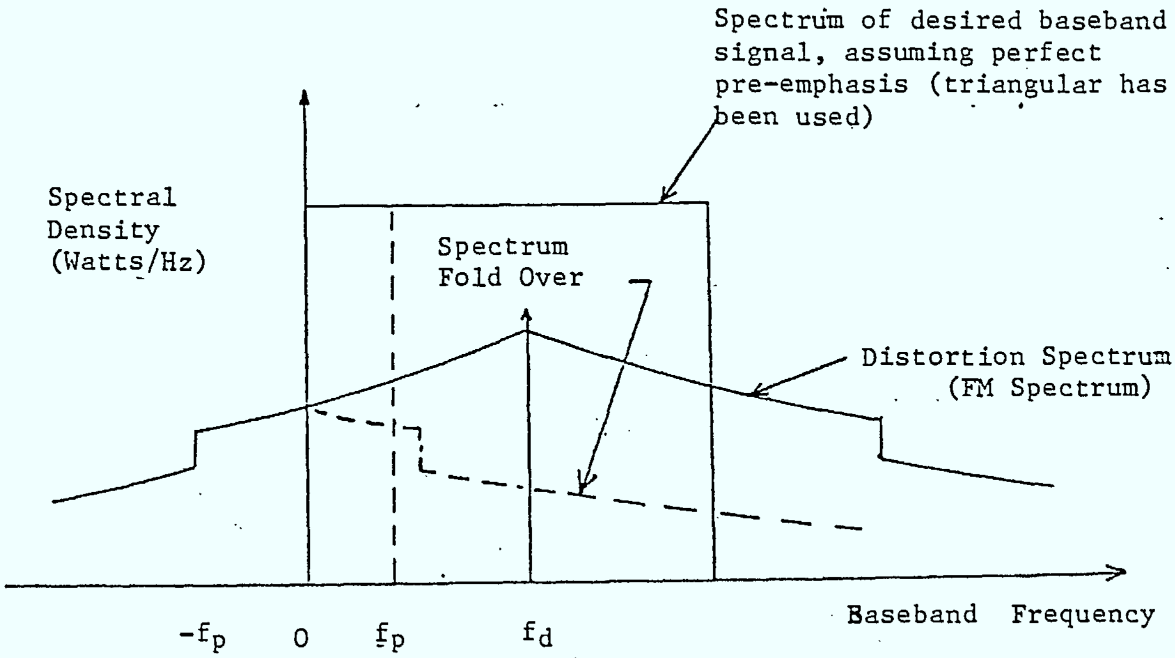
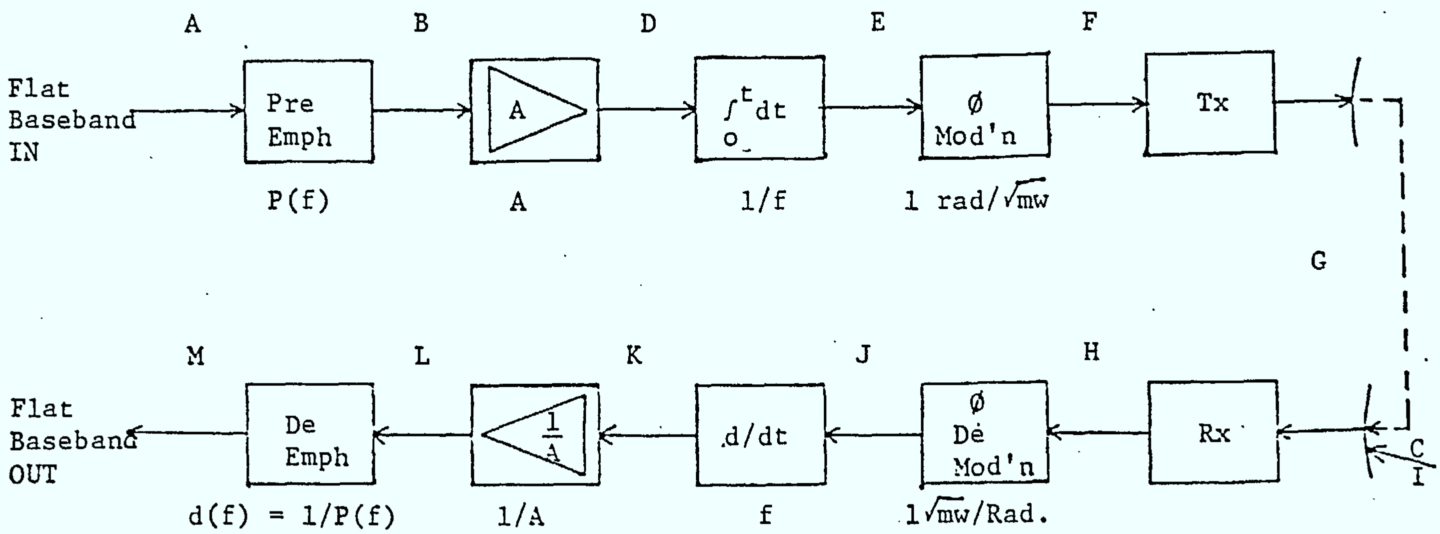


Wanted Channel Baseband Spectrum

FIGURE 1



FM Radio System Model

FIGURE 2

APPENDIX B

DERIVATION OF THE CONVOLUTION MODEL OF FM INTERFERENCE FROM THE BASEBAND SPECTRUM ADDITION MODEL.

The Baseband Addition FM-Interference Model described in Appendix A and the Convolution Interference Model [Pontano, Fuenzalida and Chitre, IEEE Transactions on Communications, June 1973, pp. 714-745] will be shown below to be identical, by a simple derivation of the convolution model from the baseband addition model.

Let $S_{bbn}(f)$ = Baseband power spectrum of the nth signal

$R_{bbn}(\tau)$ = Baseband autocorrelation function of the nth signal

$R_{rfn}(\tau)$ = RF autocorrelation function of the nth signal, after modulation

$S_{rfn}(f)$ = RF power spectrum of the nth signal, after modulation.

Preamble (Calculation of FM Power Spectrum)

The well-known Wiener-Kintchine Theorem states that the power spectrum and the autocorrelation function are Fourier Transform pairs. Thus, if one is known, the other is easily calculated (by computer, with the advent of the Fast Fourier Transform).

Stewart [Proceedings of the IEEE, October 1974, pp. 1539-1542] and Middleton ["An Introduction to Statistical Communication Theory", New York: McGraw-Hill, 1960] have shown that, for a zero-mean random baseband signal, the RF autocorrelation function of an angle-modulated signal may be obtained from the baseband autocorrelation function of the signal as follows:-

$$R_{rf}(\tau) = \exp\left[R_{bb}(\tau) - R_{bb}(o)\right]$$

It is well known that a set of random-phase sinusoids can be used to generate a noise-like baseband signal, and that this noise-like baseband signal can be used to simulate an FDM/FM baseband. The squares of the amplitudes of these baseband tones thus represent the power spectrum of the baseband signal, and the tone amplitudes are easily weighted mathematically by any desired pre-emphasis characteristic. A frequency modulation baseband signal is converted to a phase (angle) modulation baseband signal by integrating the signal. In the frequency domain, this signal integration is carried out simply by dividing each tone amplitude by the corresponding tone frequency. Reverse-Fourier Transforming the resulting integrated random-tone baseband power spectrum yields the baseband autocorrelation function. Exponentiating this then yields the RF autocorrelation function of the FM signal. Finally, forward-Fourier Transforming the RF autocorrelation function yields the desired FM power spectrum. It should be noted that this method of calculating the FM spectrum of a noise-modulated signal is not only easy to implement, but is also not restricted in the value of modulation index (total baseband loading power) used.

Derivation (of convolution model from baseband addition model).

Godier showed that an angle-modulated interfering signal appears to another angle-modulated (wanted) signal as if it were a carrier that was modulated with the baseband signals of both the interfering and the wanted signals (a power sum), the only restriction being that the (C/I) power ratio be much greater than unity. It remains then to derive the RF power spectrum of this doubly-modulated carrier.

Let the wanted-signal baseband power spectrum, after pre-emphasis and integration, be $S_{bb1}(f)$ and that of the interfering signal be $S_{bb2}(f)$. The combined-signal baseband power spectrum, $S_{bb3}(f)$, is then:

$$S_{bb3}(f) = S_{bb1}(f) + S_{bb2}(f).$$

Now, since the Fourier Transform of the sum of two signals is equal to the sum of the Fourier Transforms of each signal, it follows that the baseband autocorrelation function of the combined signal is simply the sum of the two individual autocorrelation functions, i.e.:

$$R_{bb3}(\tau) = R_{bb1}(\tau) + R_{bb2}(\tau).$$

The RF autocorrelation function of the carrier that is angle modulated by the combined signals is then:

$$\begin{aligned} R_{rf3}(\tau) &= \exp\left[R_{bb3}(\tau) - R_{bb3}(o)\right] \\ &= \exp\left[(R_{bb1}(\tau) - R_{bb1}(o)) + (R_{bb2}(\tau) - R_{bb2}(o))\right] \\ &= \exp\left[R_{bb1}(\tau) - R_{bb1}(o)\right] . \exp\left[R_{bb2}(\tau) - R_{bb2}(o)\right] \end{aligned}$$

i.e., $R_{rf3}(\tau) = R_{rf1}(\tau) \cdot R_{rf2}(\tau)$

Finally, since multiplication in the time domain is identical to convolution in the frequency domain, it follows that the RF power spectrum of the carrier that is modulated by the combined baseband signals is given by:

$$S_{rf3}(f) = S_{rf1}(f) * S_{rf2}(f),$$

where * stands for convolution.

Thus the combined-signal RF power spectrum can also be obtained by convolving the two individual RF power spectra, and this is identical to adding together their two baseband power spectra.

QED

APPENDIX C

IMPLEMENTATION OF CONVOLUTION MODEL FOR FM INTERFERENCE IN INTFER COMPUTER PROGRAM.

As is described in Appendix A, the original INTFER computer program started off by generating the wanted channel FDM/FM baseband spectrum from a set of random tones, after applying pre-emphasis and signal integration (to convert to a phase modulator spectrum). INTFER then stored the resulting baseband spectrum and went on to obtain, in the same manner, the interfering channel baseband spectrum. These two spectra were then added on a power basis, and the combined spectrum was Fourier Transformed to become the effective interference baseband autocorrelation function. The effective RF autocorrelation function of the interference was then obtained from the baseband autocorrelation function, as described in appendices A and B, and this was Fourier Transformed to become the effective RF (FM), power spectrum of the interference. Appendix A describes how the resulting wanted-channel baseband interference noise is then calculated from this effective RF spectrum.

Appendix B demonstrates that, for FDM/FM modulation, the above described baseband addition interference model used in INTFER, and the currently widely accepted RF spectrum convolution interference model, are not only similar but are one and the same thing. Since the baseband addition model is only applicable for noise like, zero mean, baseband signals, however (due to the method used to obtain the FM Spectrum from the Baseband Spectrum), it was necessary to implement the convolution method in order to extend INTFER to deal with digital and video modulation. Due to the superior computational efficiency of the baseband addition model, it was retained for FDM/FM - into - FDM/FM interference modelling.

With the convolution method, the RF (FM) power spectra of the wanted and interfering signals are separately generated and stored and are then convolved to produce the effective RF power spectrum of the interference. The existing unmodified remainder of the INTFER program then calculates and manipulates the resulting demodulated interference noise from the effective RF power spectrum, as though the baseband addition model had been used.

It should be noted that the convolution model has the advantage that it can include the effects of filtering on the wanted and interfering FM spectra (but still not the so-called DACI intelligible cross-talk caused by filer-skirt-induced AM, when it is followed by AM/PM conversion).

When the modified INTFER program is now run, the user is given the option of specifying whether the baseband addition model or the convolution model is to be used. If the convolution model is chosen, the user is then given three options for obtaining the RF spectra (separately) of the wanted and the interfering signals. The first option allows either (but not both) signal(s) to be unmodulated, in which case the effective RF spectrum is set equal to the modulated channel spectrum (ie. no convolution is required). If the wanted channel is unmodulated, a dummy pass is made through the FDM/FM baseband subroutine to establish values for the baseband frequency range of interest, the test-tone level, the type of pre-emphasis (and hence of de-emphasis), etc, to be used, as this information is needed in order to evaluate the severity of the interference noise.

Option 2 allows either (or both) channels to be modulated by an FDM/FM signal, in which case a random tone baseband spectrum is generated, as is done with the baseband addition model, and the channel RF spectrum is generated via the autocorrelation functions as described before.

Option 3 allows either (or both) channels to obtain their RF power spectra by means of a new subroutine, SPECTM. As with Option 1, a dummy pass is made through the FDM/FM baseband subroutine to set up frequency limits, etc. The subroutine SPECTM, in turn, gives the user the option of either calculating a ((sine-x)/x) -squared spectrum, calculating a digital FSK spectrum, or reading a stored (eg. video) spectrum from a data file and suitably interpolating between and extrapolating from, the stored values. Subroutine SPECTM then allows the user to specify lossless

equi-ripple

Tchebychev filtering (any bandwidth, number of poles, and in-band ripple) to be applied to the above spectrum. The ((sine-x)/x)-squared, FSK and Tchebychev equations used are given in Appendix D.

The process of spectrum convolution is carried out by means of the Fast Fourier Transform. The two RF power spectra are separately transformed to the time domain to become the RF autocorrelation functions. These two autocorrelation functions are then multiplied together (ie, corresponding array elements) and the result is transformed back to the frequency domain, to become the desired convolved RF spectrum.

Computer runs made with the modified INTFER program produce identical results for FDM-into-FDM interference for the two interference models.

APPENDIX D

EQUATIONS USED FOR

PSK AND FSK DIGITAL SPECTRA,

TCHEBYCHEV FILTERING,

AND FCC MASK.

MSK - minimum shift keying

PSK SPECTRA

For PSK digital transmitters that have equiprobable mark-space states, unity mark/space ratio, and that do not employ baseband-pulse-shaping, the output power spectrum, before output band-pass filtering, can be represented, regardless of the number of digital levels employed, by the familiar ((sine-x) / x) - squared shape.[1,2,3] This basic shape is then modified by a known transmitter output filter shape in this study. The pre-filtering PSK spectral density for a 1-watt signal is thus given by:

$$S(|f-fo|) = T.\left[\frac{\sin[\Pi|(f-fo)|.\ T]}{\Pi|(f-fo)|.T}\right]^2 \ \frac{\text{watts}}{\text{Hz}},$$

where fo is the modulated carrier frequency, Hz, and T is the reciprocal of the Baud rate, sec.

For binary PSK, the Baud rate is equal to the bit rate. For 4 PSK, since the modulation can be considered to be made up of two orthogonal binary bit streams, the Baud rate is 1/2 the bit rate (law of superposition). Similarly, for 8 PSK, the modulation can be considered to be made up of three orthogonal binary bit streams, reducing the Baud rate to 1/3 the bit rate. Continuing, the Baud rate for 16 PSK would be 1/4 the bit rate, etc, etc. Thus, for a given bit rate, the transmitted spectrum gets progressively narrower as the number of signal states or levels is increased (but the digital signal, itself, becomes progressively more susceptible to noise and interference).

FSK SPECTRA

The normalized spectrum for an N-level digital FSK signal was calculated using the method of Anderson and Salz[4], as shown below:

$$\frac{G(\beta)}{A^2T} = \frac{1}{N}\sum_{n=1}^{N}\left[\frac{1}{2}\,\frac{\sin^2\gamma n}{\gamma n^2} + \frac{1}{N}\sum_{m=1}^{N}\beta.\,\frac{\sin\gamma n}{\gamma n}\cdot\frac{\sin\gamma m}{\gamma m}\right] \qquad (2)$$

where G(f) = power spectral density at frequency f,

A = signal amplitude (arbitrary)

T = modulation Baud rate

$$\beta = \frac{(w - w_c)\ T}{2\Pi} \qquad (3)$$

$$\gamma_n = \left(\beta - \frac{\alpha_n\ .\ k}{2}\right)\Pi \qquad (4)$$

$$\alpha_n = 2_n - (N + 1) \qquad (5)$$

$$k = \frac{(W_m - W_s)\ .\ T}{2\Pi} \qquad (6)$$

W_m = mark frequency

W_s = space frequency

$$\beta = \frac{\cos(\gamma n + \gamma m) - C\alpha\cos(\gamma n + \gamma m - 2\Pi\beta)}{1 + C\alpha^2 - 2\,C\alpha.\cos 2\Pi\,\beta} \qquad (7)$$

$$C\alpha = \frac{2}{N}\sum_{n=1}^{N/2}\cos\left[k\Pi\ (2n-1)\right] \qquad (8)$$

This study was restricted to:

N = 8 level

$k = \frac{1}{N} = 0.125$

TCHEBYCHEV FILTER (infinite - Q)

$$\text{Loss, dB} = 10.\ \log_{10} \left\{ 1. + \varepsilon.\ \cos^2 \left[n.\ \cos^{-1} \left(\frac{w'}{w_1'} \right) \right] \right\} \quad w' \leq w_1' \tag{12}$$

$$= 10.\ \log_{10} \left\{ 1. + \varepsilon.\ \cosh^2 \left[n.\ \cosh^{-1} \left(\frac{w'}{w_1'} \right) \right] \right\} \quad w' > w_1' \tag{13}$$

where: n = number of low-pass-equivalent filter poles, (14)

$w' = \left| (w-wo) \right|$

$w_1' = \frac{1}{2}$ ripple bandwidth (low-pass equivalent)

$$\varepsilon = \left[\text{antilog}_{10} \left(\frac{Lr}{10.} \right) - 1. \right], \quad ? \tag{15}$$

and Lr = ripple, dB (in-band).

In this study, n and Lr were arbitrarily fixed at 5 poles and 0.1 dB, respectively. Then for each PSK and FSK signal (Baud rate), and for each licensed bandwidth, a minimum value for w_1' was empirically determined, which was just sufficient to make the filtered transmitter spectrum lie below the FCC Mask with a nominal 1 dB margin at the closest point. Finite filter Q's would then increase this margin several dB, for frequency drifts etc. ?

16 dB return loss

FCC MASK[5] (Below 15 GHz)

Let B_n = licensed bandwidth, or necessary bandwidth in MHz.

The FCC rules then require the mean out-of-band power of any emissions in a 4 kHz bandwidth to be attenuated below the mean total output power of the transmitter by the following amounts: -

a) At frequencies that are removed from the assigned frequency by $>50\%$ of B_n out to $\leq 250\%$ of B_n:

$$\text{Atten, dB} = 35. + 0.8\ (P-50.) + 10.\log_{10} B_n$$

where P = percent removed from the assigned (carrier) frequency, and providing $50 \leq \text{Atten} \leq 80$ dB

b) At frequencies removed from the assigned frequency by $>250\%$ B_n:

$(43 + 10 \log P_w) \leq \text{Atten} \leq 80$ dB.

Where P_w = mean output power, watts (= 1 watt in this study).

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APPENDIX E

TABLES AND DERIVATIONS OF SIGNAL PARAMETERS FOR C/I CURVES.

Tables 1 through 3 give the parameters used in the study for the FDM/FM wanted signal analogue loadings, and the PSK and FSK interfering signal digital loadings, respectively. The latter two tables include the filter parameters that were found necessary to force the digital spectra to conform, with 1 dB margin, to the FCC Masks for the channel bandwidths given in the left-hand columns. The filter parameters are for infinite-Q. Finite-Q filters will provide additional margin, to allow for signal frequency tolerances and drifts, filter tolerances and drifts, etc.

FDM/FM Loadings

For each licensed bandwidth in Table 1, the minimum channel loading (no. of voice channels) was taken from the relevant SRSP, and the maximum channel loading (when not given in the SRSP) was calculated using the necessary bandwidth formulas in the latest version of RSP-113. The standard CCIR test-tone deviations and baseband frequency limits were used in each case, the frequency limits in general being adjusted slightly to become integral multiples of convenient simulation-tone (see Appendix A) frequency spacings. CCIR pre-and de-emphasis characteristics were used, with the pre-emphasis fmax made to correspond to the actual baseband fmax used, and with the total baseband loading power being measured after pre-emphasis. In each case, the number of simulation tones used (for the FDM/FM baseband signal), and the frequency spacing between tones, were chosen to allow a total simulation bandwidth in the storage arrays (2048 points) that was sufficiently large to accomodate the broadest (after convolution) digital spectrum (highest Baud rate) at the required adjacent channel frequency spacing.

