application note 175-1

microwave link measurement series

differential phase & gain at work

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FOREWORD

This Application Note is intended to provide a more complete understanding of the swept frequency measurements performed by the hp Microwave Link Analyzer (MLA). It is hoped that the Note will form the basis of more advanced techniques in microwave link optimisation, since the relationships between telephony baseband distortion and swept measurements are clearly defined. However, this is not the end of the story, since some assumptions and approximations have been made, particularly in the case of multihop transmissions. Therefore it is also hoped that further application material will be forthcoming which will improve our understanding of a very complicated subject.

ACKNOWLEDGEMENT

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INTRODUCTION

The rapid growth in wideband traffic carried by microwave radio has drastically reduced the available spectrum space in the lower part of the 'microwave window'. To cope with this growth in traffic, designers have the choice of either using higher transmission frequencies or increasing the amount of traffic carried in the part of the spectrum already in use. Much of the new design work is centred on using higher transmission frequencies. However, the very large investment in equipment employed at present transmission frequencies makes it highly desirable to utilise the full traffic potential of this equipment. It is this realisation of equipment traffic potential that is the subject of this application note.

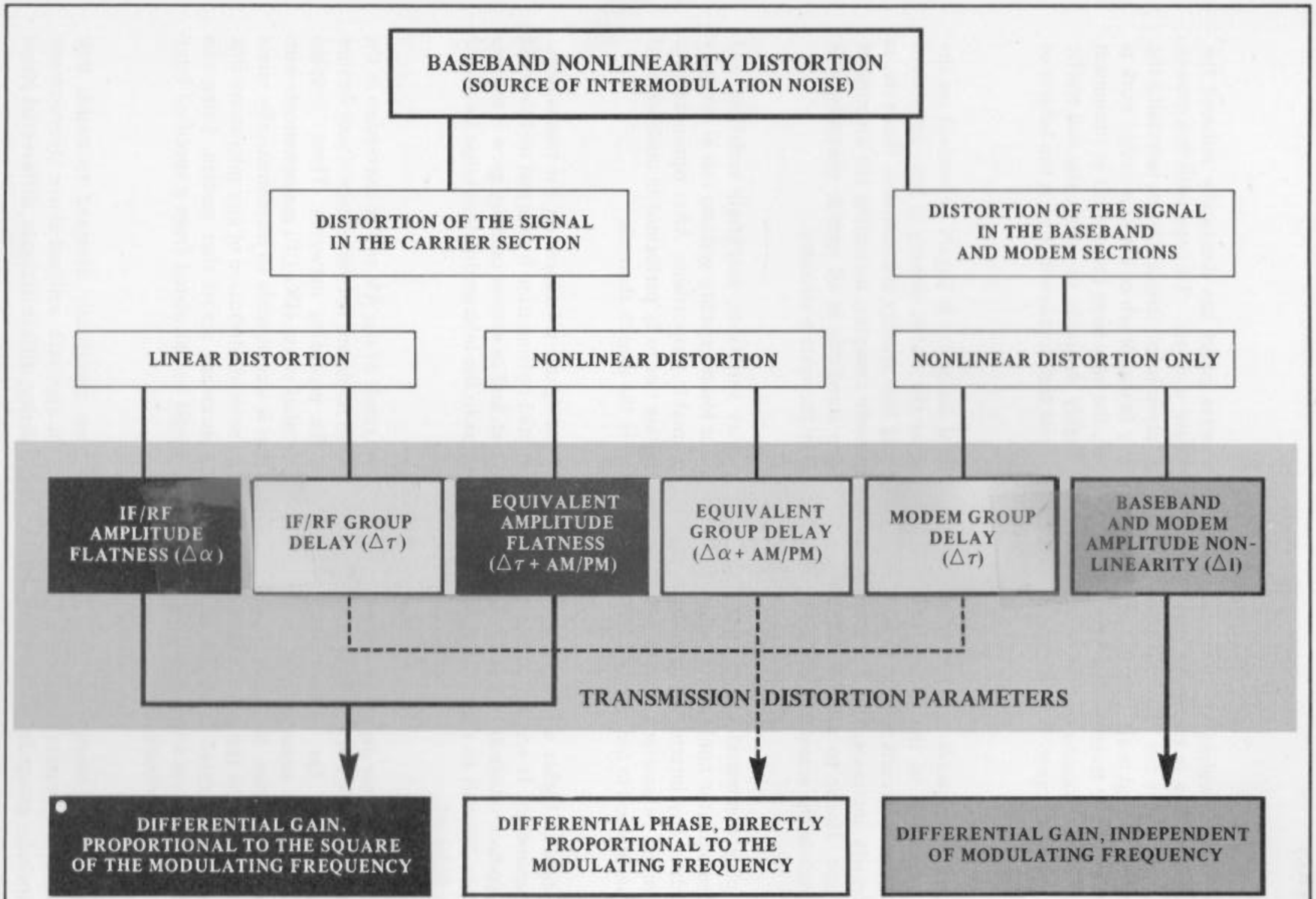
In multi-channel microwave radio relay systems, the signal distortion is largely dependent on the transmission capability of the circuits involved. As long as the traffic density is low, distortions introduced by these circuits will be small enough to permit high quality transmission. However, as the traffic density increases, the FM spectrum becomes more complex, increasing the amount of signal distortion. Hence to achieve the same transmission standards in all speech channels, it is necessary to reduce the amount of distortion contributed by the system circuitry.

Normally, during commissioning of equipment, group delay variations, amplitude variations and modulator/demodulator nonlinearity will be optimised. For lower capacity systems this is normally enough to reduce any intermodulation distortions to acceptable proportions. After optimising the various equipment parameters, 'white noise loading tests' are normally performed in multi-channel telephony transmission in order to ascertain the noise level in the speech channels.

The introduction of higher capacity systems called for even tighter tolerances on the transmission equipment parameters. It was found however, that attempted optimisation by normal methods did not always produce satisfactory results on a white noise test, and as system capacity grew it became more and more difficult to relate the equipment parameters to the intermodulation noise measured by the white noise set.

The main source of these discrepancies were found to be a result of the AM to PM conversion in the transmission path of the FM signal. This AM/PM, occuring in nonlinear networks, introduces further intermodulation from the signal deviations arising in the preceding networks. These 'coupled responses' can only be assessed by differential gain/differential phase (DG/DP) measurements with high frequency test tones. Moreover, it was discovered that it was possible to mathematically relate the DG/DP responses to the noise contribution. Thus the noise contribution of any microwave link section could be calculated from the DG/DP responses measured across that section. Using this technique, the maximum acceptable DG/DP deviations could be calculated from a specified intermodulation noise contribution.

The Hewlett-Packard Microwave Link Analyzer has been specifically designed to enable link manufacturers and operators to optimise their links using the approach outlined above. Measurement capabilities include: group delay distortion, modem nonlinearity, differential gain, differential phase and return loss.



● The dependence on the modulating frequency can be greater than power 2, see conclusions of mathematical analysis in Appendix A1.

Figure 1-1 Sources of Baseband Intermodulation Noise

SECTION I – DISTORTION SOURCES IN FM SYSTEMS

The primary objective in any microwave transmission system is the transfer of wideband information - carrying signals with the minimum amount of distortion. This distortion appears in different forms depending on the type of traffic carried, eg, TV, telephony, digital, etc. In Section IV some effects of distortion on the TV signal are discussed. However, the analysis presented in this Application Note is largely confined to the intermodulation performance of systems carrying telephony traffic. Here, the noise contributed by the system to the telephony signal must be kept within acceptable limits and involves a traffic capacity trade-off, ie, the higher the noise value the lower the traffic capacity. In the case of intermodulation noise, a reduction of the circuit distortions will allow an increase in the channel capacity. With the ever increasing demand for higher density traffic, maximizing the channel capacity by reducing the circuit distortions has become an important consideration for microwave link manufacturers and operators.

Normally, the link performance is evaluated by measuring the noise contribution using a commercial white noise test set. This is a final 'go – no go' acceptance test whereby the transmission quality of the system between baseband (BB) terminals is determined. Although this measurement will separate out the basic noise and intermodulation noise components, it does not localise the noise source.

Figure 1-1 shows the main sources of baseband intermodulation noise in an FM transmission system comprising, BB, modem, carrier IF and carrier RF parts. It can be seen that the contributions to the BB intermodulation noise are from two main sources:

a. the carrier sections – sometimes termed 'Dynamic Distortions'.

b. the BB and modem sections – sometimes termed 'Static Distortions'.

Distortion arising in the carrier sections may be divided into linear and nonlinear distortions. Distortion in the BB and modem sections is nonlinear only. Further subdivision is possible showing the transmission parameters causing these distortions.

It has been recognised for some time that variations in the amplitude and group delay responses result in linear distortions in the carrier section. In FM systems these linear distortions will in turn introduce nonlinear distortion, ie, intermodulation noise in the BB section.

Investigations over the last decade have revealed another source of baseband distortion occurring in the carrier section – the so called 'equivalent amplitude' and 'equivalent group delay' distortions. These distortions are caused by linear/nonlinear cascaded network pairs; the linear network having amplitude and/or delay variations, and the nonlinear network having an AM/PM conversion characteristic.

The BB and modem parts have two distortion sources, ie, the linearity characteristic, and – for the transfer characteristics of the modem only – the group delay characteristic.

Figure 1-1 also shows the types of differential characteristic originating from the various transmission parameters, together with their dependence on the frequency of the baseband test-tone used in the measurement of the differential characteristic.

These relationships are investigated in greater detail in SECTION II.

SECTION II – DIFFERENTIAL GAIN AND PHASE CHARACTERISTICS RELATED TO IF/RF TRANSMISSION PARAMETERS

This section shows how the transmission distortion parameters relate to differential gain and phase in various types of network.

Differential gain and phase can be defined as:

DIFFERENTIAL GAIN — The difference in gain encountered by a low-level, high-frequency sinusoid at two stated instantaneous amplitudes of a superimposed high-level, low-frequency signal.

DIFFERENTIAL PHASE — The difference in phase shift encountered by a low-level, high-frequency sinusoid at two stated instantaneous amplitudes of a superimposed high-level, low-frequency signal.

Thus, the phase and amplitude variations of a baseband test tone are considered at different amplitudes of a low-frequency signal on which it is superimposed: or, when frequency modulated onto a carrier, the phase and amplitude modulation of the test tone. Differential gain and phase are therefore a measure of intermodulation at two specific frequencies.