As stated in our proposal, for each digital interference case, an unmodulated wanted-channel case was included in order to obtain a best-case C/I curve in the region of the adjacent frequency spacing. The removal of the wanted channel modulation has the effect of stopping the spectrum convolution process, and assumes that the wanted channel receiver filtering removes the bulk of the interfering channel signal power (removes all power outside the wanted signal channel bandwidth), leaving only the direct "spill-over" power, ie., that part of the interfering signal spectrum which falls inside the wanted signal channel bandwidth and hence cannot be attenuated by wanted channel filtering. This total "spill-over" power is assumed to be sufficiently low that it can be treated simply as an increase in the wanted channel receiver "front-end" thermal noise.

Video Loading

Video modulation was applied to the wanted signal, but only for the 20 MHz and 29.65 MHz licensed bandwidth channels. The video FM spectrum was generated using another in-house computer simulation program (FMSIM). The video modulating waveform used was a standard de-saturated colour bar test pattern (excluding the vertical synchronizing intervals). The colour bar baseband signal was low-pass filtered at approximately 5 MHz, and a 6.8 MHz audio program channel was added, at 284 kHz peak frequency deviation (of the main radio). Standard 525-line CCIR video pre-and de-emphasis networks were used.

PSK Loadings

The absolute maximum Baud rate for a given channel (licensed) bandwidth was arbitrarily taken as 10% below the Nyquist rate, or in other words,

$$\text{Max Baud rate} = \frac{\text{licensed bandwidth}}{1.1}$$

This is the value used for the first entry for each licensed bandwidth in Table 2. Note that these absolute maximum Baud rates are not constrained to be multiples of standard bit streams.

For a general digital radio system, an "implementation factor" of 1.3 rather than 1.1 was used for determining the maximum practical Baud rate for each licensed bandwidth. Then the maximum bit rates for four and eight digital levels were calculated from the Baud rates (see Appendix D). Then the nearest integral multiples of standard digital bit streams (DS1, DS2, or DS3), after allowing for 2% bit-stuffing, that would produce no more than the calculated maximum bit rates, were used to determine the maximum practical Baud rates for the 4 PSK and 8 PSK digital signals. These values were given in Table 1 of our proposal. When the interference C/I curves were subsequently computed and plotted however for these 4 PSK and 8 PSK Baud rates, it was found that the resulting curves were essentially superimposed. For this reason, only a single curve was plotted then, for each such interference case, using a Baud rate equal to the average of the values obtained for 4 PSK and 8 PSK. These values are given as the 2nd entry for each licensed bandwidth in Table 2. It was in order to broaden the scope of the study, then, that the absolute maximum Baud rate cases explained at the beginning of this section, were added (1st entries in Table 2), and these can also be taken to apply to both 4 PSK and 8 PSK (and also 2 PSK, 16 PSK, etc). An example calculation for the 3.5 MHz licensed bandwidth case is given below:

$$\text{Max Baud Rate} = \frac{3.5}{1.3} = 2.692 \text{ MBs}$$

$$4 \text{ PSK} = 2^2 \text{ digital levels}$$

$$\therefore \text{Max 4 PSK bit rate} = 2 \times 2.692 = 5.385 \text{ MBs}$$

$$\text{Max no. of DS1 streams} = \frac{5.385}{1.02 \times 1.544} = 3$$

$$\therefore \text{Actual 4 PSK Baud Rate} = \frac{3 \times 1.02 \times 1.544}{2} = 2.362 \text{ MBs}$$

$$8 \text{ PSK} = 2^3 \text{ digital levels}$$

$$\text{Max 8 PSK bit rate} = 3 \times 2.692 = 8.076 \text{ MBs}$$

$$\text{Max. No. of DS1 streams} = \frac{8.076}{1.02 \times 1.544} = 5$$

$$\text{Actual 8 PSK Baud Rate} = \frac{5 \times 1.02 \times 1.544}{3} = 2.625 \text{ MBs}$$

$$\therefore \text{Average Baud Rate} = \frac{2.362 + 2.625}{2} = 2.49 \text{ MBs}$$

To further broaden the scope of the C/I curves, a set of lower-bound Baud rates were worked out and appear as the 3rd and 4th entries for each licensed bandwidth. These entries are, respectively, the 4 PSK and 8 PSK Baud rates corresponding to the minimum number of standard bit streams desired to be carried in the given licensed bandwidths, as given in the 6th column of Table 2.

8FSK Loadings

It was stated in our proposal, since the FSK spectrum appeared to fall off relatively fast away from the carrier, that no filtering would be used with the FSK signals. However, in making some trial computer runs, it was found that, if Baud rates appoaching those used for the PSK signals were desired, some filtering would be needed. To best facilitate comparison of FSK and PSK interference, then, it was decided to use the same Baud rates for FSK as were used for PSK, and also the same complexity of filtering. The resulting FSK C/I curves were found to lie so close to the corresponding PSK C/I curves that only the absolute maximum Baud rate cases were run with the FSK signals. Table 3 lists the Baud rates used and the empirically determined filter parameters (bandwidths) required to meet the FCC Masks with 1 dB margin.

QPRS Loading

Only the Northern Telecom Ltd. DRS-8 transmitter spectrum was modelled, as stated in our proposal. Although the DRS-8 signal occupies a 40.74 MHz bandwidth, the now-obsolete 8 GHz Analogue Radio SRSP limited the analogue radios in that band to 29.65 MHz bandwidths/channel spacings, and to 600 - to - 1800 voice channel loadings. As these analogue radios could still be in use, the DRS-8 QPRS interference was analyzed for the 29.65 MHz bandwidth analogue loadings of Table 1.

The QPRS signal parameters used are given below:-

Baud Rate = 45.52 MBs $\left(\left[\frac{\text{sinex}}{x}\right]^2\right)$ How does this compare with FCC mask?

Transmit Filter Bandwidth = $\pm$ 14 MHz

No. of Poles = 5

Ripple = 0.06 dB

TABLE 1

FDM/FM ANALOG CHANNEL

Licensed Bandwidth MHz	Modulation Type	Fmax MHz	Fmin MHz	RMS T.T. MHz	Number of Tones	Total Loading dBm0
3.5	48 ch.	.250	.060	.200	25	+6.1
	6 ch.	.084	.060	.200	14	+4.8
	Unmod.	.250	.060	.200	25	___
7.0	120 ch.	.540	.060	.200	27	+7.3
	60 ch.	.300	.060	.200	15	+6.1
	Unmod.	.540	.060	.200	27	___
9.75	300 ch.	1.300	.057	.200	46	+9.8
	120 ch.	.540	.057	.200	19	+7.3
	Unmod.	1.300	.057	.200	46	___
20.	1260 ch.	5.772	.315	.140	99	+16.0
	600 ch.	2.690	.0585	.140	46	+12.8
	Unmod.	5.772	.315	.140	99	___
29.65	1800 ch.	8.210	.312	.140	95	+17.6
	600 ch.	2.680	.0865	.140	31	+12.8
	Unmod.	8.210	.312	.140	95	___
40.	2700 ch.	12.400	.316	.140	106	+19.3
	1260 ch.	5.740	.316	.140	49	+16.0
	Unmod.	12.400	.316	.140	106	___

TABLE 2

PSK DIGITAL CHANNEL

Licensed Bandwidth MHz	Baud Rate MBs	Filter Bandwidth MHz		No. of Poles	Ripple dB MHz	
	3.18	±1.19		5	.1	
3.5	2.49	±1.26		5	.1	
	0.772	±1.39	4PSK			
	0.515	±1.38	8PSK	5	.1	(1)DS1
	6.36	±2.50		5	.1	
7.0	4.77	±2.69		5	.1	
	1.575	±3.00	4PSK			
	1.05	±2.94	8PSK	5	.1	(2)DS1
	8.86	±3.57		5	.1	
9.75	6.44	±3.88		5	.1	
	3.156	±3.90	4PSK			
	2.104	±4.10	8PSK	5	.1	(1)DS2
	18.18	±7.72		5	.1	
20.	13.90	±8.20		5	.1	
	9.657	±9.50	4PSK		.1	
	6.438	±8.40	8PSK	5	.1	(3)DS2
	26.95	±11.75		5	.1	
29.65	22.14	±12.20		5	.1	
	12.876	±13.65	4PSK			
	8.584	±13.85	8PSK	5	.1	(4)DS2
	36.36	±15.94		5	.1	
40.	26.62	±17.30		5	.1	
	22.368	±18.90	4PSK			
	14.912	±17.65	8PSK	5	.1	(1)DS3

1.544
2
3.088
3.0
1.0

TABLE 3

8-LEVEL; k=0.125; FSK DIGITAL CHANNEL

Licensed Bandwidth MHz	Baud Rate MBs	Filter Bandwidth MHz	No. of Poles	Ripple dB
3.5	3.18	$\pm$1.27	5	.1
7.0	6.36	$\pm$2.66	5	.1
9.75	8.86	$\pm$3.80	5	.1
20.	18.18	$\pm$8.22	5	.1
29.65	26.95	$\pm$12.5	5	.1
40.	36.36	$\pm$17.0	5	.1

APPENDIX F

DERIVATION OF

$\frac{C}{I}$ OBJECTIVES

As stated in our proposal, the Trans Canada Guidelines[1] have been used as a basis for deriving the $\frac{C}{I}$ interference objectives.

Long-Haul FDM/FM Objective

This is given [1] as a maximum of 4 dBrnc0 of interference noise in the noisiest voice channel (baseband "slot") per interference exposure.

The INTFER program outputs a plot of the noisiest-slot interference noise in dBm0 at a user-specified $\frac{C}{I}$ ratio, as the carrier difference frequency is uniformly incremented from zero (co-channel) to a high value ($\geq$ adjacent channel spacing). The computer scans the complete baseband, at each value of carrier separation frequency, to determine the noisiest slot interference.

Normalized interference noise outputs are obtained by feeding in a value of 0dB for the $\frac{C}{I}$. The interference noise then varies inversely dB-for-dB with the $\frac{C}{I}$ as long as one stays away from threshold and DACI effects.

Let us say we measure a worst-slot interference noise power of N dBm0 at a 0 dB $\frac{C}{I}$ and a particular carrier difference frequency.

We want only -84 dBm0 noise (=4 dBrnc0), so we need to increase the $\frac{C}{I}$ to:

$$\frac{C}{I} = N - (-84) = \underline{N + 84} \text{ dB} \qquad (1)$$

If, however, we had fed in a value of -84 dB for the $\frac{C}{I}$ (this doesn't violate the restriction that the model only holds true for large $\frac{C}{I}$ values, since the computer program simply linearly scales the normalized noise value by the $\frac{C}{I}$ value) we would have measured an interference power of:

Interference noise = <u>N + 84</u> dBm0 (2) ?

Thus equations (1) and (2) are identical, and we see that if we use a value of -84 dB for the $\frac{C}{I}$ ratio when running INTFER for the interference into FDM/FM, the values that are plotted by INTFER for the interference noise in dBm0, versus carrier separation frequency, will be numerically equal to the required values of $\frac{C}{I}$ for 4 dBrnc0 interference noise. No further scaling is thus necessary.

It should be noted that, although the term $\frac{C}{I}$ is referred to as the "carrier-to-interference" ratio, we are talking here about total signal powers (in the full IF bandwidth), rather than carrier components, as implied.

<u>Short-Haul FDM/FM Objective</u>

This is given[1] as:

"Short-haul interference requirements state that an interference signal may not cause the noise in the desired system to increase more than 1 dB under <u>fully - faded</u> conditions".

It will be shown, below, that this short-haul objective is quite consistent with the 4 dBrnc0 long-haul objective:

Typical fully-faded protection-switching trip point +52 dBrnc0 top-slot noise.

For 1 dB degradation of this, the interference noise must be 6 dB below this, or +46 dBrnc0.

Typical modern microwave radio fade margin 42 dB

Therefore, the typical interference noise during non-faded conditions = +46 −42 = +4 dBrnc0, the same value as used for long-haul.

Thus the long-haul interference $\frac{C}{I}$ curves can be used for short haul interference as well. Since the interference noise varies inversely dB-for-dB with the $\frac{C}{I}$ it is a simple matter to scale the long-haul $\frac{C}{I}$ curves to apply to a particular short-haul microwave radio design (different protection swtiching trip point and/or fade margin).

<u>Video Interference Objective</u>

<u>Method 1</u>

Equate the 4 dBrnc0 FDM/FM noise to a video $\left(\frac{S}{N}\right)$, treating the interference noise as <u>triangular</u> noise.

Assuming the FDM/FM loading is 1260 channels at 140 kHz test-tone, we have:

$$-84 \text{ dBm0} = 20 \log \left(\frac{5636}{140}\right) + 10 \log 3100 - 4 - \left(\frac{C}{N}\right)$$

where the 4 dB is the top-slot pre-emphasis improvement, and the $\left(\frac{C}{N}\right)$ is the ratio of the total received carrier (signal) power to the receiver front-end thermal noise power density.

Solving (3) we get:

$$\left(\frac{C}{n}\right) = 147 \text{ dB Hz}$$

The corresponding unweighted video peak-to-peak picture (excluding synch) to total rms noise ratio is then:

$$\left(\frac{S_{ppp}}{N_{rms}}\right) = 7.78 + 147 + 20 \log (4 \times 10^6) - 30 \log (4.2 \times 10^6) + 2.9 \qquad (4)$$

where the 7.78 dB is simply 10 log 6,

and the 2.9 dB is the pre-emphasis improvement.

Solving (4) we get:

$$\left(\frac{S_{ppp}}{N_{rms}}\right) = 91 \text{ dB}$$

$\therefore$ The weighted $\left(\frac{S_{ppp}}{N_{rms}}\right) = 91 + 9.8 = \underline{100.8 \text{ dB}}$ (CCIR)

<u>Method 2</u>

Equate the 4 dBrnc0 FDM/FM noise to a video $\left(\frac{S}{N}\right)$, treating the interference noise as <u>flat</u> noise.

Total interference noise excluding de-emphasis

$$= -84 \text{ dBm0} + 4 \text{ dB} + 10 \log\left(\frac{4.2 \times 10^6}{3100}\right) \qquad (5)$$

$$= -48.7 \text{ dBm0}$$

Peak-to-peak video frequency deviation = 8 MHz.

Picture portion of this $= 8 \times \frac{100}{140} = 5.714$ MHz

$$= 5714 \text{ kHz}$$

$$= 20 \log \left(\frac{5714}{140}\right) = +32.2 \text{ dBm0}.$$

$\therefore$ Video $\left(\frac{S_{ppp}}{N_{rms}}\right) = 32.2 - - 48.7 = 80.9$ dB

Weighting + de-emphasis improvement of flat noise is only 2.4 dB.

$\therefore$ Weighted Video $\left(\frac{S_{ppp}}{N_{rms}}\right) = 80.9 + 2.4$

$= \underline{\underline{83.3\text{ dB}}}$ (CCIR)

Method 3

The CCIR-weighted video $\left(\frac{S_{ppp}}{N_{rms}}\right)$ at the fully-faded protection switching trip point is typically in the neighbourhood of 46 dB. For 1 dB degradation of this, the $\left(\frac{S_{ppp}}{N_{rms}}\right)$ due to the interference must be 46 + 6 = 52 dB.

The microwave radio fade margin for video is typically greater than the fade margin for FDM/FM. Assuming a fade margin of 45 dB, we get a value for the weighted video $\left(\frac{S_{ppp}}{N_{rms}}\right)$ due to the interference, in non-faded conditions, of:

$$\left(\frac{S_{ppp}}{N_{rms}}\right) = 52 + 45 = \underline{\underline{97\text{ dB}}}\text{ (CCIR)}$$

For the purposes of this report, we will use a compromise round-figure value of 90 dB for the video $\left(\frac{S_{ppp}}{N_{rms}}\right)$ ratio objective. The $\frac{C}{I}$ curves can then be easily scaled dB-for-dB for other desired objectives.

The INTFER computer program performs a power-summation of the total interference noise power over a wanted-channel baseband frequency range of ≈ 0 to 4.2 MHz, after shaping the noise spectrum by the CCIR 525-line video de-emphasis and weighting characteristics. This total noise power is measured in terms of dBm0 relative to 140 kHz rms test-tone deviation. In order to convert this into an $\left(\frac{S_{ppp}}{N_{rms}}\right)$ power ratio, INTFER makes use of a reference S_{ppp} power level calculated as follows: -

$$S_{ppp} = 20 \log_{10}\left[\frac{8000 \text{ kHz p-p}}{140 \text{ kHz rms}} \times \frac{100 \text{ IRE}}{140 \text{ IRE}}\right] = +32.22 \text{ dBm0}$$

Similar to the FDM/FM case, INTFER computes a normalized video $\left(\frac{S}{N}\right)$ if a $\frac{C}{I}$ ratio of 0 dB is fed into the program.

Let us say that we measure an $\left(\frac{S}{N}\right)$ of S dB at 0 dB $\frac{C}{I}$ and at a particular carrier offset frequency. Since we require $\left(\frac{S}{N}\right) = 90$ dB, we must increase the incoming $\frac{C}{I}$ to:

$$\frac{C}{I} = \underline{90-S} \text{ dB.}$$

Now, if we had fed in a $\frac{C}{I}$ ratio of -90 dB, we would have measured an interference $\left(\frac{S}{N}\right)$ of:

$$\left(\frac{S}{N}\right) = S - 90 \text{ dB}$$
$$= -\underline{(90 - S)} \text{ dB}$$

Comparing equations (6) and (7) we see that, if we use a value of -90 dB for the $\frac{C}{I}$ ratio when running INTFER for interference into FM-video, the values that are plotted by INTFER for the interference $\left(\frac{S}{N}\right)$ versus carrier separation frequency will be numerically equal, and of opposite sign, to the required values of $\frac{C}{I}$ for our $\left(\frac{S_{rms}}{R_{rms}}\right)$ objective of 90 dB. Again, these curves can be directly scaled dB-for-dB for other desired interference objectives.

<u>Reference:</u>

1. See reference 3. of Appendix A.

APPENDIX G

PSK INTO FDM/FM CURVES

APPENDIX G

Figures G-1(a) through G-6(c) present the results of our study of PSK interference into FDM/FM signals, in the form of C/I objective curves versus carrier separation frequency for the 4 dBrnc0 interference objective derived in Appendix F. The figures are grouped in sets of three, according to licensed bandwidth. There is one figure for each of the wanted channel FDM/FM loadings given in Table 1 of Appendix E. There are four curves in each figure, covering the four interfering channel PSK Baud rates for the corresponding licensed bandwidth, as given in Table 2 of Appendix E. The rationale behind the choice of analogue and digital loadings in the tables is also given in Appendix E. The first two figures in each set of three represent the worst-case, no-receiver-filtering, C/I objectives for the maximum and minimum FDM/FM loadings, respectively, in the corresponding licensed bandwidth, and were obtained with convolution (see Section 3 of this report). As discussed in sections 2 and 3 of this report, and in Appendix E, the third figure (0 channel loading) in each set of three represents the best-case, perfect-receiver-filtering, C/I objectives for the maximum FDM/FM loading in the corresponding licensed bandwidth. These best-case curves were obtained by running INTFER with the wanted channel modulation removed (but with the baseband frequency loading scanning limits, the de-emphasis network and the test-tone level set up for the maximum FDM/FM loading case) in order to stop the RF spectrum convolution process and thus make use of only the digital RF spectrum in the interference calculations. These best-case curves are applicable at large values of carrier separation frequency, for example at adjacent channel spacings, although they are shown as extending to zero frequency separation. The figures are listed below (see Tables 1 and 2, Appendix E, for more complete details).