Single Cascaded Linear/Nonlinear Network

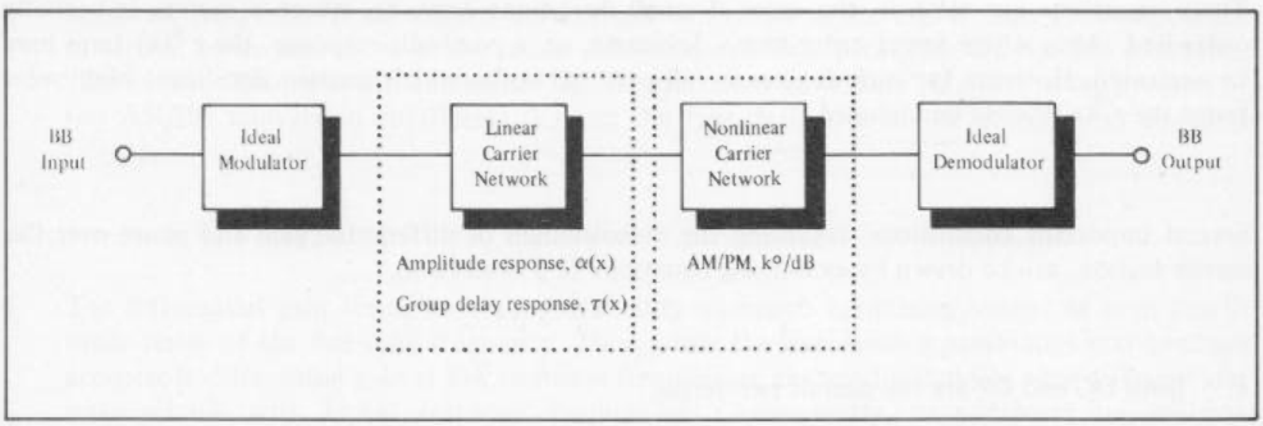


Figure 2-1 Single Cascaded Linear/Nonlinear Network

It is convenient to express the relationship between the BB differential characteristic, and the characteristic originating from the direct and coupled distortions in the carrier part of the transmission system shown in Figure 2-1. This system comprises an ideal modulator and an ideal demodulator, having flat differential gain and phase characteristics between baseband input and output terminals in a direct back-to-back connection at IF. The carrier section, which can be considered as an IF or RF test item, consists of two cascaded two-port networks; the first being a linear network having amplitude and/or group delay variations, the second a nonlinear network having AM/PM conversion. Under these conditions it can be concluded that differential measurements made between BB terminals will be characteristic of the carrier section, and according to the analysis presented in Appendix A1, may be computed from the carrier part responses as follows:

$$DG(x) = 1 + [\alpha''(x) + k\tau'(x)] \frac{\omega_m^2}{2} - \tau'^2(x) \frac{\omega_m^4}{8} \quad \text{(2-1)}$$

$$DP(x) = [\tau(x) - k\alpha'(x)] \omega_m \quad \text{(2-2)}$$

where:

$DG(x)$ is the baseband differential gain characteristic.

$DP(x)$ is the baseband differential phase characteristic.

$\alpha(x)$ is the normalised amplitude response of the linear network.

$\tau(x)$ is the group delay response of the linear network.

x is the swept carrier frequency.

k is the AM/PM factor of the nonlinear network.

ω_m is the test-tone frequency.

$'$ is the first derivative with respect to x.

$''$ is the second derivative with respect to x.

These equations are valid in the case of small deviations only, eg, where a system is partially optimised. Also, where lower order terms dominate, eg, a parabolic response, the $\tau'^2(x)$ term may be neglected. However for such devices as 'all-pass' networks which contain significant high order terms the $\tau'^2(x)$ should be included.

Several important conclusions, regarding the measurement of differential gain and phase over the carrier section, can be drawn by examining equations (2-1) and (2-2).

1. Both DG and DP are the sum of two terms.

   a. The first term originates from the transmission characteristics of the linear network (direct response).

   b. The second term is the combined effect of the linear network characteristics and the nonlinear network AM/PM conversion characteristic (coupled response).

2. In the special case of no AM/PM conversion following the linear network, the DP will be proportional to the group delay variations of the carrier section; the proportionality constant being the test-tone frequency. This is the well known Nyquist principle, used traditionally for group delay measurements.

3. In the general case however, the differential phase characteristic will reveal the combined effect of the linear and nonlinear networks. The nonlinear section will contribute to the DP if there are components of second order and above in the amplitude characteristic preceding the AM/PM converter. This indirect contribution, sometimes called 'equivalent delay', is equal to the product of the AM/PM conversion coefficient (k) and the first derivative of the amplitude response $[\alpha'(x)]$ and will again produce DP which is proportional to the test-tone frequency.

4. In the special case where no AM/PM conversion follows the linear network, the DG will be proportional to the second derivative (α'') of the carrier section amplitude response; the proportionality constant being half the square of the test-tone frequency. As a result, only components of third order (eg, cubic response) and above in the carrier amplitude response will contribute to the DG response between BB terminals.

   Note: This component of DG will occur regardless of subsequent amplitude limiting.

   As already stated the term proportional to the square of the first derivative (τ'^2) of the carrier section group delay response can in some cases be neglected.

5. In the general case, the DG characteristic will reveal the combined effect of the linear and nonlinear networks. The nonlinear section will contribute to the DG if there are components of second order and above in the group delay characteristic preceding the AM/PM converter. This indirect contribution, sometimes called 'equivalent amplitude', is equal to the product of the AM/PM conversion coefficient (k) and the first derivative of the group delay response $[\tau'(x)]$.

6. The differential gain terms have proportionality constants containing second or even fourth order terms of the test-tone frequency. Thus, while the transmission parameters may produce acceptable differential gain at low test-tone frequencies, the results could be very different and unaccpetable with higher test-tone frequencies. Consequently, in wideband transmission systems any alignment using low-frequency test-tone measurements will not necessarily optimise the system for wideband performance.

The conclusions drawn so far are valid between baseband terminals only if an ideal modulator and demodulator are used. It has already been stated in Section I (Figure 1-1), that modem nonlinearity contributes to differential gain. In general this is a direct relationship and so is independent of test-tone frequency. Also modem group delay affects the differential responses in the same manner as group delay in the carrier section. Thus, for accurate differential measurements over the carrier section it is necessary to use a modem which is practically ideal, eg, the hp Microwave Link Analyzer.

The modulator of the MLA Transmitter and the demodulator of the MLA Receiver can be regarded as ideal as far as differential measurements are concerned. The results will not therefore be significantly affected by contributions from the modem of the MLA and equations (2-1) and (2-2) for DG and DP over the carrier section, will relate to measurements made by the MLA.

Multiple Cascaded Linear/Nonlinear Networks

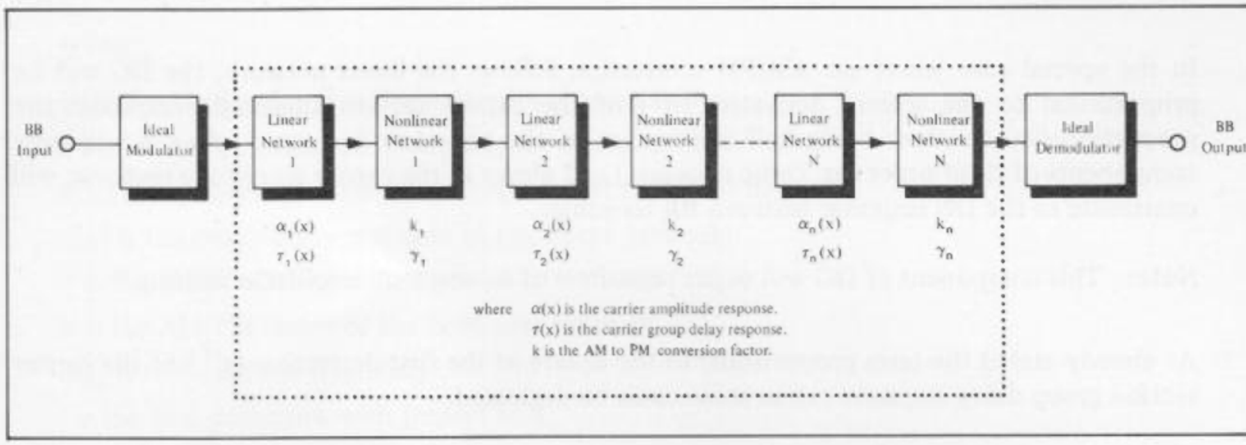


Figure 2-2 Multiple Cascaded Linear/Nonlinear Networks

Consider the same ideal modulator and demodulator as used in Figure 2-1, but with an inserted carrier part consisting of several cascaded linear/nonlinear network pairs. The linear networks have amplitude and group delay variations as before, while the nonlinear network not only has AM to PM conversion but also has a compression coefficient. A system such as that just described is shown in Figure 2-2, and according to the analysis presented in Appendix A1 will yield the following DG and DP equations.

$$DG(x) = 1 + \sum_{r=1}^{n} [\alpha''_r(x) + Keff_r\, \tau'_r(x)] \frac{\omega_m^2}{2} - \tau'^2_r(x) \frac{\omega_m^4}{8} \quad (2\text{-}3)$$

$$DP(x) = \sum_{r=1}^{n} [\tau_r(x) - Keff_r \alpha'_r(x)]\, \omega_m \quad (2\text{-}4)$$

where:

$\alpha_r(x)$ is the normalised amplitude response of the r-th linear network

$\tau_r(x)$ is the group delay response of the r-th linear network

K_r is the AM/PM conversion factor of the r-th nonlinear network

γ_r is the AM compression factor of the r-th nonlinear network

$Keff_r = K_r + \gamma_r K_{r+1} + \gamma_r \gamma_{r+1} K_{r+2} + \ldots\ldots\ldots + \gamma_r \ldots\ldots\ldots \gamma_{n-1} K_n$

is the effective conversion factor of the r-th nonlinear network

As before, several important conclusions can be drawn from equations (2-3) and (2-4).

1. The DG and DP are each represented by a series, in which the number of terms corresponds to the number of cascaded linear/nonlinear networks.

2. Each term of the series has the same form as their counterparts in the single linear/nonlinear network equations (2-1) and (2-2), with the exception of the conversion coefficient. Due to AM compression and limiting in the nonlinear networks an 'effective conversion factor' has to be substituted instead of the actual conversion factor of the network in question. In practice, for cases of more than 30dB compression – typical for limiters in FM radio relay systems – only the AM/PM converter following the last limiter will have an effect.