FIGURE	LICENSED BANDWIDTH MHz	FDM/FM CHANNELS	PSK BAUD RATES MBs
G-1 (a)	3.5	48	3.18/2.49/0.772/0.515
G-1 (b)	"	6	" " " "
G-1 (c)	"	0	" " " "
G-2 (a)	7.0	120	6.36/4.77/1.575/1.05
G-2 (b)	"	60	" " " "
G-2 (c)	"	0	" " " "
G-3 (a)	9.75	300	8.86/6.44/3.156/2.104
G-3 (b)	"	120	" " " "
G-3 (c)	"	0	" " " "
G-4 (a)	20.0	1260	18.18/13.90/9.657/6.438
G-4 (b)	"	600	" " " "
G-4 (c)	"	0	" " " "
G-5 (a)	29.65	1800	26.95/22.14/12.876/8.584
G-5 (b)	"	600	" " " "
G-5 (c)	"	0	" " " "
G-6 (a)	40.0	2700	36.36/26.62/22.368/14.912
G-6 (b)	"	1260	" " " "
G-6 (c)	"	0	" " " "

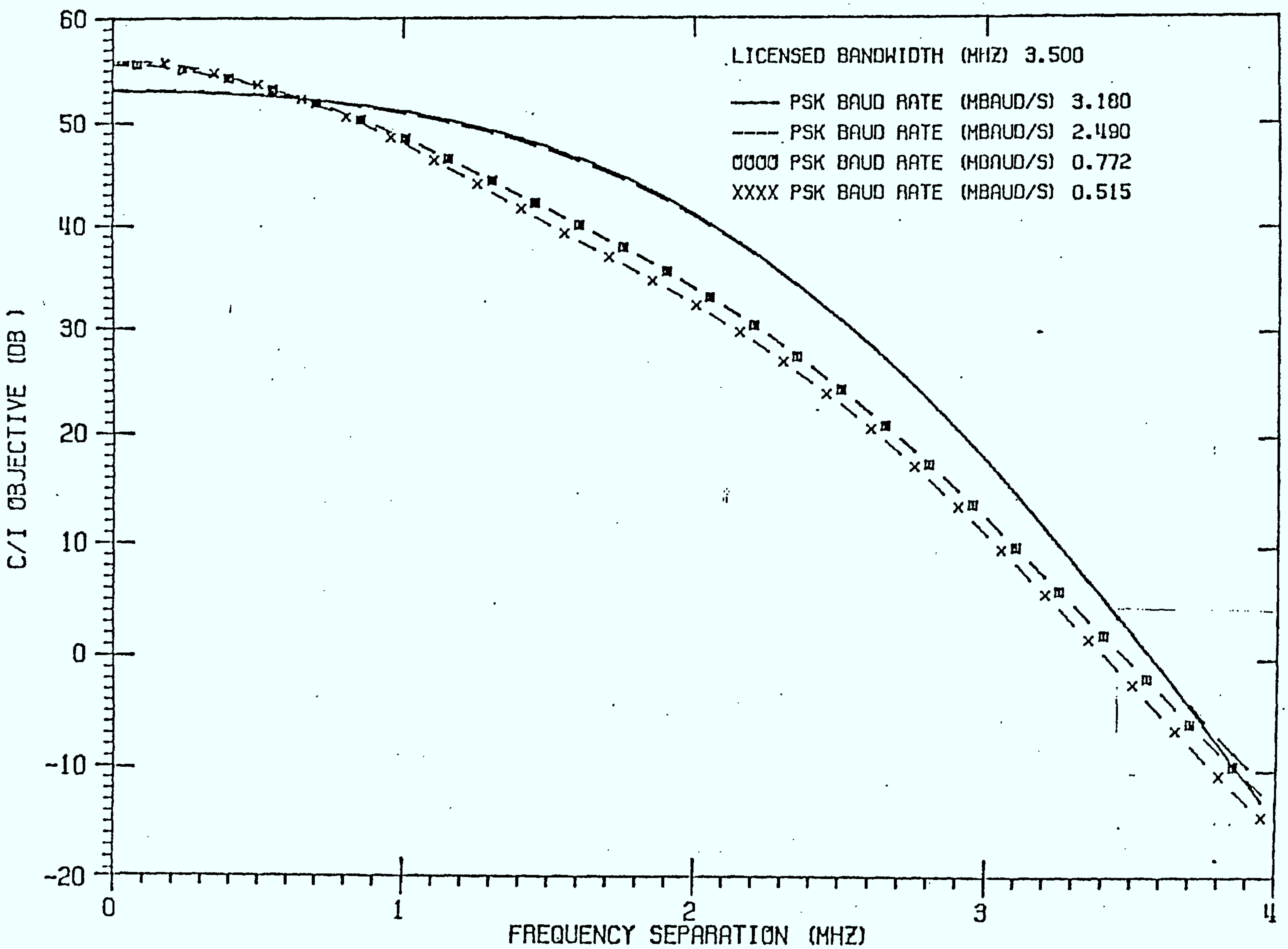


FIGURE G-1(A) C/I OBJECTIVE VS. FREQUENCY SEPARATION FOR 48 CHANNEL LOADING FOR

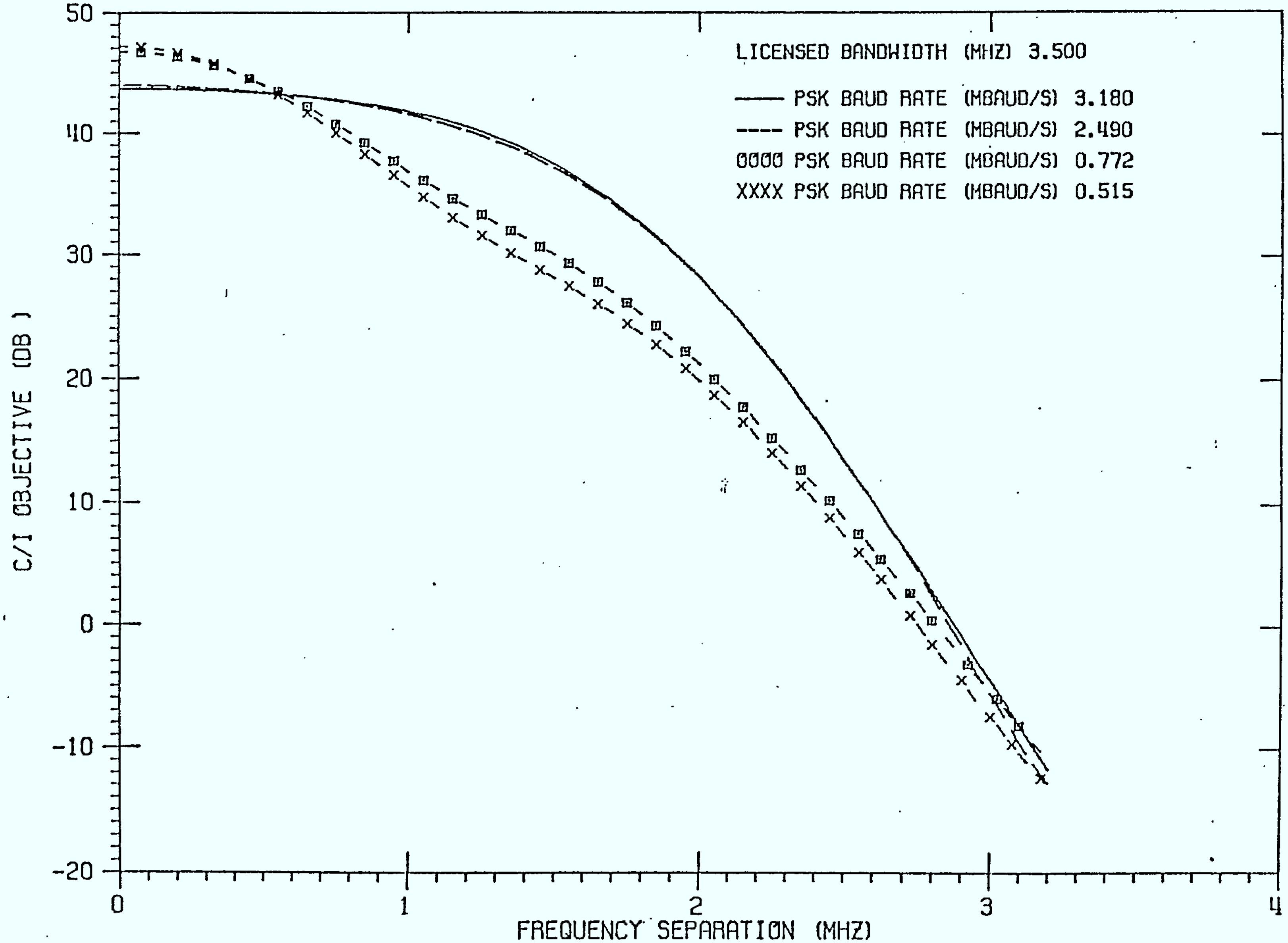


FIGURE G-1(B) C/I OBJECTIVE VS. FREQUENCY SEPARATION FOR 6 CHANNEL LOADING FOR UPPER... [illegible]

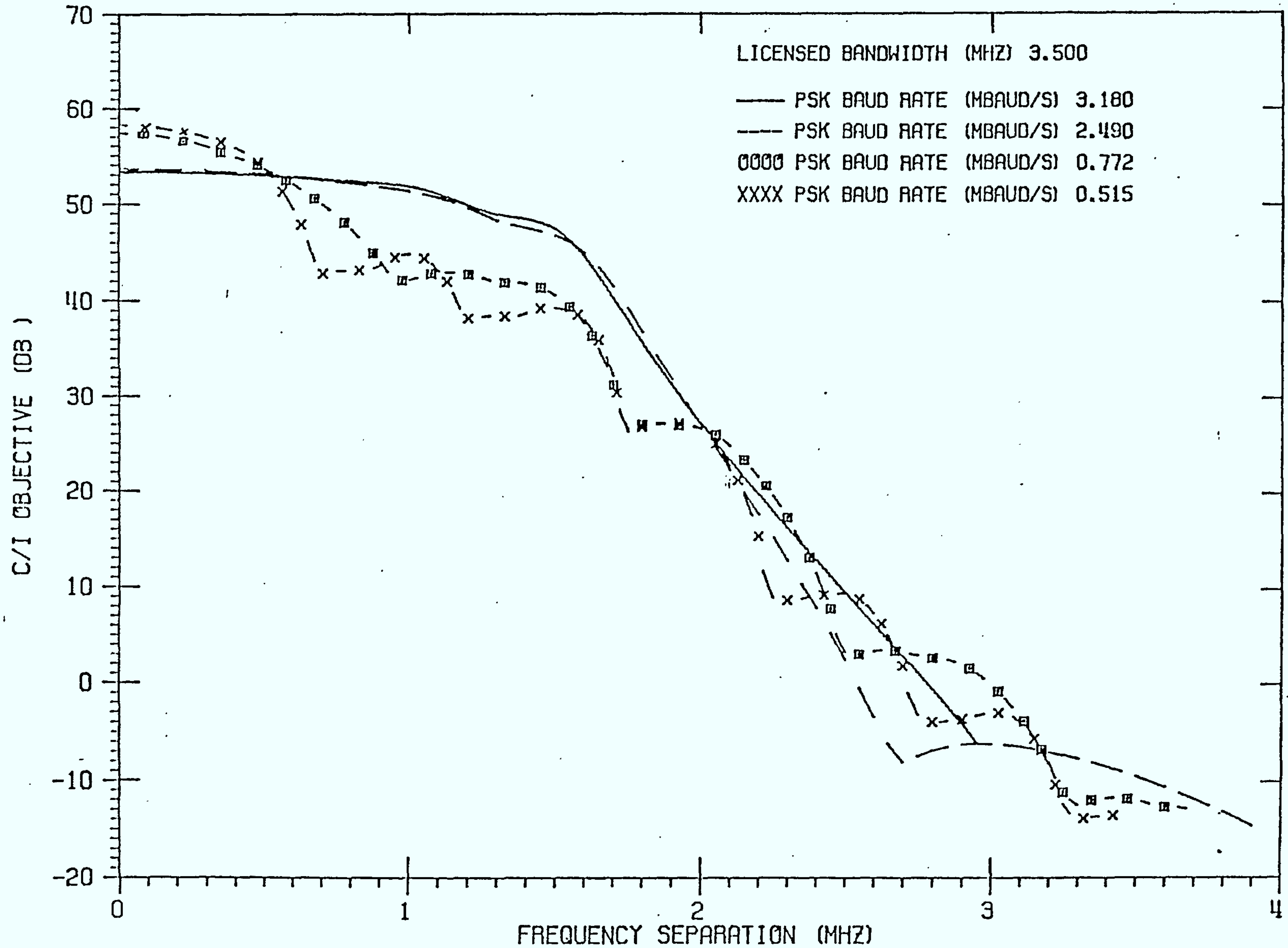


FIGURE G-1(c) C/I OBJECTIVE VS. FREQUENCY SEPARATION FOR O

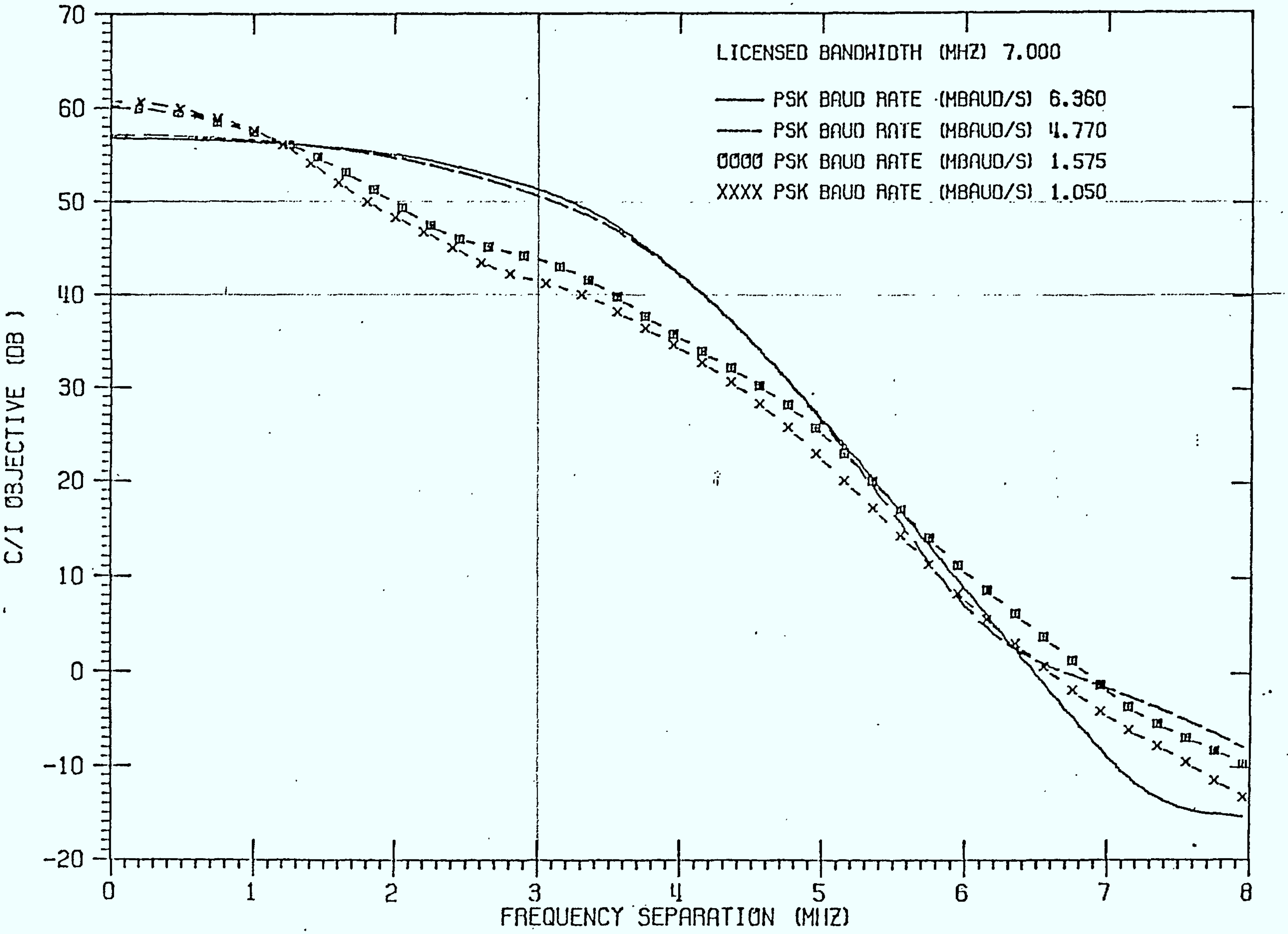


FIGURE G-2(A) C/I OBJECTIVE VS. FREQUENCY SEPARATION FOR 120

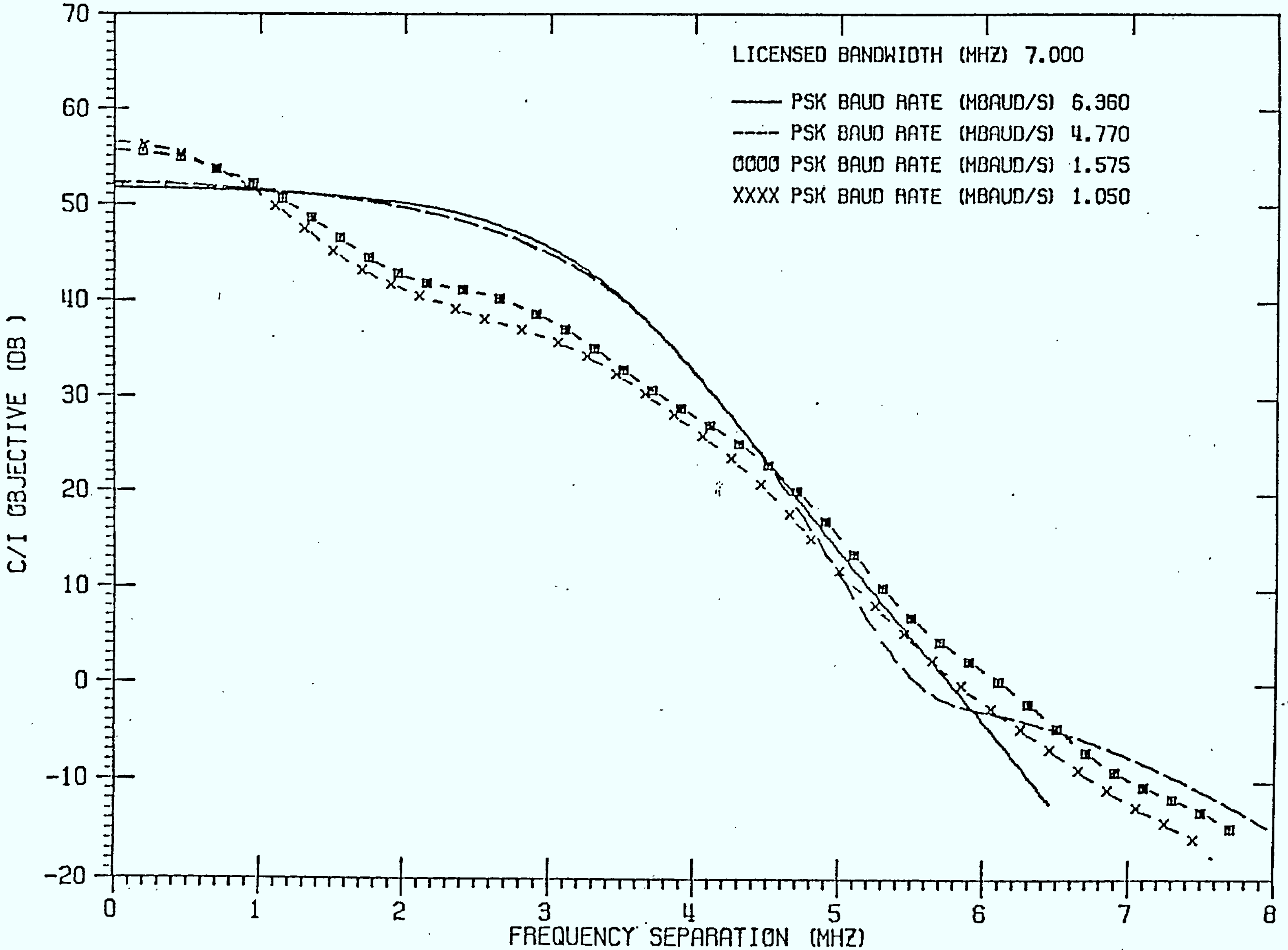


FIGURE G-2(B) C/I OBJECTIVE VS. FREQUENCY SEPARATION FOR 60

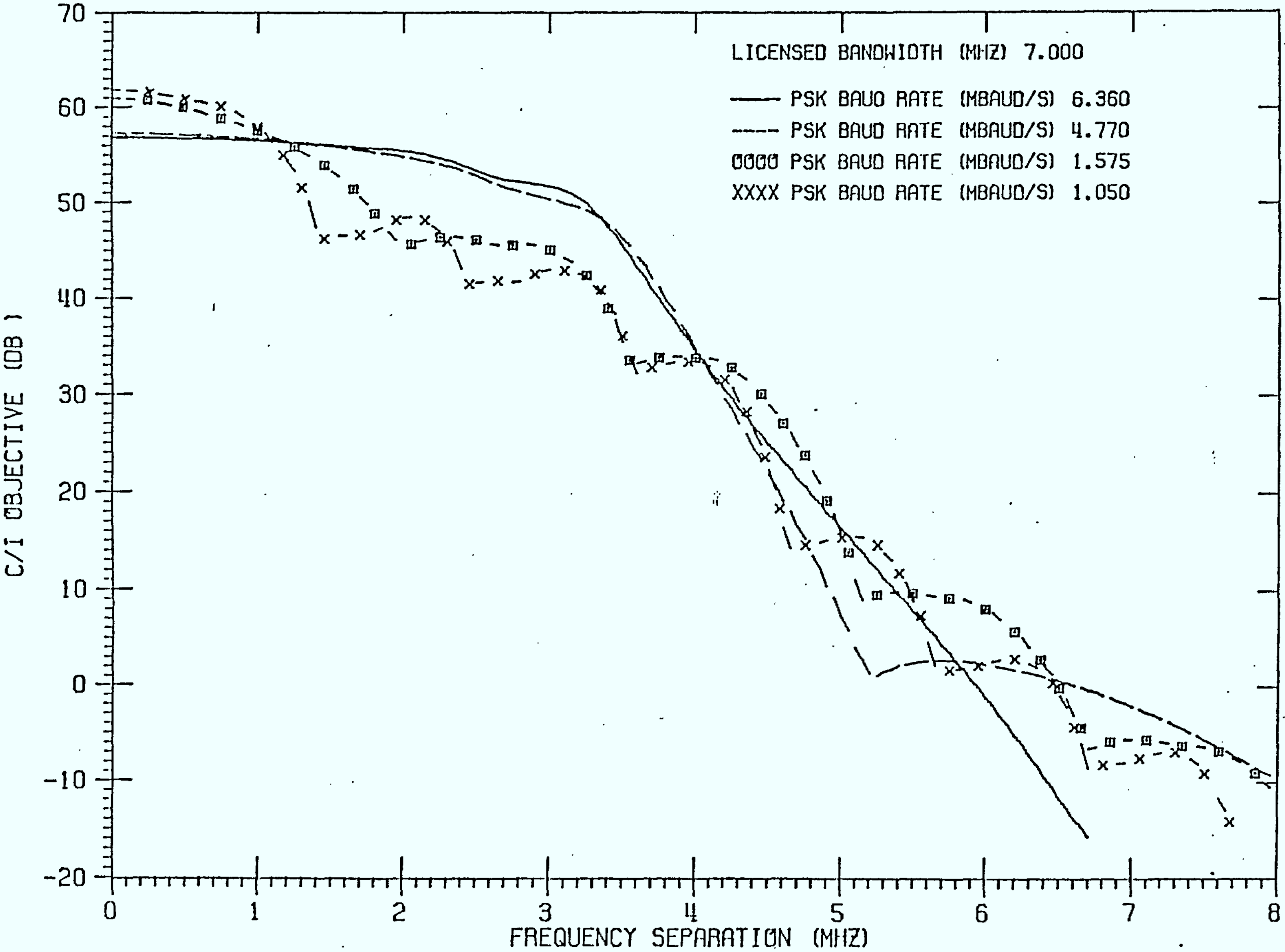


FIGURE G-2(c)

C/I OBJECTIVE VS. FREQUENCY SEPARATION FOR 0 CHANNEL LOADING FOR URBAN NOISE

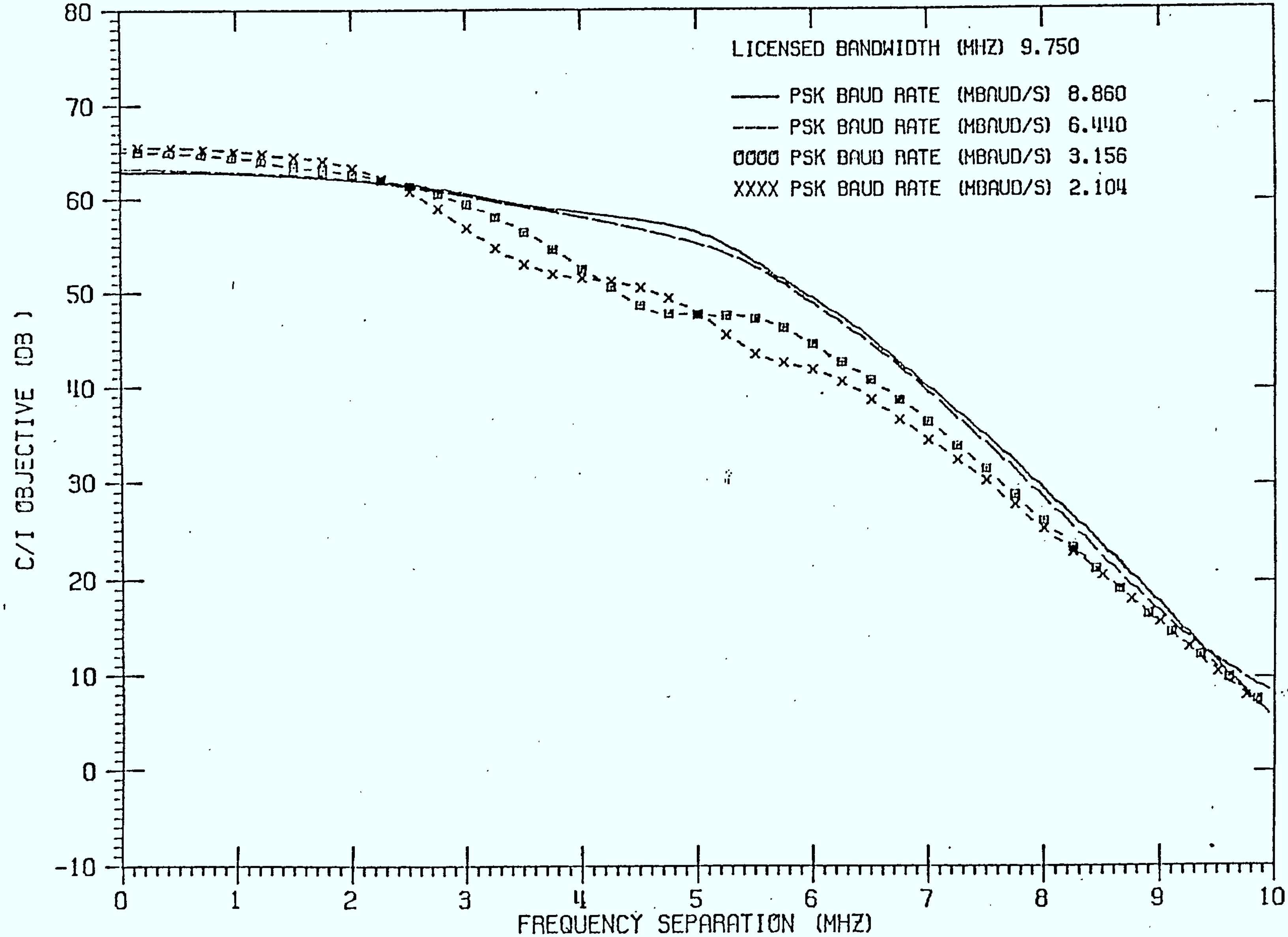
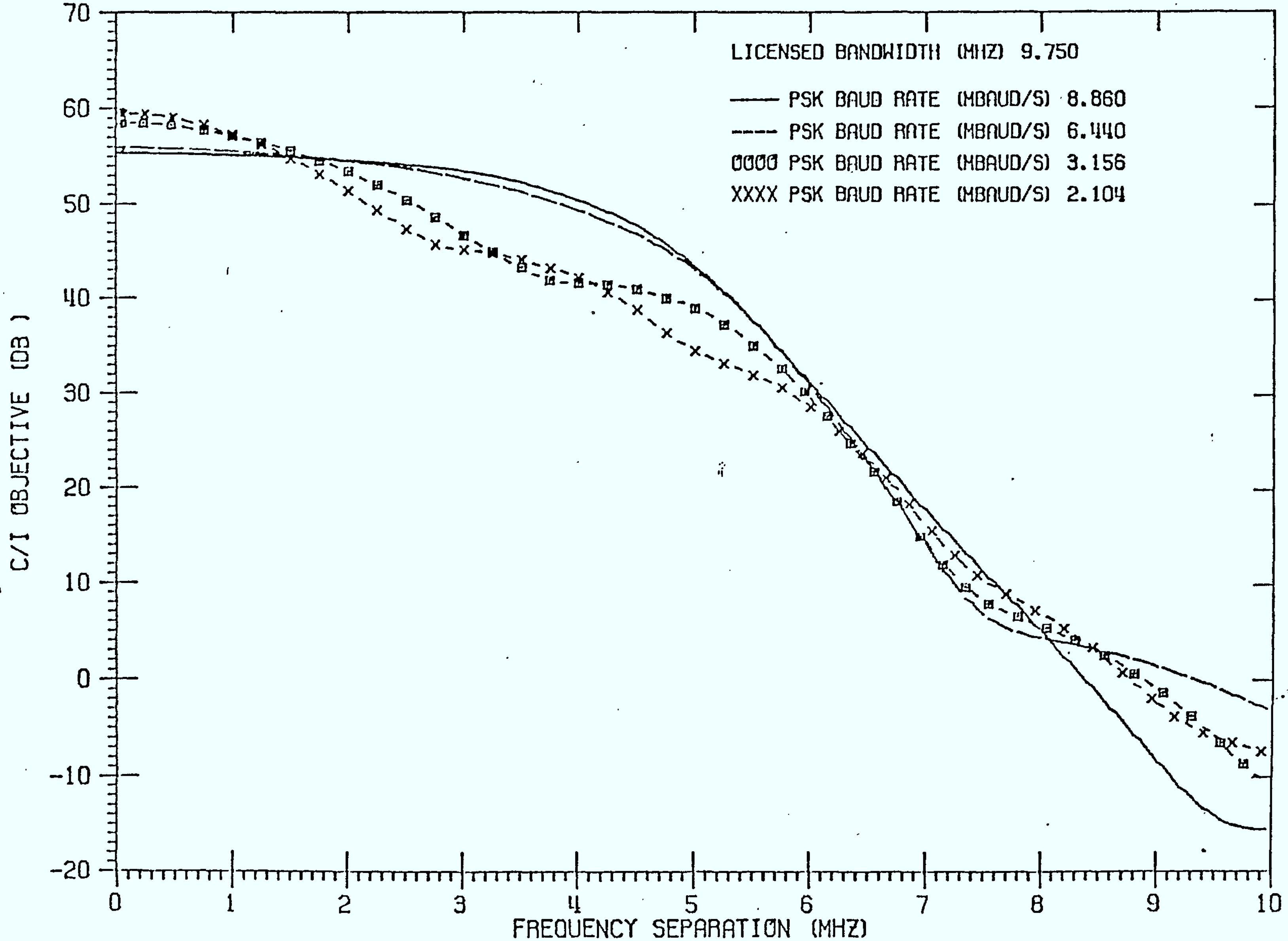