In practical transmission systems, the individual parameters of the cascaded networks, as shown in Figure 2-2, are not all measurable since individual networks do not necessarily have suitable terminals for the connection of test instruments. However, recent investigations have shown that in the case of microwave FM radio relay systems, a knowledge of individual network parameters is not necessary for the determination of the baseband intermodulation noise, since this noise is uniquely determined by the BB differential responses. Thus, the intermodulation noise originating from any section of a microwave link can be predicted from measurements made using the hp MLA.

SECTION III – BASEBAND INTERMODULATION NOISE RELATED TO RF/IF TRANSMISSION PARAMETERS

Section II mentioned that the transmitter and receiver parts of the hp MLA can respectively be regarded as ideal modulators and demodulators when performing DG and DP measurements, enabling the models shown in Figures 2-1 and 2-2 to be realised in practice. Hence, important conclusions regarding carrier part linear and nonlinear network parameters can be drawn from the DG/DP measurement results, as outlined in Section II. It was also stated that there is a unique relationship between the DG/DP responses and the BB intermodulation noise. This Section will investigate the relationship in greater detail.

The computation of BB intermodulation noise originating from IF/RF transmission parameters is of fundamental importance in the noise assessment of microwave link sections. This is particularly true where the noise contribution of the section is small. It may not be possible to ascertain the intermodulation noise by a direct BB noise measurement (ie, with the section to be measured inserted between IF terminals of a modem, and a white noise test set connected to the BB terminals of this modem), since the white noise test set would then measure the total noise of both the carrier network under test and of the modem. Even the highest quality test modems have intermodulation noise powers of between 20 and 40 pWOp at different slot frequencies in a back-to-back IF connection and could result in the modem noise concealing the test item noise. In general, test modems are not close enough to the ideal for measuring low-noise carrier networks, whereas the transmitter and receiver parts of the MLA can be regarded as a combination of an ideal modulator and an ideal demodulator, even for the measurement of small deviations in responses. Noise computed in this way from the measured responses (using an MLA) could be much more accurate than the noise measured with a modem.

These considerations show that using just one test system – the hp MLA – it is possible to fully categorise the parameters of a microwave link carrier section causing intermodulation noise. Thus, for a required noise performance, the parameters of the carrier section can be given specific values. Conversely, the intermodulation noise can be predicted from MLA measurements of the section parameters.

Noise Computation Methods

Calculation of baseband intermodulation noise from different kinds of transmission characteristics is based on the mathematical representation of stochastic* processes. To simplify these calculations, a small noise contribution from the transmission equipment is generally assumed. Under this condition the 'first order theory' is applicable. According to this theory, the intermodulation noise power originating from the response deviations depends quadratically on these deviations from the ideal flat response, and can be expressed as a sum of four power-additive terms. From these terms, two will depend on the asymmetrical and two on the symmetrical components of the DG and DP characteristics. Asymmetrical components will result in even order intermodulation products and symmetrical components will result in odd order products. However, under operational conditions of FM radio relay systems, only second and third order products will be substantial. The relative level of these products may be ascertained when making white noise measurements, by changing the noise load level by 1dB in the vicinity of the nominal load. The change in intermodulation noise introduced by this load change will depend on the type of characteristic, as shown in Table 3-1.

* See Glossary of Terms

Table 3-1 Differential Gain and Phase Characteristics

RESPONSE COMPONENTS GIVING	DIFF GAIN	DIFF PHASE	DIFF GAIN	DIFF PHASE
POWER ADDITIVE NOISES	ASYMMETRICAL COMPONENT		SYMMETRICAL COMPONENT	
TYPE OF RESPONSE				
ORDER OF INTERMODULATION DISTORTION	SECOND ORDER		THIRD ORDER	
INTERMODULATION NOISE CHANGE PRODUCED BY THE INCREASE OF THE NOISE LOAD* NPR pWOp	 -dB/dB +2dB/dB		 -2dB/dB +3dB/dB	
* IN THE VICINITY OF THE NOMINAL NOISE LOAD LEVEL. AT HIGHER LEVELS (EG, AT THE OVERLOAD LEVEL OF +6dB FREQUENTLY USED) HIGHER ORDER INTERMODULATION PRODUCTS WILL APPEAR AND THE RELATIONSHIPS GIVEN WILL NOT BE VALID.				

The computation methods based on the first order theory approximation may be divided into two groups.

Computer Analysis of Closed-form Expressions

In this group, intermodulation noise is expressed by some closed-form expression which may be applied for any type of characteristic, eg, for responses on equalised link sections which may not be approximated even by high-degree polynomials. For this method use is made of the response values measured at suitable intervals, eg, 2MHz (using the MLA marker comb facility). Calculation of closed-form expressions requires numerical integration via a computer. A suitable computer program for intermodulation noise calculation, will be available as part of the hp Contributed Software Catalogue at some later date.

Polynomial Analysis by Power Series Expansion

Instead of a closed-form expression, the response function is expanded into a power series, and the intermodulation noise is expressed by the coefficents of this series.

If the transmission characteristic can be approximated with reasonable accuracy by a low-order polynomial, then only a few terms of the power series have to be taken into account. This method has the advantage that no computer is required and the intermodulation noise is given by a relatively simple formula.

The main field of application for noise computation from polynomial characteristics is the system check-out of microwave FM equipment. This requires the determination of noise contributions from single equipment parts such as filters, amplifers, mixers, limiters etc. having characteristics which can be approximated by low-order polynomials. The remainder of this chapter shows how these noise contributions can be calculated using derived equations with practical units (%, dB, ns, etc).