FIGURE G-3(A) C/I OBJECTIVE VS. FREQUENCY SEPARATION FOR 300



FIGURE

C/I OBJECTIVE VS. FREQUENCY SEPARATION FOR 120

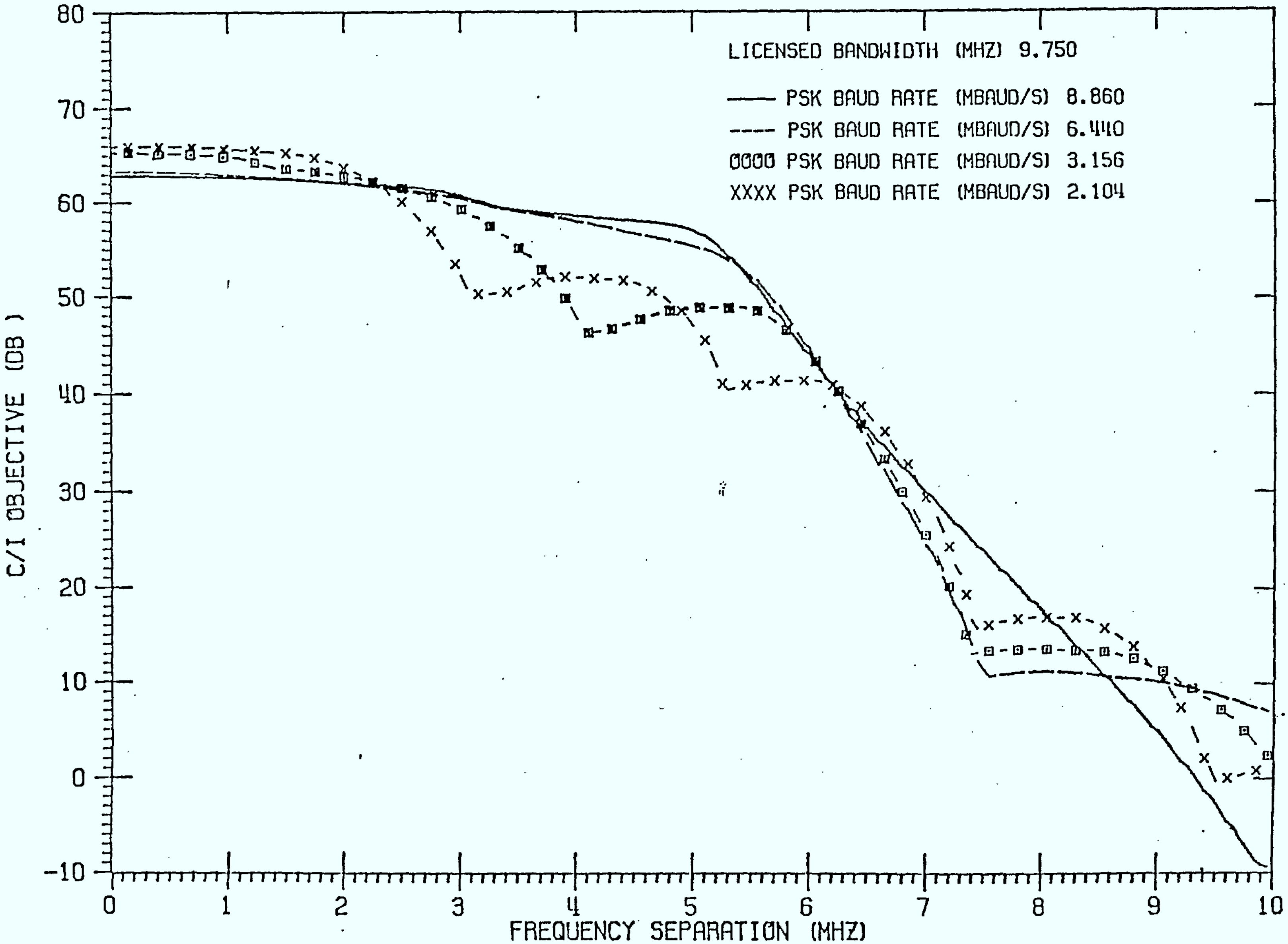


FIGURE G-3(c) C/I OBJECTIVE VS. FREQUENCY SEPARATION FOR 0 CHANNEL LOADING FOR UDRRNCA NOISE

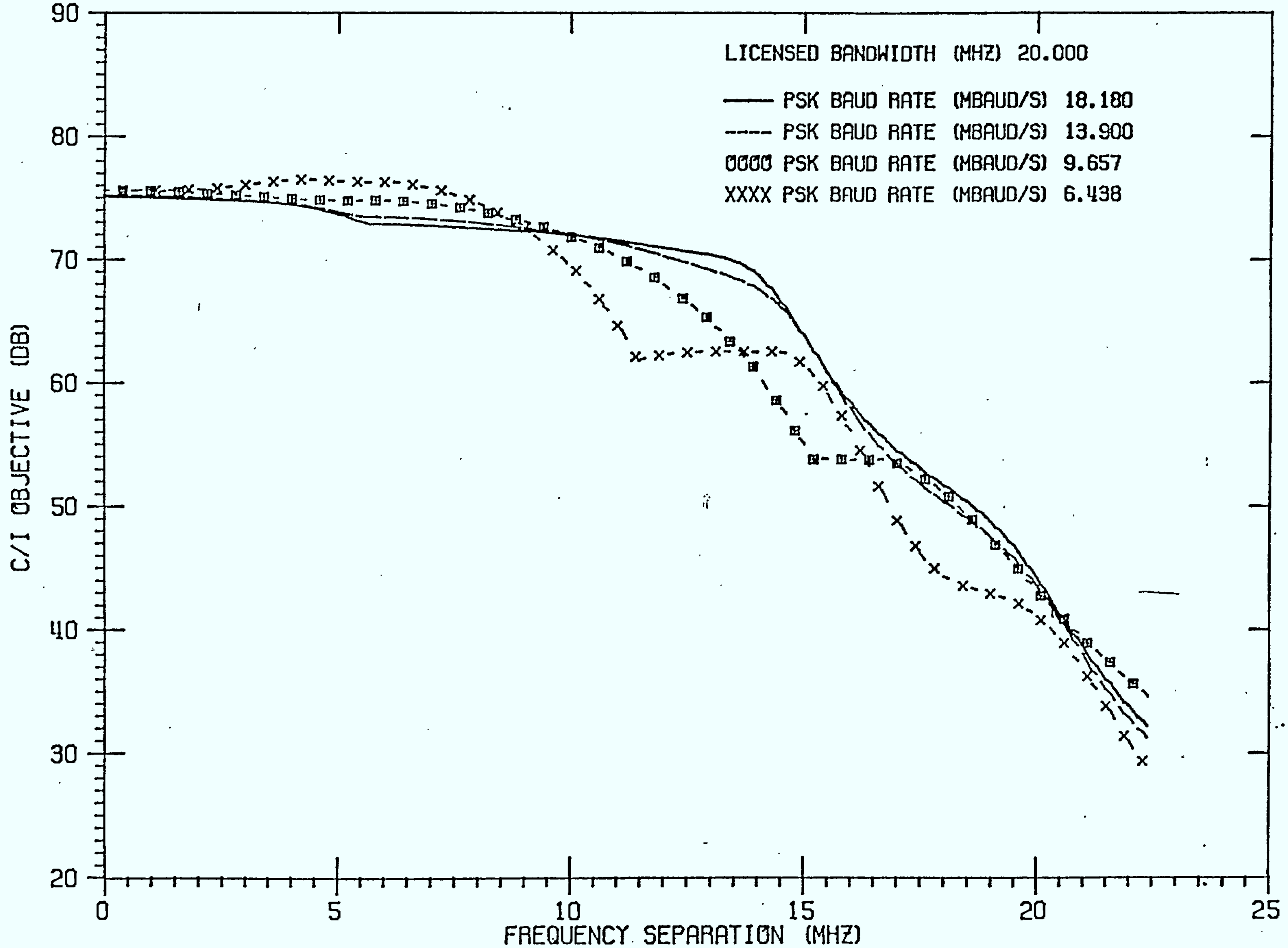


FIGURE G-4(A) C/I OBJECTIVE VS. FREQUENCY SEPARATION FOR 1260 CHANNEL LOADING FOR 4DBRNC0 NOISE

G-12

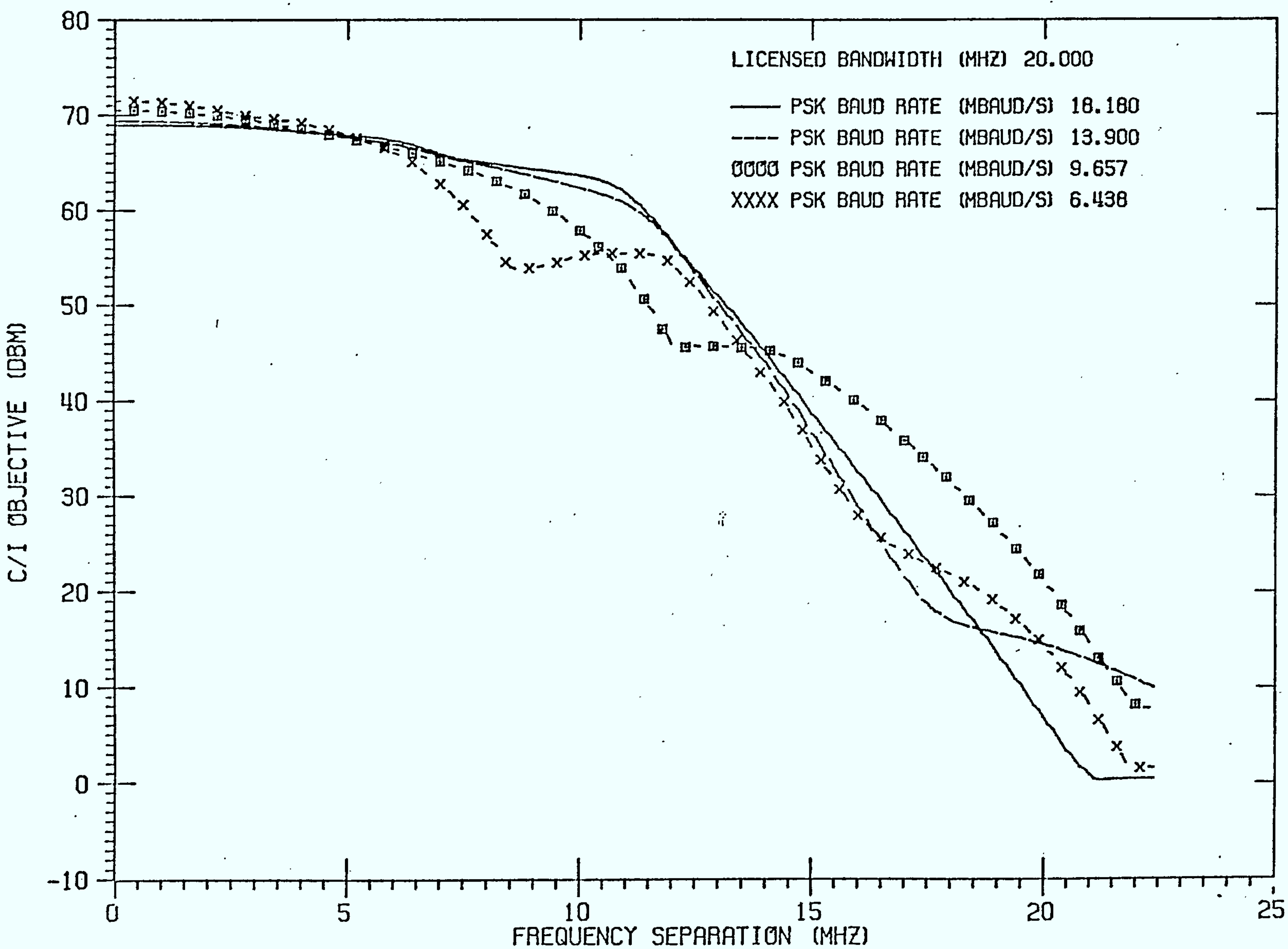


FIGURE G-4(B) C/I OBJECTIVE VS. FREQUENCY SEPARATION FOR 600 CHANNEL LOADING FOR 4DBRNC0 NOISE

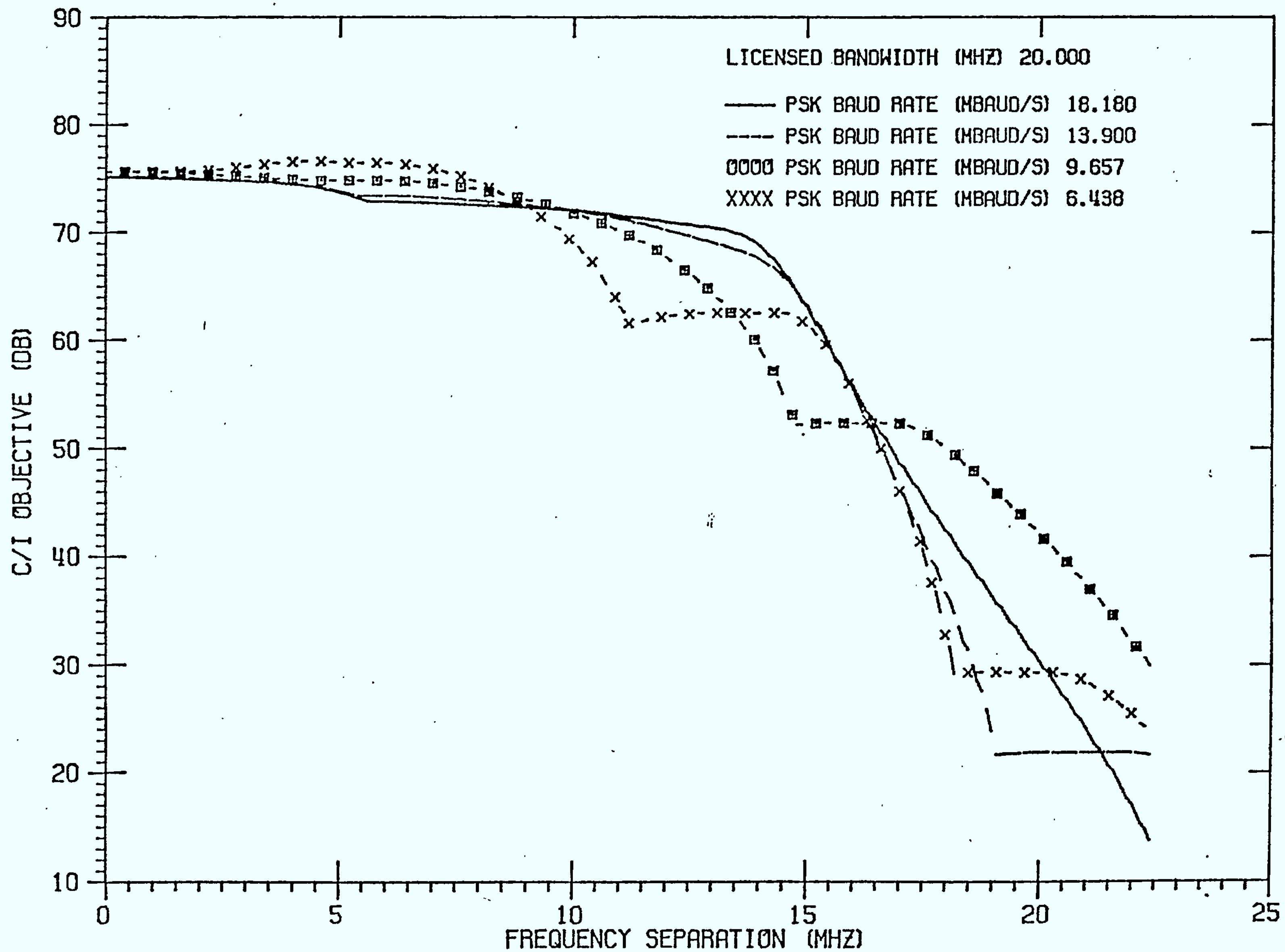
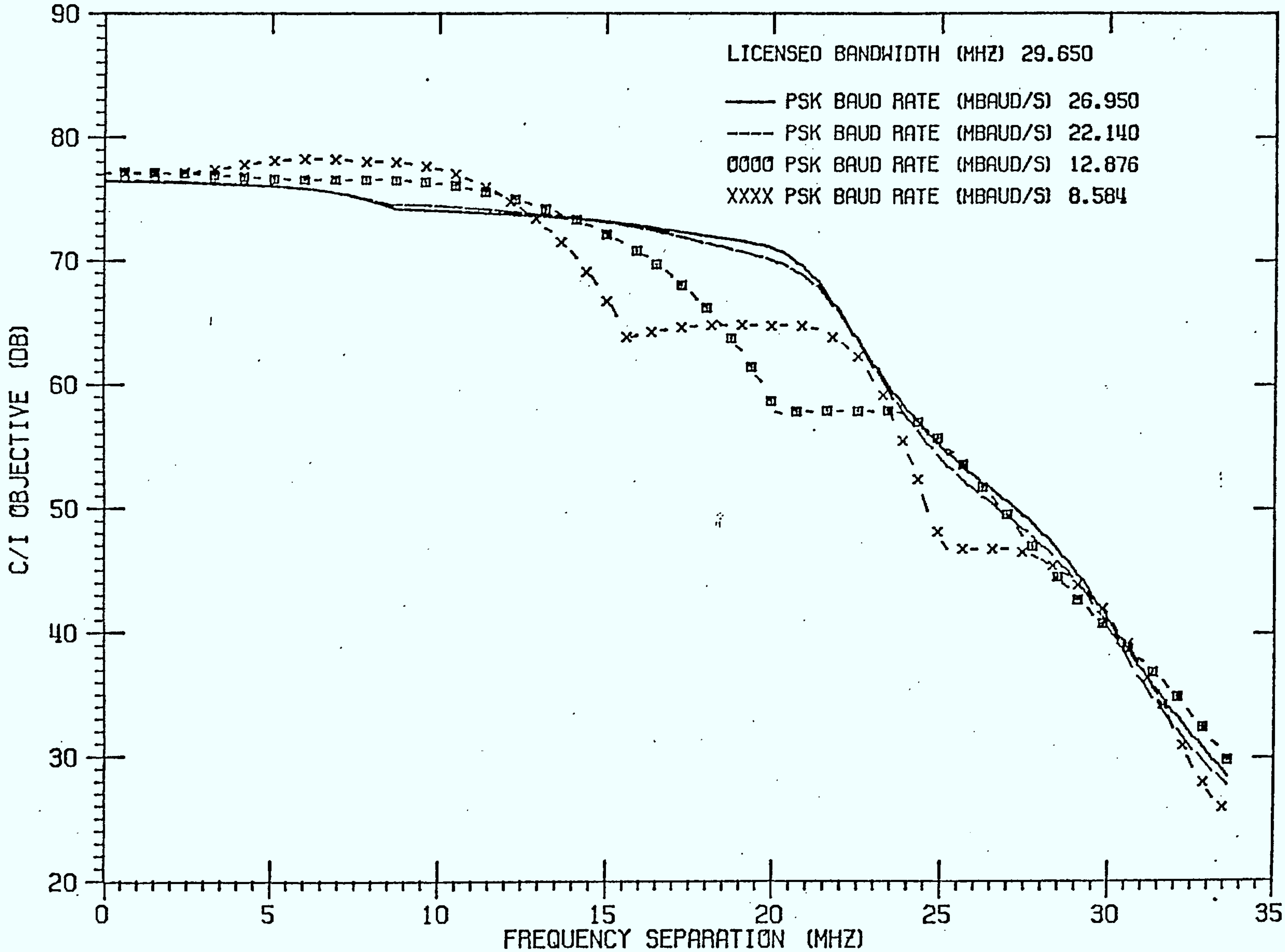


FIGURE G-4(c)

C/I OBJECTIVE VS. FREQUENCY SEPARATION FOR 0 CHANNEL LOADING FOR 4DBRNC0 NOISE



C/I = 40
→ 84

FIGURE G-5(A) C/I OBJECTIVE VS. FREQUENCY SEPARATION FOR 1800 CHANNEL LOADING FOR 40DBRNC0 NOISE

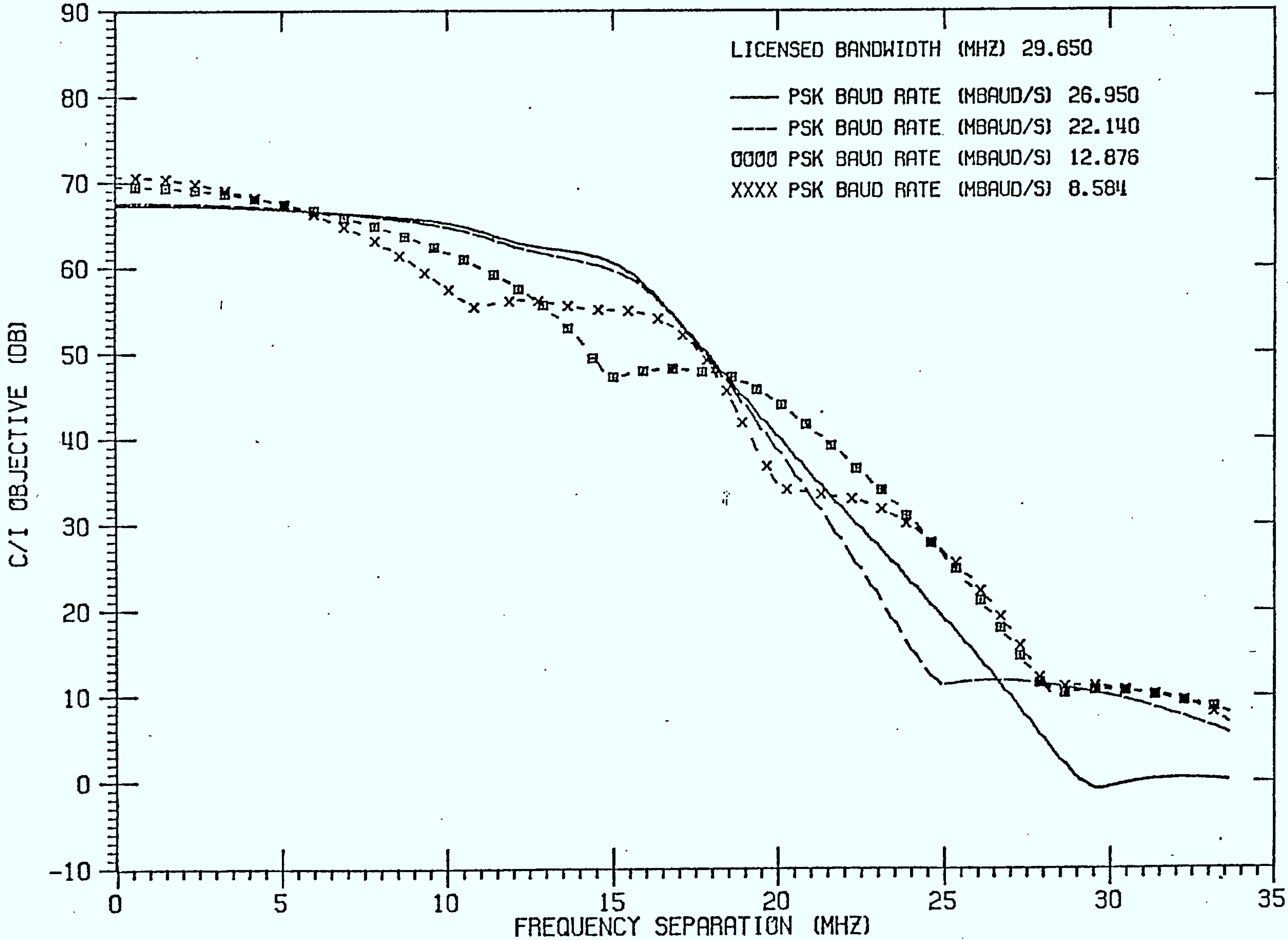


FIGURE
G-5(B)

C/I OBJECTIVE VS. FREQUENCY SEPARATION FOR 600
CHANNEL LOADING FOR 4DBRNCO NOISE

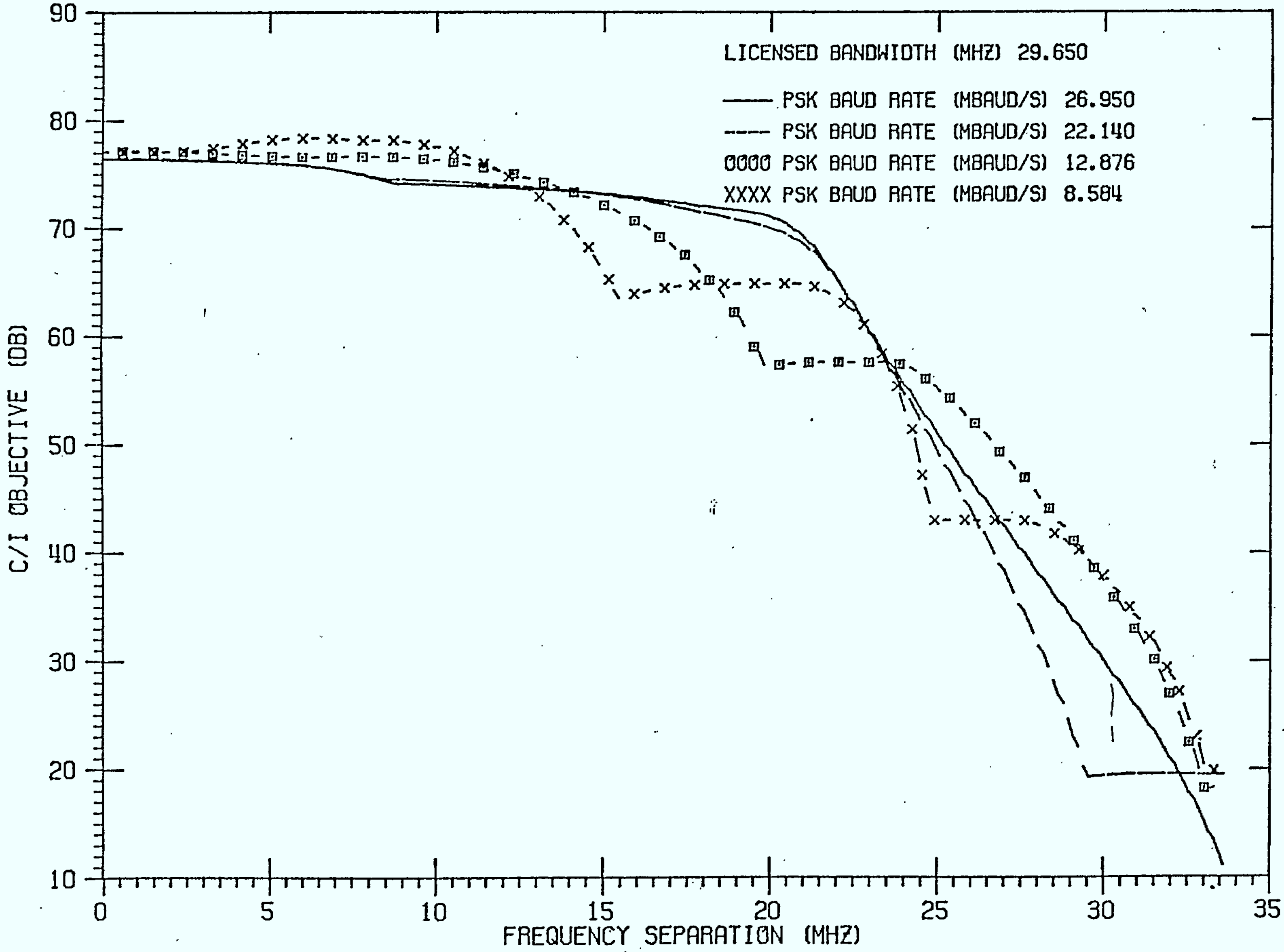


FIGURE G-5(c) C/I OBJECTIVE VS. FREQUENCY SEPARATION FOR 0 CHANNEL LOADING FOR 4DBRNCO NOISE

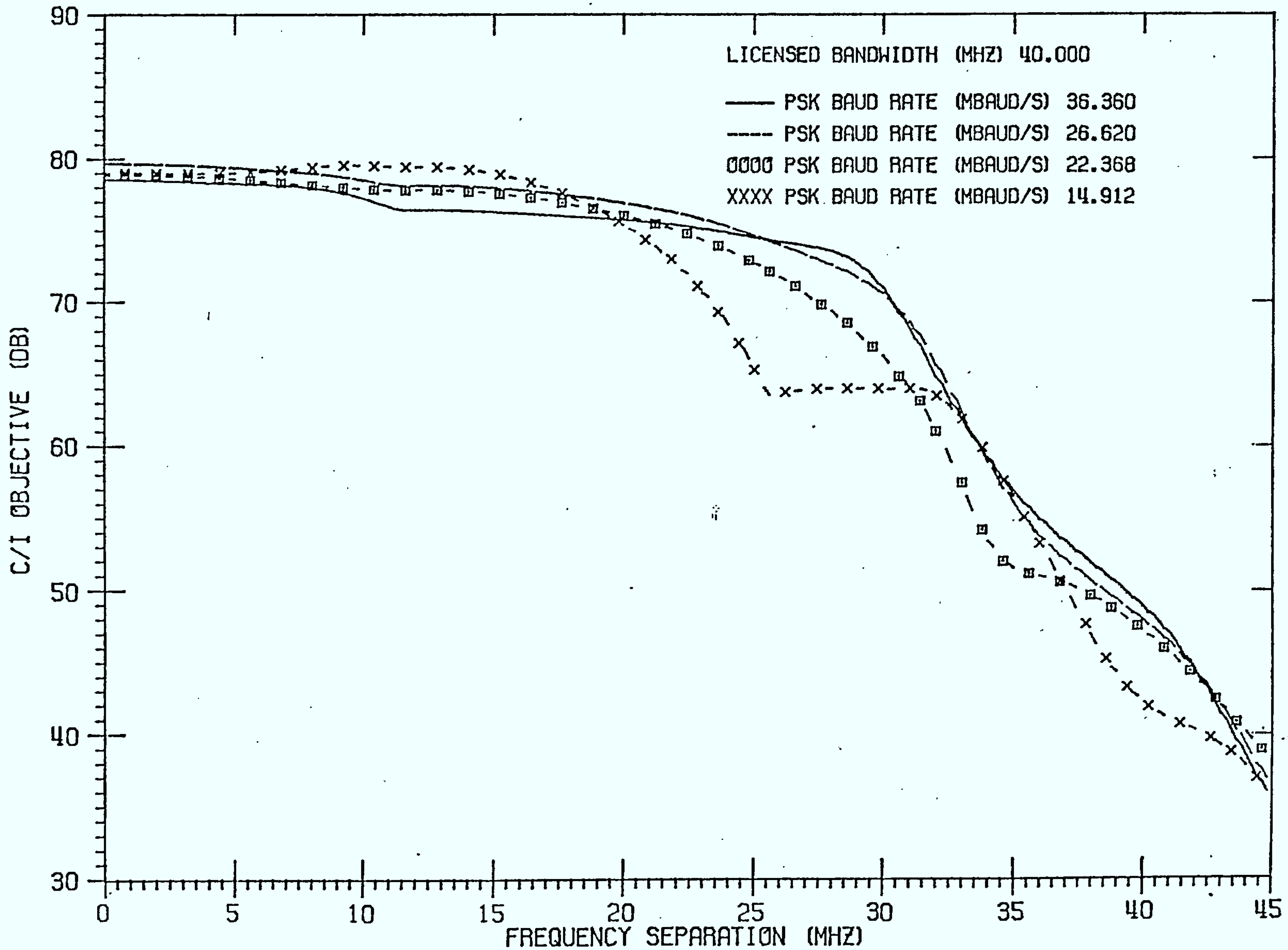


FIGURE G-6(A) C/I OBJECTIVE VS. FREQUENCY SEPARATION FOR 2700 CHANNEL LOADING FOR 4DBRNCO NOISE

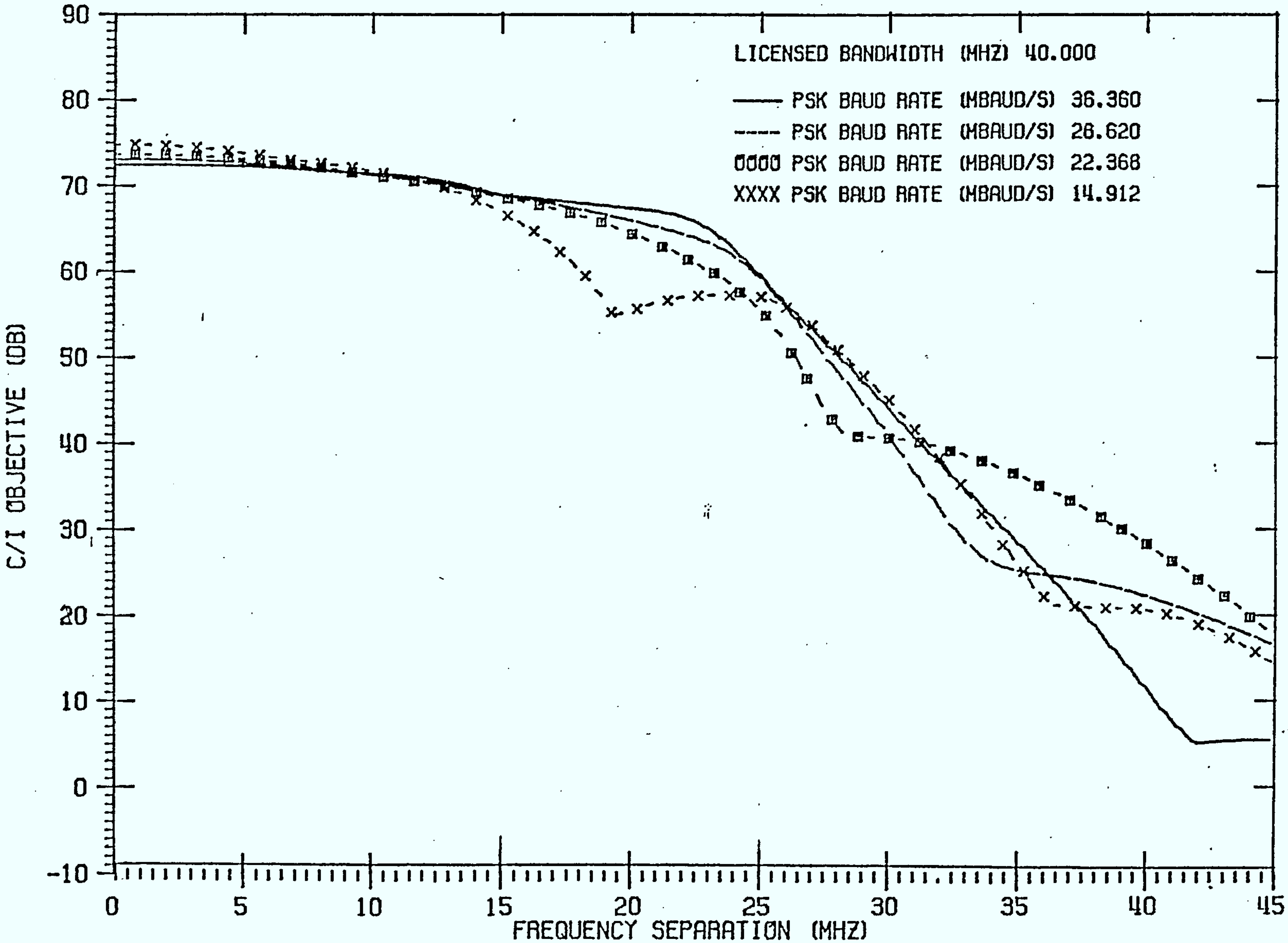


FIGURE G-6(B) C/I OBJECTIVE VS. FREQUENCY SEPARATION FOR 1260 CHANNEL LOADING FOR 4DBRNCO NOISE

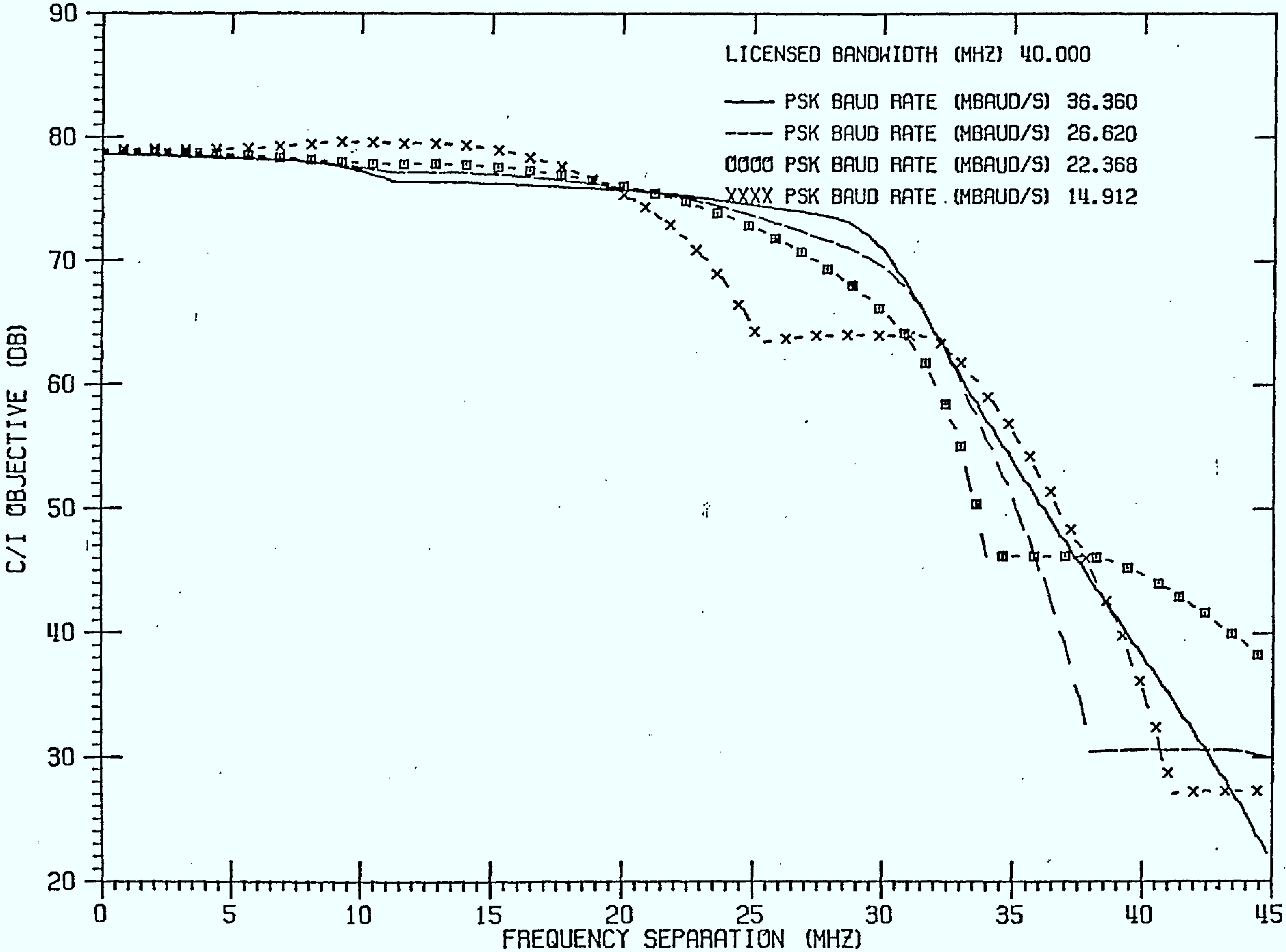


FIGURE G-6(c) C/I OBJECTIVE VS. FREQUENCY SEPARATION FOR 0 CHANNEL LOADING FOR 4DBRNC0 NOISE

APPENDIX H

Figures H-1 through H-6 present the results of our study of 8-level, 0.125 mod-index, FSK interference into FDM/FM signals, in the form of C/I objective curves versus carrier separation frequency for the 4 dBrnc0 interference objective derived in Appendix F. There is one figure for each licensed bandwidth covered in the study, and also one FSK Baud rate for each licensed bandwidth. The Baud rates are given in Table 3, Appendix E, and are deliberately made equal to the maximum PSK Baud rates given in Table 2, Appendix E, in order to facilitate comparison of the FSK curves in this appendix with the PSK curves in Appendix G. It was because of the similarity between the FSK and PSK results that only the maximum Baud rate cases were run for the FSK interference. The three curves on each figure cover the three FDM/FM loadings for each licensed bandwidth, as given in Table 1, Appendix E. As in the case of the PSK curves in Appendix G, the unmodulated wanted channel curves represent the best-case, perfect-receiver-filtering, C/I objectives for the maximum FDM/FM loadings in the corresponding licensed bandwidths. The FSK figures are listed below (see Figures 1 and 3, Appendix E, for more complete details):

FIGURE	LICENSED BANDWIDTH MHz	FSK Baud Rate MBs	FDM/FM Channel Loadings
H-1	3.5	3.18	48 / 6 / 0
H-2	7.0	6.36	120 / 60 / 0
H-3	9.75	8.86	300 / 120 / 0
H-4	20.0	18.18	1260 / 600 / 0
H-5	29.65	26.95	1800 / 600 / 0
H-6	40.0	36.36	2700 / 1260 / 0