Polynomial Equations Using Practical Units

Table 3-2 gives the required equations for intermodulation noise calculations using practical units. These equations show the relationship between intermodulation noise and the various parameters which can be measured using an hp MLA. When used in conjunction with Tables A4-1 to A4-4 or the Nomograms given in Appendix A5, these equations allow determination of:

a. Intermodulation Noise from MLA measurements.

b. Maximum allowable parameter values for a specific intermodulation noise requirement with a given network configuration.

The equations are used as follows:

Equation 3-1 is used for relating intermodulation noise to the DG/DP responses of the carrier part. The carrier part may comprise linear networks; or linear/nonlinear network pairs; or any combinations thereof. The individual network responses need not be known when using this equation.

Equation 3-2 is used for relating intermodulation noise to the amplitude and group delay responses of the carrier part, where the carrier part comprises linear networks only (direct distortion).

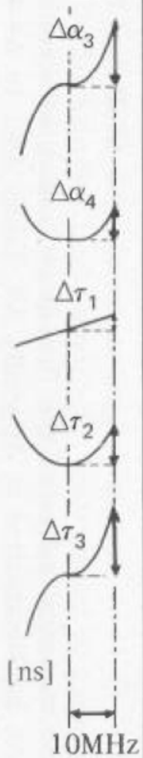
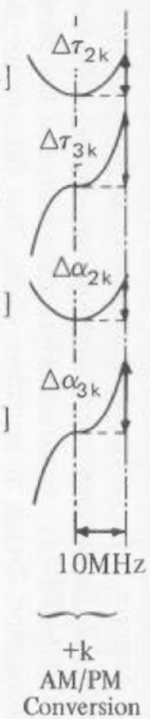
Equation 3-3 is used for relating intermodulation noise to the amplitude and group delay responses of a carrier part comprising a linear/nonlinear network pair; the nonlinear network causing AM/PM conversion (coupled distortion). This equation is used in conjunction with equation (3-2) since the direct distortions still apply.

An example of how these equations can be used is given later in this section.

Equivalent Networks

The equations given in Table 3-2 use practical units corresponding to the MLA trace calibration setting. Further, instead of using response coefficients which are difficult to assess from the MLA display, deviations at ±10MHz are used, since the MLA display will show these directly. The equations will give the noise power (pWOp) at a point of zero relative level, ie, at a point where the test-tone level is 0dBm. This is useful since most white noise test sets read directly in pWOp, and the actual test-tone level at the point where the noise receiver is connected is taken into account by an attenuator setting in the noise receiver.

Table 3-2 Noise Power Equations

BB INTERMODULATION NOISE ORIGINATING FROM:		
BB Differential Characteristics	Direct Transmission Deviations	Coupled Transmission Deviations
$P(\omega)$ [pWOp] =	$P(\omega)$ [pWOp] =	$P(\omega)$ [pWOp] =
$G_1(\omega) DG_1{}^2{}_{(10)}$ [%]	$\alpha_3(\omega)\,\Delta\alpha_3{}^2{}_{(10)}$ [dB]	$\tau_{2k}(\omega)\Delta\tau_{2k}{}^2{}_{(10)}$ [ns] k^2 [°/dB]
$+G_2(\omega) DG_2{}^2{}_{(10)}$ [%]	$+\alpha_4(\omega)\,\Delta\alpha_4{}^2{}_{(10)}$ [dB]	$+\tau_{3k}(\omega)\Delta\tau_{3k}{}^2{}_{(10)}$ [ns] k^2 [°/dB]
$+P_1(\omega)\, DP_1{}^2{}_{(10)}$ [°]	$+\tau_1(\omega)\,\Delta\tau_1{}^2{}_{(10)}$ [ns]	$+\alpha_{2k}(\omega)\Delta\alpha_{2k}{}^2{}_{(10)}$ [dB] k^2 [°/dB]
$+P_2(\omega) DP_2{}^2{}_{(10)}$ [°]	$+\tau_2(\omega)\,\Delta\tau_2{}^2{}_{(10)}$ [ns]	$+\alpha_{3k}(\omega)\Delta\alpha_{3k}{}^2{}_{(10)}$ [dB] k^2 [°/dB]
	$+\tau_3(\omega)\,\Delta\tau_3{}^2{}_{(10)}$ [ns]	
	$+\tau_{13}(\omega)\Delta\tau_{1(10)}$ [ns]$\Delta\tau_{3(10)}$ [ns] (Linear/Cubic Interaction Term)	
		
Equation 3-1	*Equation 3-2*	*Equation 3-3*

where G_1, G_2, P_1, P_2, etc are constants depending on sweep width, capacity, etc and are listed in Appendix A4.

In all cases, noise power is calculated as a sum of several terms, and each term will give the noise contribution of a definite order polynomial response, as shown by the response shapes given for each term in Table 3-2. Each term is a product of two factors. The second factor is the squared response deviation at ±10MHz in equations (3-1) and (3-2), and the product of the squares of the deviation and conversion factor in equation (3-3). The first factor depends on the following parameters, all given in CCIR recommendations:

a. channel number (N).

   multichannel baseband frequency limits.

   baseband noise measuring frequency (slot frequency).

   CCIR Rec. 399-1

b. test-tone deviation.

   CCIR Rec. 404-2

c. multichannel loading nominal
   = test-tone level + (-15 + 10 log N) dB.