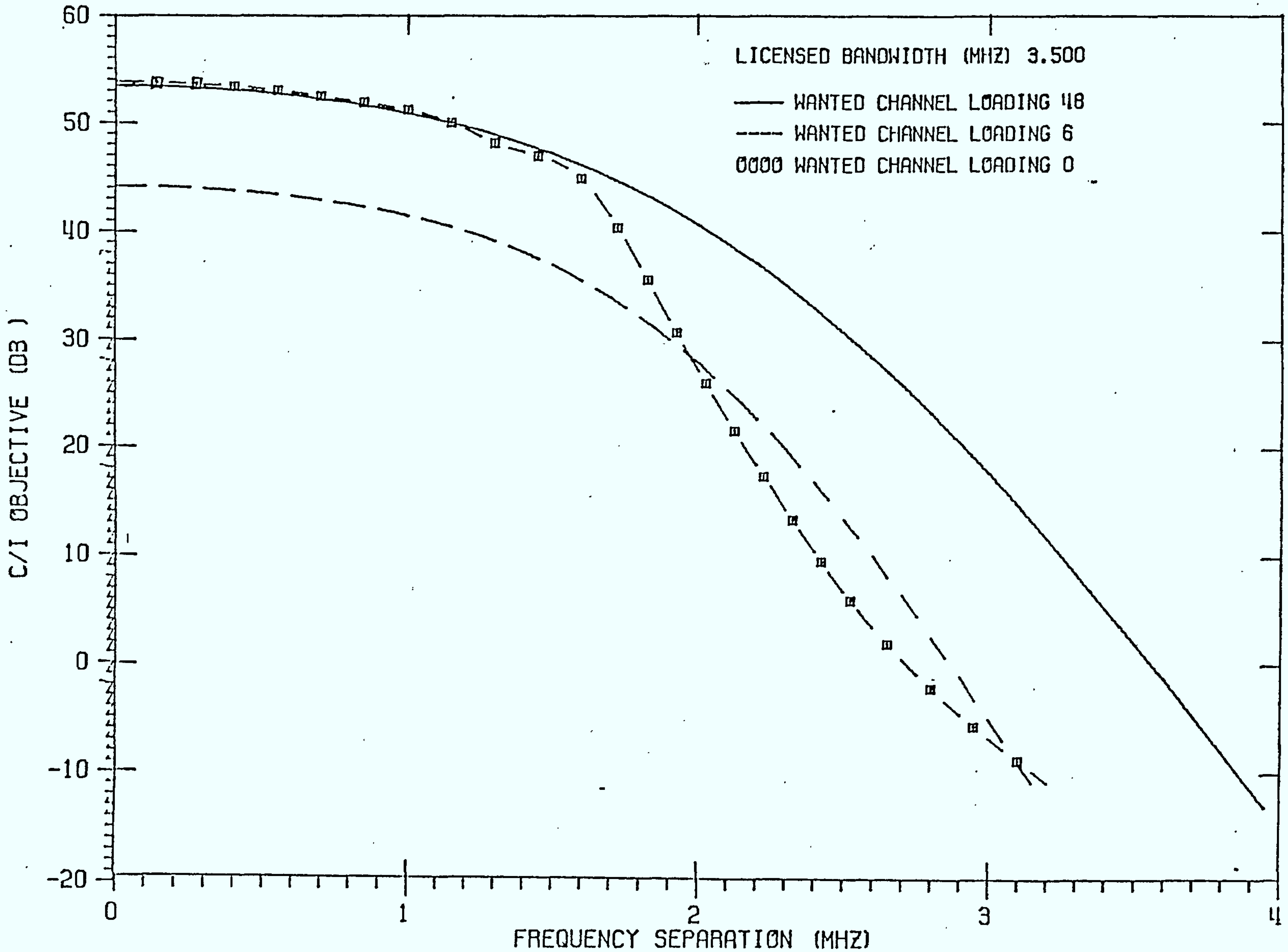


FIGURE H-1 C/I OBJECTIVE VS. FREQUENCY SEPARATION FOR 4 DBRNC0 NOISE FOR FSK INTERFERENCE BAUD RATE (MBAUD/S) 3.180

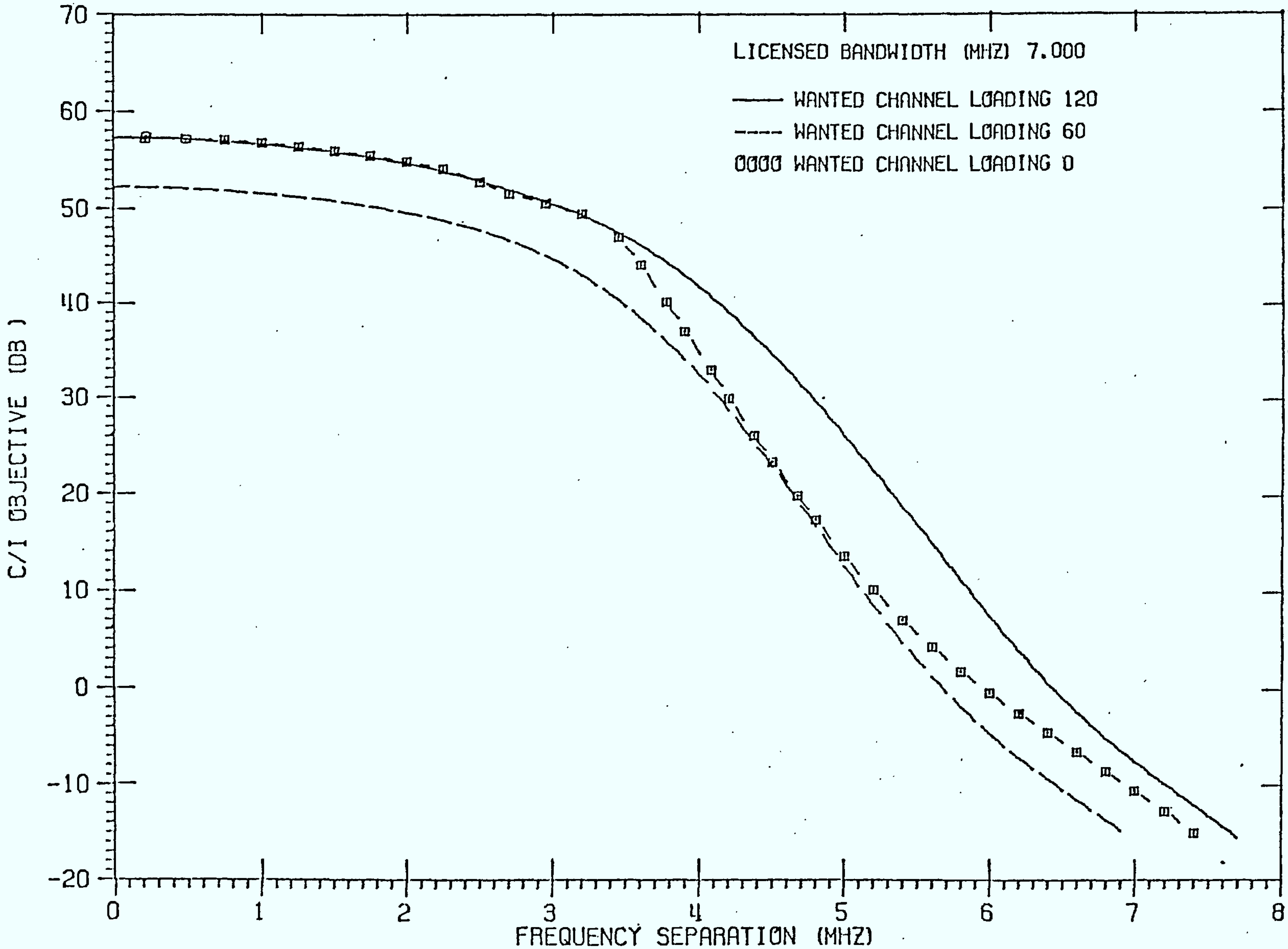


FIGURE H-2 C/I OBJECTIVE VS. FREQUENCY SEPARATION FOR 4 DBRNC0 NOISE FOR FSK INTERFERENCE BAUD RATE (MBAUD/S) 6.360

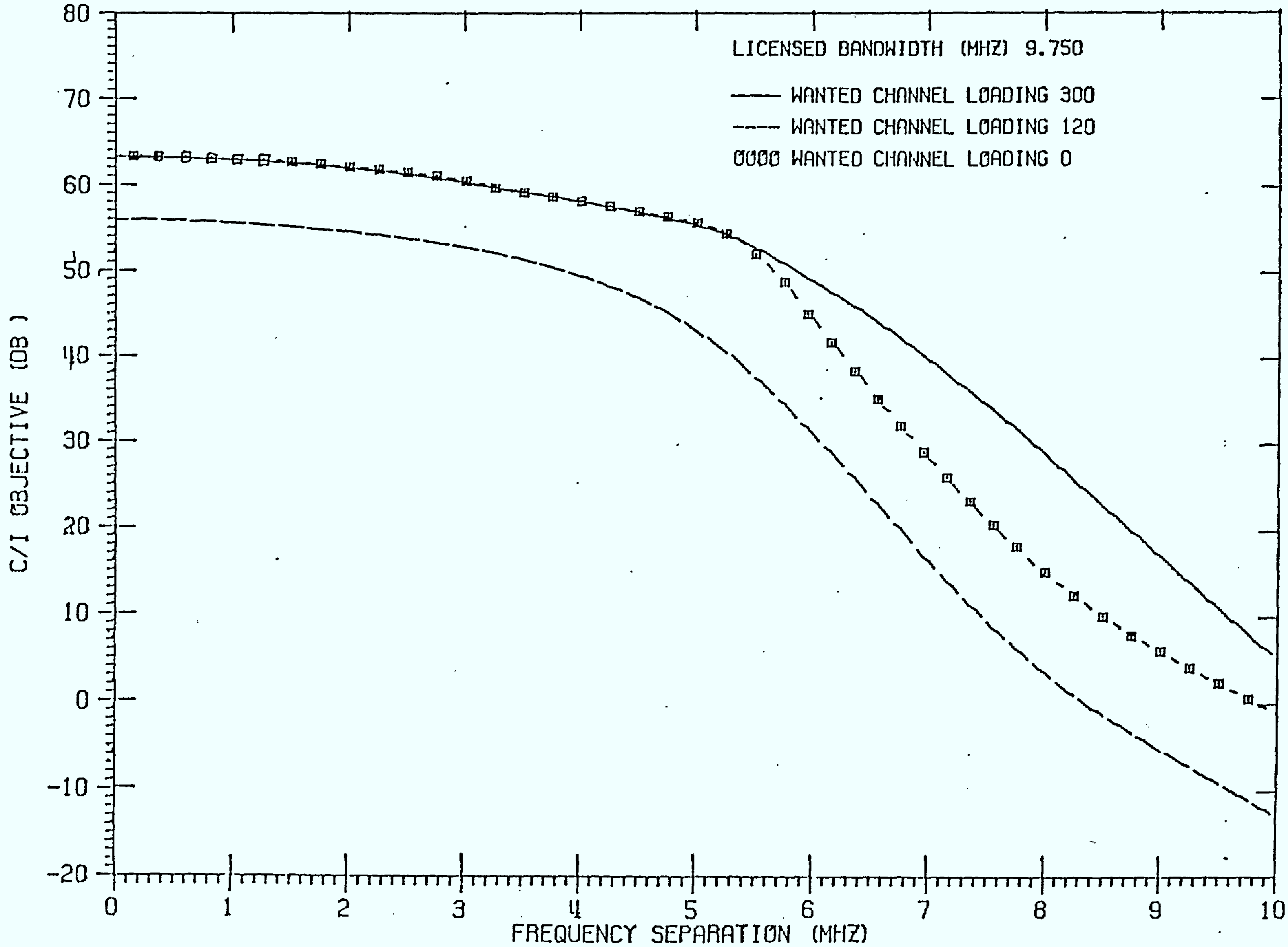


FIGURE H-3 C/I ØBJECTIVE VS. FREQUENCY SEPARATIØN FØR 4 DBRNCØ NØISE
FØR FSK INTERFERENCE BAUD RATE

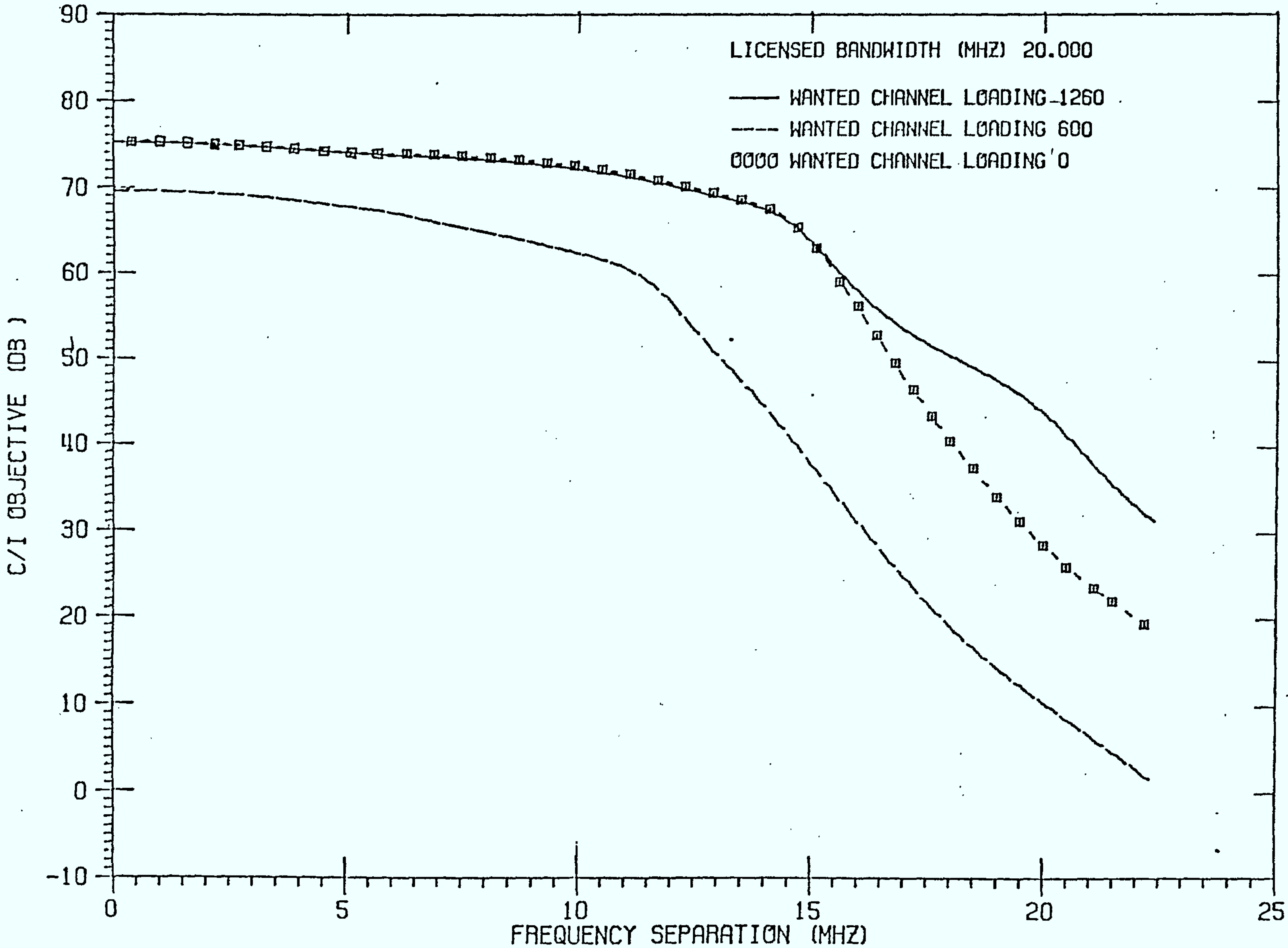


FIGURE II-4 C/I OBJECTIVE VS. FREQUENCY SEPARATION FOR 4 DBRNCO NOISE
FOR FSK [illegible]

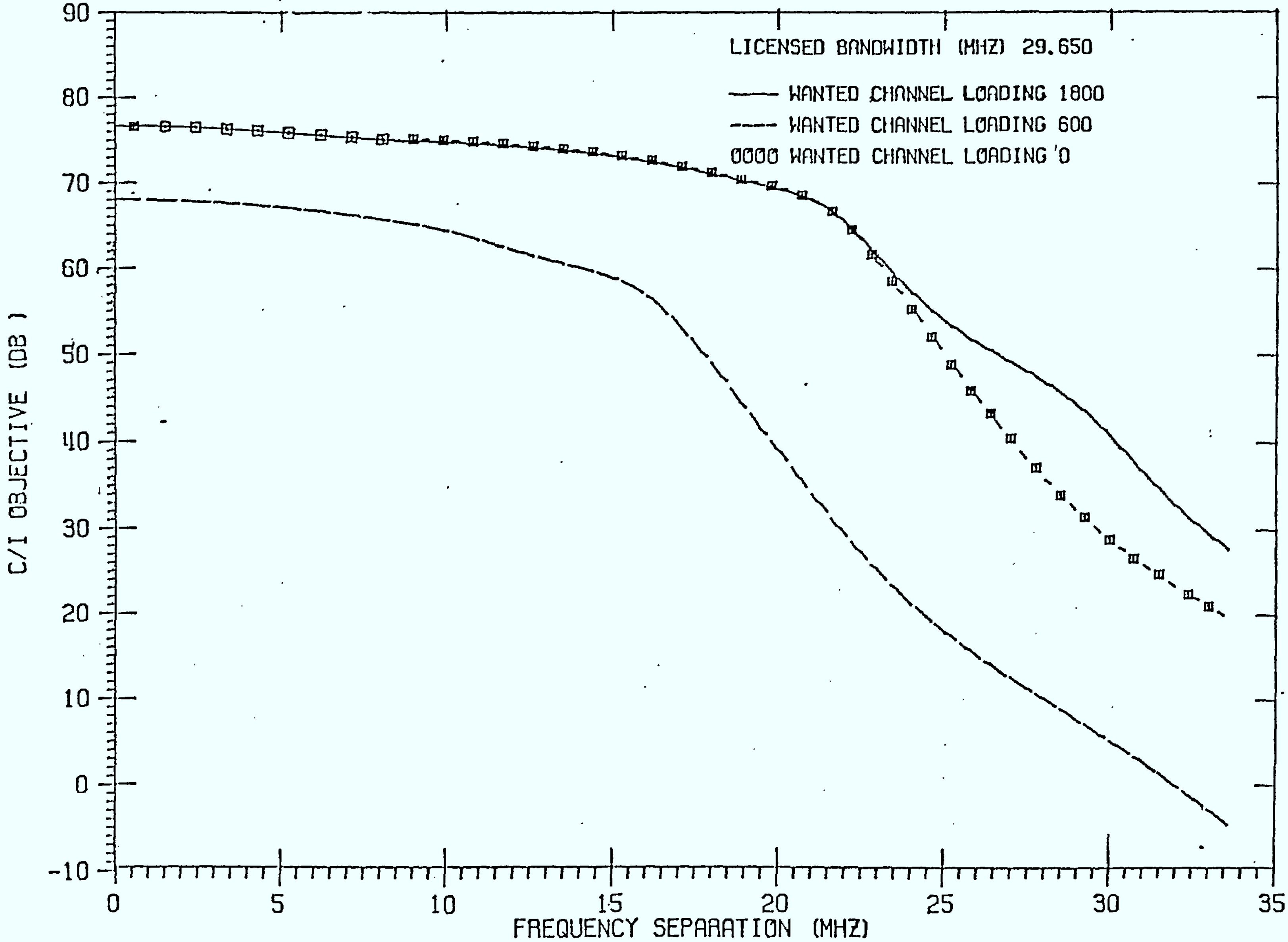


FIGURE C/I OBJECTIVE VS. FREQUENCY SEPARATION FOR 4 DBRNC0 NOISE

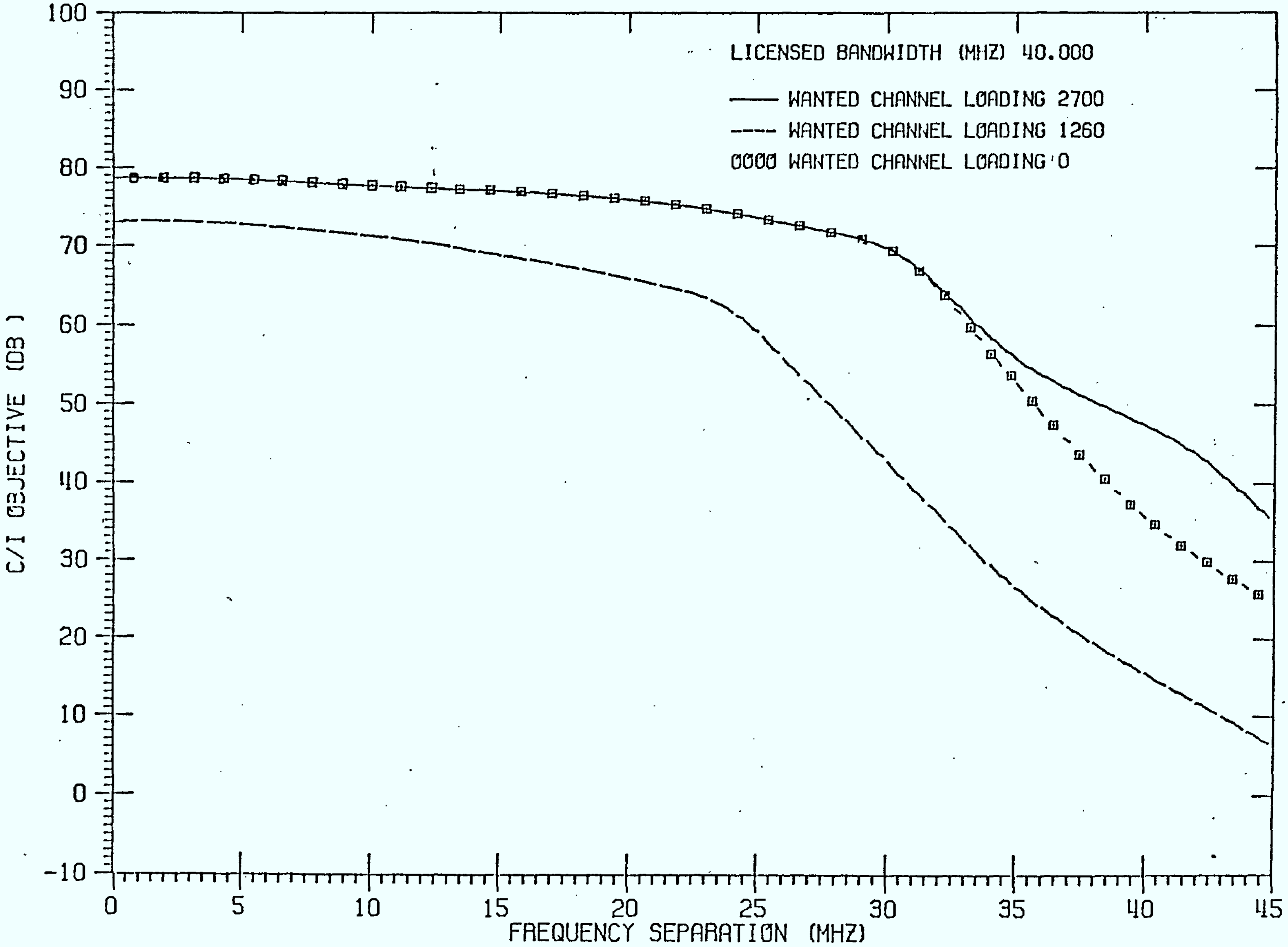


FIGURE
II-6

C/I ØBJECTIVE VS. FREQUENCY SEPARATIØN FØR 4 DBRNCØ NØISE
FØR FSK INTERFERENCE

APPENDIX I

QPRS INTO FDM/FM CURVES

As shown in section 2 of this report, and as discussed in Appendix E, only a single QPRS interferer was modeled for this study, namely the medium-capacity (91 Mb/s), Northern Telecom DRS-8 (8 GHz) digital radio system. The parameters used to perform this modelling, and the FDM/FM radio parameters used for the wanted channel signals, are given in the 4 QPRS section of Appendix E. The results of this study of QPRS into FDM/FM interference are presented in Figure I-1, in the form of C/I objectives versus carrier separation frequency for the 4 dBrnc0 interference objective derived in Appendix F. There are three curves in Figure I-1, one for each of the three FDM/FM loadings given in the QPRS section of Appendix E. As in the case of the PSK curves in Appendix G, the unmodulated wanted channel curve in Figure I-1, represents the best-case, perfect-receiver-filtering C/I objective for the 91 Mb/s QPRS interference into the maximum (1800 channel) FDM/FM loading for the given bandwidths (29.65 MHz FDM/FM bandwidth and 40.74 MHz QPRS bandwidth).

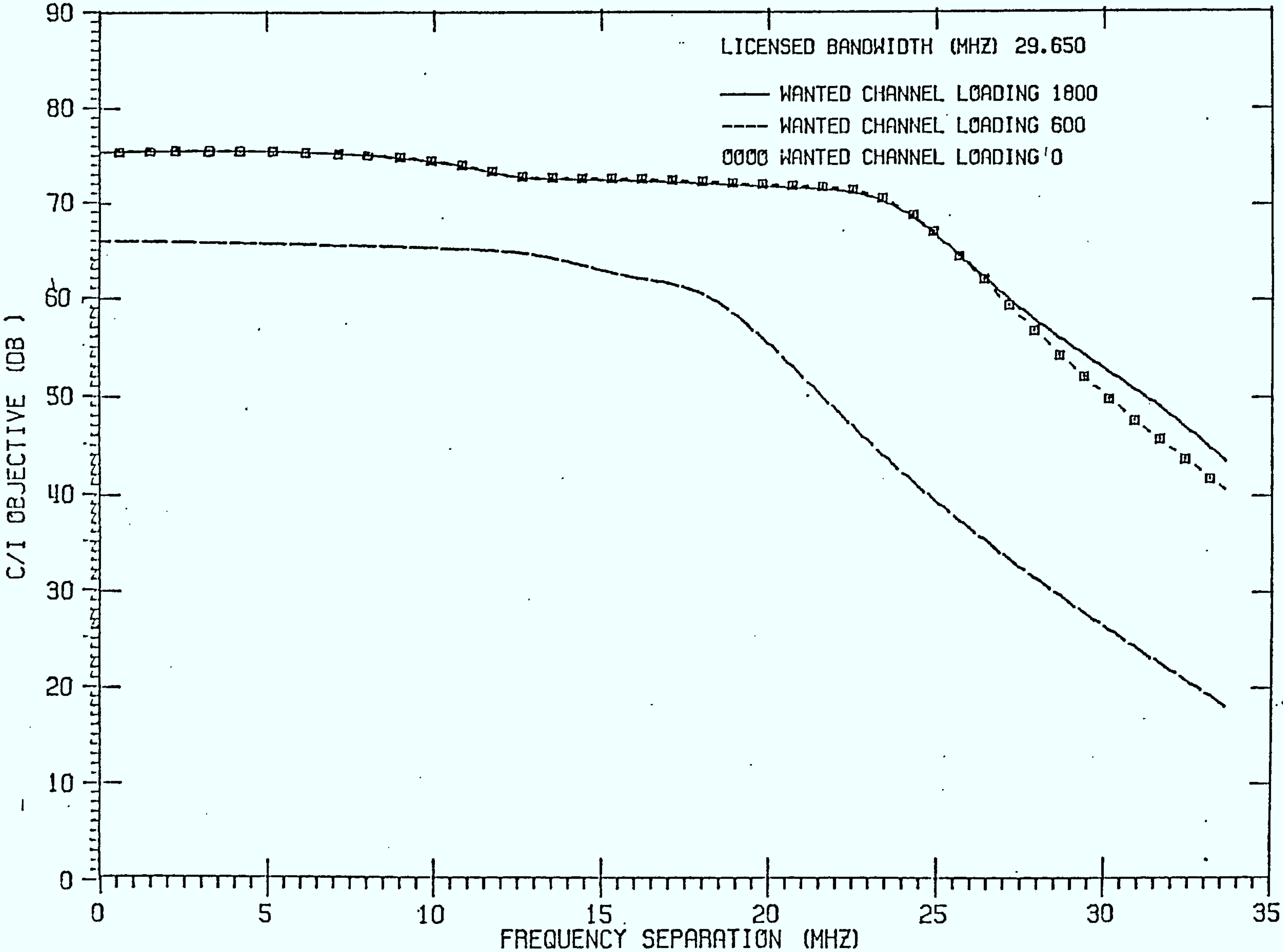


FIGURE I-1 C/I OBJECTIVE VS. FREQUENCY SEPARATION FOR 4 DBRNCO NOISE

APPENDIX J

PSK AND QPRS INTO VIDEO CURVES

APPENDIX J

As discussed in section 2 and 3 of this report, and in the Video Loading section of Appendix E, a stored colour bar spectrum was used for the wanted channel video signal, and the PSK signals for the 20 MHz and 29.65 MHz licensed bandwidths were used as interference, as well as the 40.74 MHz QPRS signal. The results of the study are presented in Figures J-1 through J-3 in the form of $\frac{C}{I}$ objectives versus carrier separation on frequency for the 90 dB video $\frac{S}{N}$ objective derived in Appendix F. Refer to Appendix E for the signal parameters. The figures are listed below.