   CCIR Rec. 393-1

d. pre-emphasis characteristic.

   CCIR Rec. 275-2

All the data and numerical values of the first factor, calculated according to the above CCIR recommendations, are summarized in Appendix A4, Tables A4-1 to A4-4. The data is given for noise loads corresponding to 960, 1260, 1800 and 2700 channels and for all slot frequencies specified by the CCIR. All numbers are pWOp values for unity response deviation (ie, 1ns, 1dB, 1%) and should be multiplied by the squared response deviation as given by the second factor of the equation terms.

Numbers lower than 0.01 are not contained in the tables since these would in practical systems normally give less than 1pWOp noise power.

In the case of pure characteristics (eg, pure linear or quadratic), appropriate terms of the noise power equations have to be used, eg:

Linear differential gain and quadratic differential phase – use 1st and 4th rows of equation (3-1).

Fourth degree amplitude and cubic group delay – use 2nd and 5th rows of equation (3-2).

For mixed responses, first use Figure 3-2 to determine the 1st, 2nd, 3rd and 4th degree component responses, and then substitute these into the appropriate terms.

With one exception, the terms in the noise power equations will be positive since it is the squared value of the deviations which are substituted. Thus, noise powers arising from different types of differential terms will add directly to give the total noise power. However, the component parts of each differential term add with a voltage relationship.

For example, if two carrier sections are lumped together, then their individual DG_1 terms are added together with respect to voltage. This is true whether the DG_1 terms comprise linear (cubic amplitude) or nonlinear (parabolic delay +AM/PM) parts or both.

The one exception for the noise power term being positive is for a delay response having linear and cubic components. For delay in this form, a special interaction term has to be added to the separate linear and cubic terms, the last term in equation (3-2). This interaction term is a product and may be negative or positive. For instance, with a positive linear and negative cubic (or negative linear and positive cubic) deviation, the interaction term will be negative. Thus, the resulting noise power from the interaction term must be added or subtracted to the other terms accordingly.

Nomograms

As an aid to the rapid determination of intermodulation noise, equations (3-1), (3-2) and (3-3) have been plotted as a series of nomograms. These nomograms will give the pWOp value of the intermodulation noise power as a function of the response deviations at ±10MHz. Log-log co-ordinate systems are used throughout and the quadratic dependence of noise power on the response deviation results in straight line plots. In addition to the logarithmic pWOp scale, a linear dB scale is provided which enables direct 'noise power ratio' (NPR) reading if required. The abscissa scale is calibrated in practical deviation units (ns, dB, %, deg), corresponding to equations (3-1), (3-2) and (3-3), which can be directly read off the MLA display. Each nomogram is accompanied by a simple block diagram and response shapes as a quick reminder of the field of application and of the notation.

The nomograms are intended to evaluate the noise originating from different kinds of distortion for a specific channel number. As an example consider N = 1800 channels which is representative of modern high capacity microwave links. Nomograms shown in Figures A5-12, 13, 14 and 15, entitled "EQUIVALENT DISTORTIONS", are suitable for determining noise originating from the equivalent characteristics as explained previously in this section. In each of these figures, terms shown in one row of equations (3-1), (3-2) and (3-3) are plotted – Figure A5-12 presents the terms in the first row, Figure A5-13 the terms in the second row etc. Rapid comparison of noise contributions originating from equivalent characteristics is thus possible.

The nomograms corresponding to equivalent characteristics have a special feature:

For the same amount of intermodulation noise any response deviation can be assessed from another equivalent response deviation, according to the following:

DG (%) $\longleftrightarrow$ $\Delta\alpha$ [dB] $\longleftrightarrow$ $\Delta\tau_k$ (ns) k(°/dB)

DP (°) $\longleftrightarrow$ $\Delta\tau$ [ns] $\longleftrightarrow$ $\Delta\alpha_k$ (dB) k(°/dB)

Figure A5-1 explains how to extract the equivalent deviations from the "EQUIVALENT DISTORTIONS" nomograms.

In Figure A5-16, all group delay terms are plotted on a single drawing, including the linear/cubic interaction term. As already explained this is the only term which may be either positive or negative. The term is positive if $\Delta\tau_1$ and $\Delta\tau_3$ have the same sign, and negative if they have opposite signs. The ±sign on the interaction plot is a reminder to choose the correct sign.