Figure J-1 - 20 MHz bandwidth, 18.18/13.90/9.657/6.438 MBs PSK interference into video colour bar.

Figure J-2 - 29.65 MHz bandwidth, 26.95/22.140/12.876/8.584 MBs PSK interference into video colour bar.

Figure J-3 - 40.74 MHz bandwidth, 45.52 MBs QPRS interference into video colour bar.

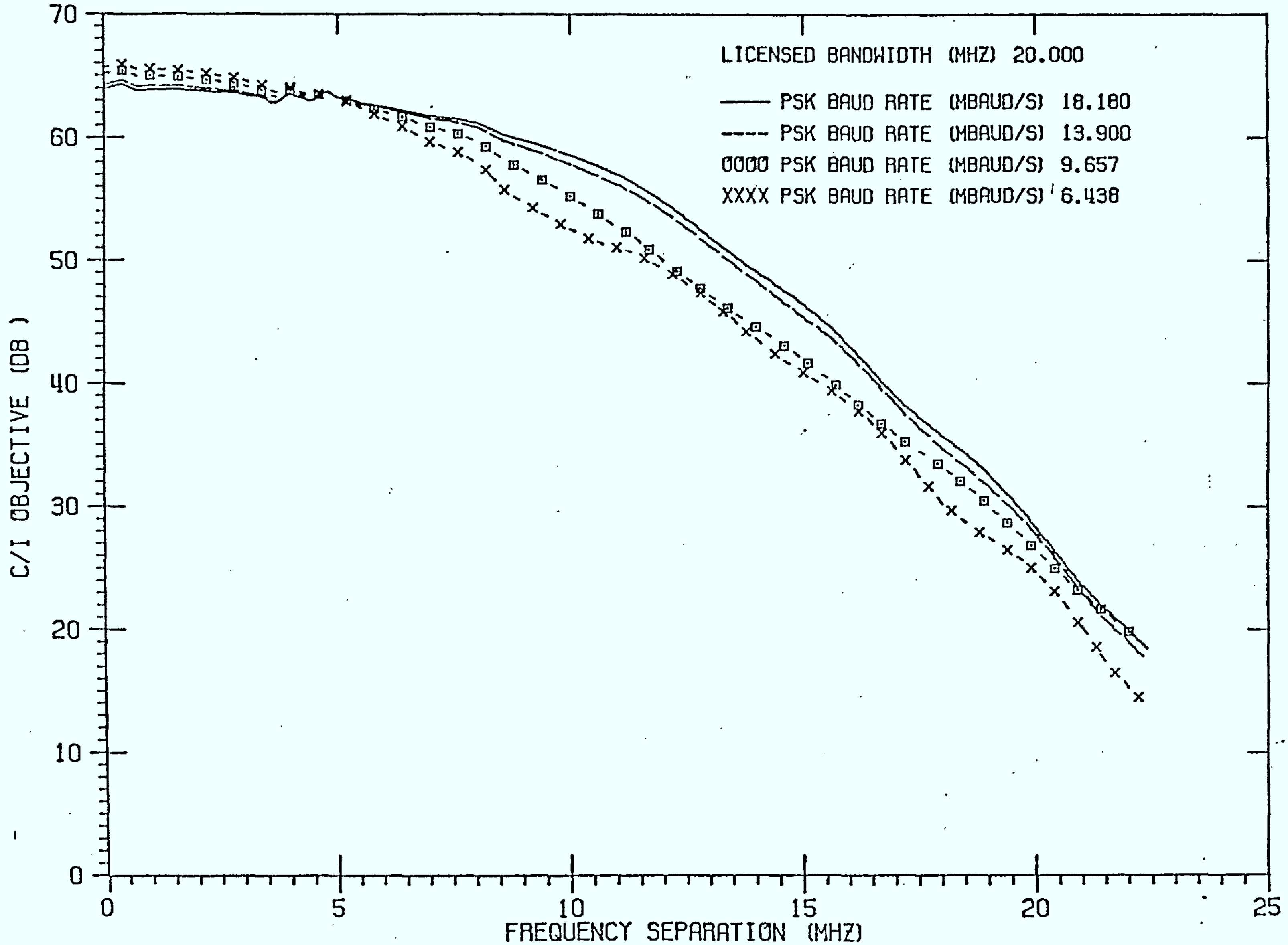


FIGURE J-1 C/I OBJECTIVE VS. FREQUENCY SEPARATION FOR 90DB WEIGHTED VIDEO S/N

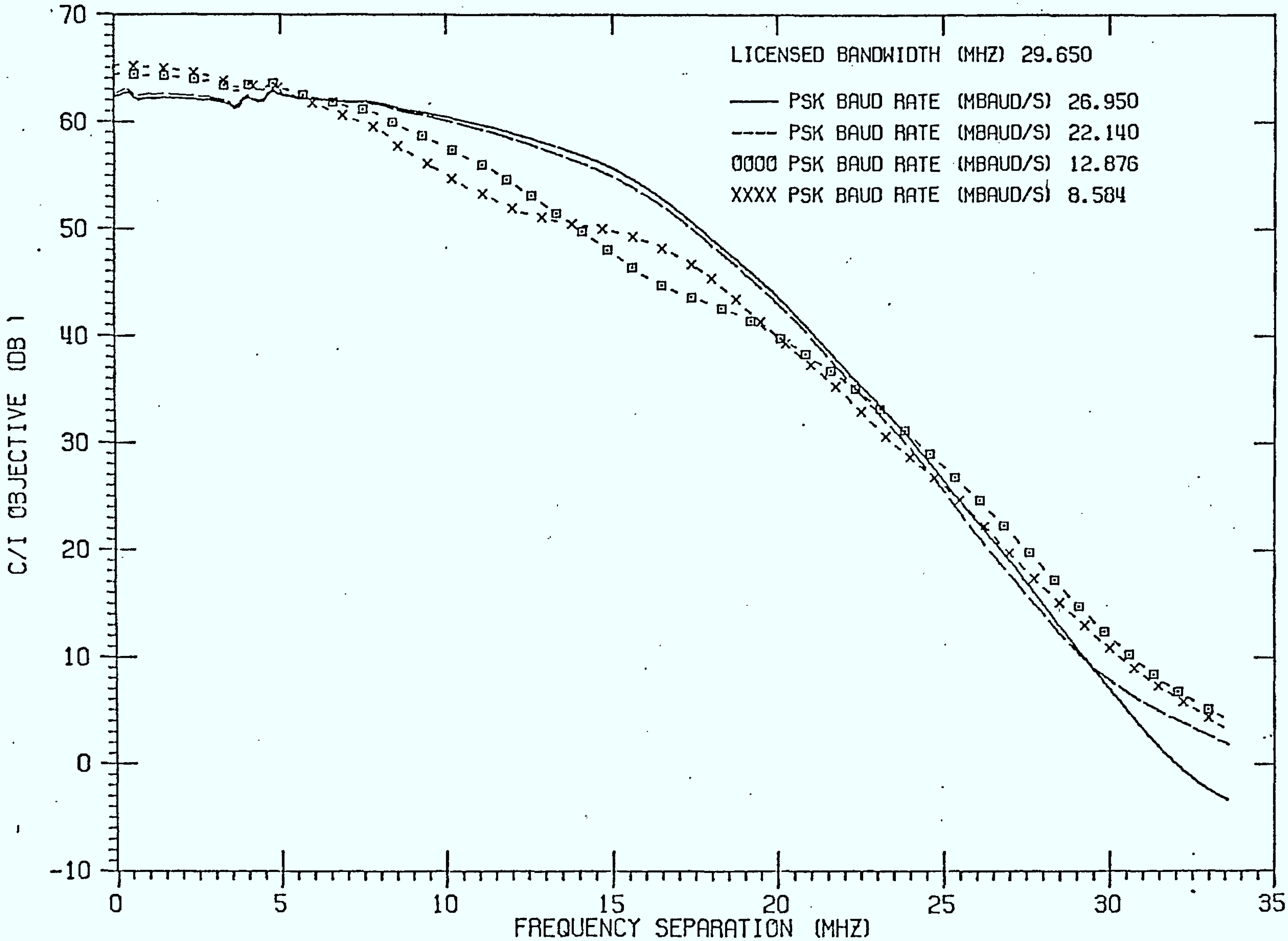


FIGURE J-2 C/I OBJECTIVE VS. FREQUENCY SEPARATION FOR 90DB WEIGHTED VIDEO S/N

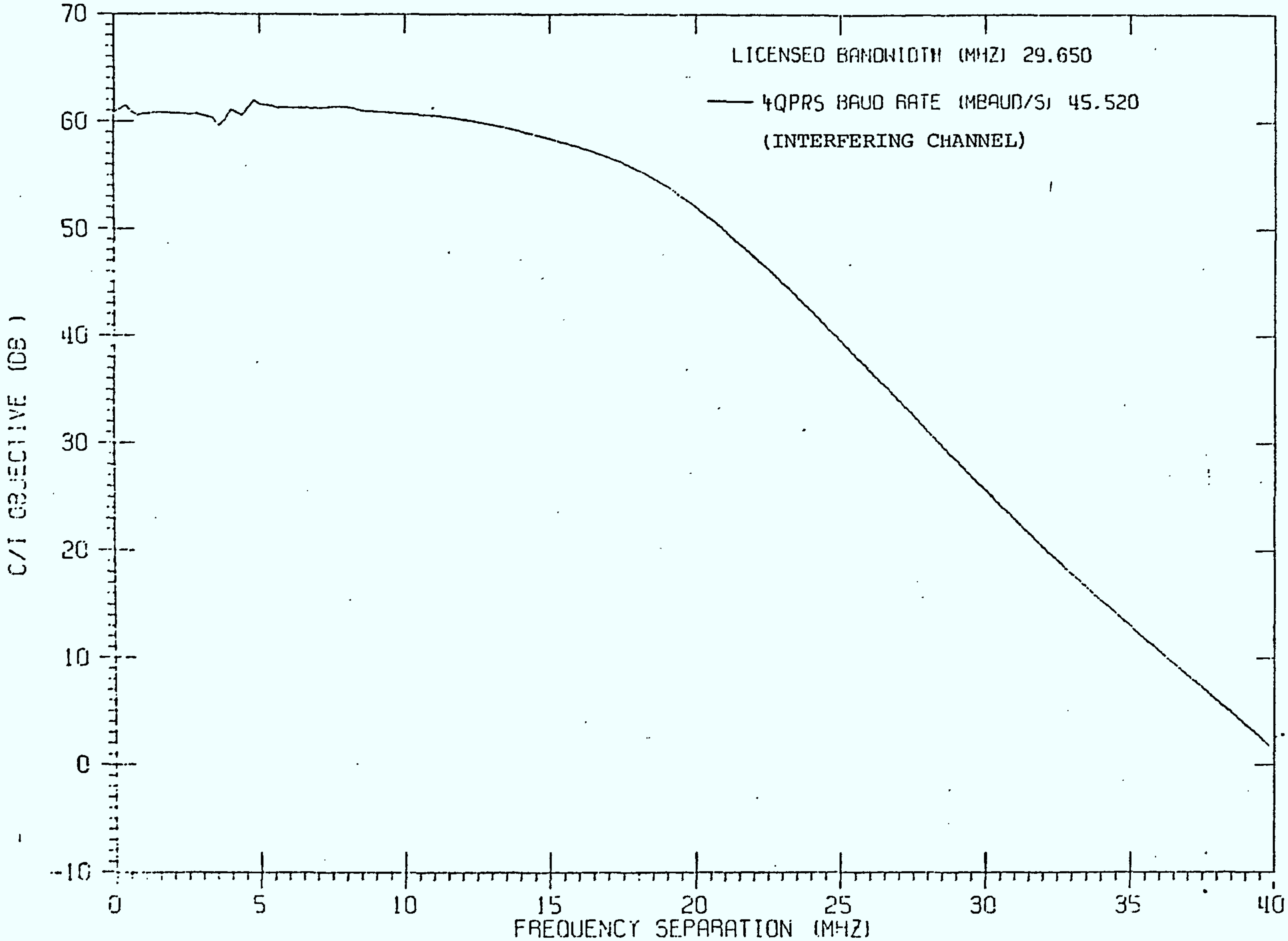


FIGURE J-3 C/I OBJECTIVE VS. FREQUENCY SEPARATION FOR 90DB WEIGHTED VIDEO S/N

APPENDIX K

COMPARISON OF PSK AND FDM/FM

INTO FDM/FM CURVES

APPENDIX K

In this appendix a comparison is made of the relative severity of PSK and FDM/FM interference into FDM/FM channels. The 20 MHz licensed bandwidth case was chosen for this comparison, and only the maximum analogue and digital modulations for this bandwidth were used, namely 1260 channel, 140 kHz test-tone FDM/FM and 18.18 MBs baud rate PSK (filtered for the 20 MHz FCC Mask). The results are presented in Figure K-1, in the form of two labelled curves of C/I objectives versus carrier separation frequency for the 4 dBrnc0 FDM/FM interference objective derived in Appendix F. Again, the signal parameters used are given in Appendix E, Tables 1 and 2.

It is seen from Figure K-1 that, for small carrier frequency separations, FDM/FM interference is much more severe than PSK interference, due to the strong FDM/FM carrier component (but this strong interference is confined to a small number of voice channels in the wanted FDM/FM radio channel). It is conversely seen that, for large carrier frequency separations, PSK interference is much more severe than FDM/FM interference, due to the wide bandwidth of the PSK signals.

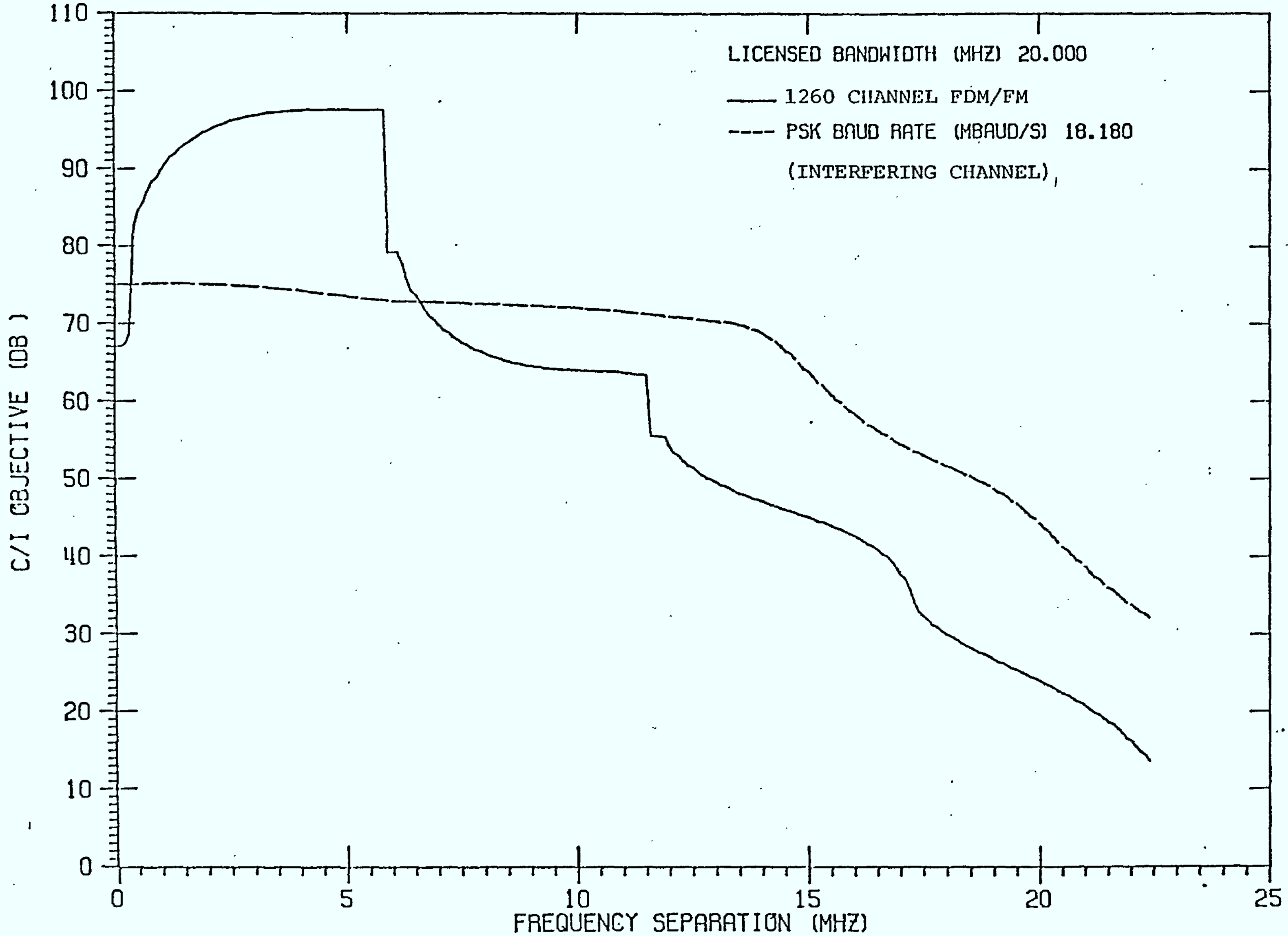


FIGURE K-1 C/I OBJECTIVE VS. FREQUENCY SEPARATION FOR 1260 CHANNEL LOADING FOR 4DBRNC0 NOISE

APPENDIX L

FDM/FM, PCM, FSK AND VIDEO

POWER SPECTRA

APPENDIX L

This appendix provides a comparison of FDM/FM, PSK, FSK and VIDEO RF spectra. The 20 MHz licensed bandwidth case was chosen for the comparison, and only the maximum modulations for this bandwidth were used. Figure L-1 contains labelled curves for all four spectra, in power density per 3.1 kHz spectrum analyzer bandwidth, expressed in dB below total transmitter power. It should be noted that the FDM/FM spectrum is not filtered at all, whereas the PSK and FSK spectra are severely filtered in order to meet the FCC out-of-band emission mask. The video colour bar spectrum reflects normal analogue transmitter output filtering.

It is seen that the PSK, FSK and Video spectra are quite comparable over the 20 MHz bandwidth displayed in Figure L-1, and that the FDM/FM spectrum is comparatively less severe except in the region of the strong FDM/FM carrier component.