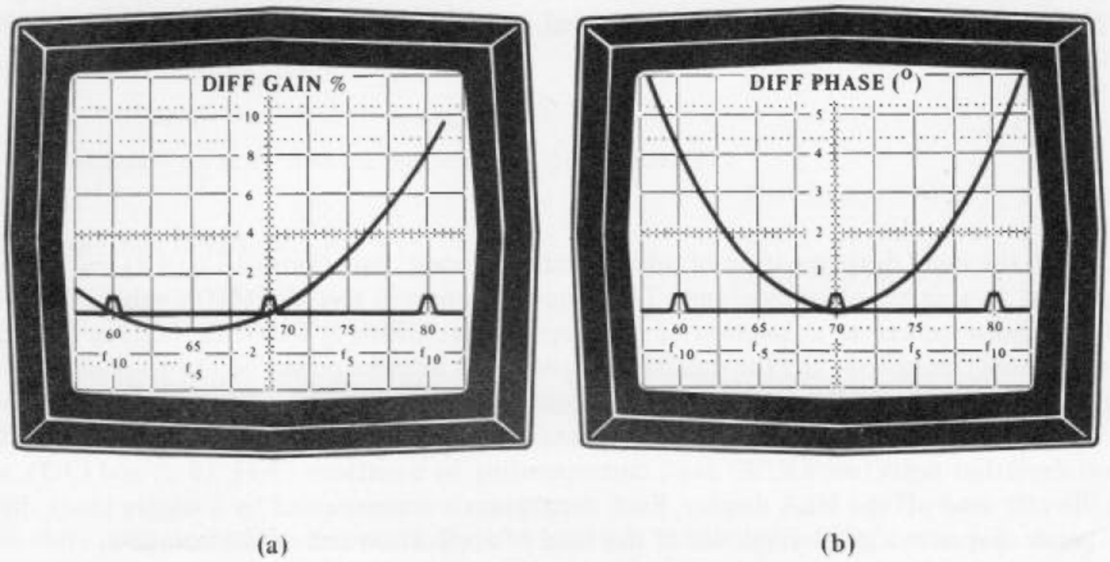
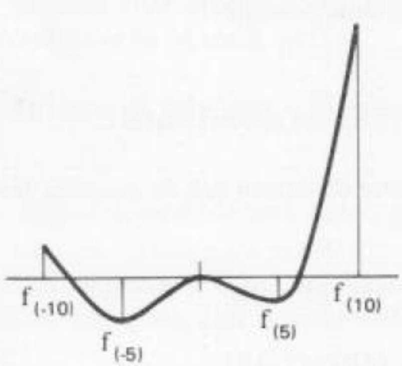


Figure 3-1 Differential Gain and Differential Phase Responses



$$f_{1(10)} = \frac{8(f_5 - f_{-5}) - (f_{10} - f_{-10})}{6}$$

$$f_{2(10)} = \frac{16(f_5 + f_{-5}) - (f_{10} + f_{-10})}{6}$$

$$f_{3(10)} = \frac{2(f_{10} - f_{-10}) - 4(f_5 - f_{-5})}{3}$$

$$f_{4(10)} = \frac{2(f_{10} + f_{-10}) - 8(f_5 + f_{-5})}{3}$$

Figure 3-2 Formulae for Calculating Component Deviations at 10MHz, for Mixed Polynomial Responses

The nomograms given in Appendix A5 can be used to determine the intermodulation noises for the top-slot frequencies. However, noises at any slot frequency for any of the channel numbers 960, 1260, 1800 and 2700 can be easily plotted on a log-log scale co-ordinate system. The plot will be a straight line passing through the pWOp value given in the Tables at unity deviation (eg, 1ns, 1dB, 1%) and having a slope of 2 decades/decade. Also, noises for deviations other than ±10MHz and test-tone frequencies other than 2.4MHz are easily plotted by multiplying the pWOp values given in the Tables by the correction factor given in the last column.

Example 1

Consider the differential gain and differential phase responses shown in Figure 3-1, measured across an 1800 channel carrier section using the MLA. The carrier is swept over ±10MHz and the test-tone frequency is 2.4MHz at 140kHz rms deviation.

The component deviations in equation (3-1), ie, DG_1, DG_2, DP_1 and DP_2 are calculated using the formula given in Figure 3-2 (the centre frequency is used as the datum or zero value).

From Figure 3-1(a):

$$DG1_{(10)} = \frac{8(f_5 - f_{-5}) - (f_{10} - f_{-10})}{6} = \frac{8[3 - (-1)] - [8 - 0]}{6} = 4$$

$$DG2_{(10)} = \frac{16(f_5 + f_{-5}) - (f_{10} + f_{-10})}{6} = \frac{16[3 + (-1)] - [8 + 0]}{6} = 4$$

$$DG3_{(10)} = \frac{2(f_{10} - f_{-10}) - 4(f_5 - f_{-5})}{3} = \frac{2[8 - 0] - 4[3 - (-1)]}{3} = 0$$

$$DG4_{(10)} = \frac{2(f_{10} + f_{-10}) - 8(f_5 + f_{-5})}{3} = \frac{2[8 + 0] - 8[3 + (-1)]}{3} = 0$$

Thus the differential gain response shown in Figure 3-1 is a combined linear/parabolic response.

Similarly from Figure 3-1 (b):

$$DP1_{(10)} = \frac{8(f_5 - f_{-5}) - (f_{10} - f_{-10})}{6} = \frac{8(1 - 1) - (4 - 4)}{6} = 0$$

$$DP2_{(10)} = \frac{16(f_5 + f_{-5}) - (f_{10} + f_{-10})}{6} = \frac{16(1 + 1) - (4 + 4)}{6} = 4$$

$$DP3_{(10)} = 0$$

$$DP4_{(10)} = 0$$

Thus the differential phase response shown in Figure 3-1(b) is purely parabolic.

Referring to Table A4-3 – for a top slot frequency of 7600kHz,

$G_1(\omega) = 115; \quad G_2(\omega) = 3.07; \quad P_2(\omega) = 0.935$

Substituting these values in equation (3-1):

$$P(\omega) = G_1(\omega)\, DG_1{}^2{}_{(10)} + G_2(\omega)\, DG_2{}^2{}_{(10)} + P_2(\omega)\, DP_2{}^2{}_{(10)}$$

$$= 115 \times 4^2 + 3.07 \times 4^2 + 0.935 \times 4^2$$

$$= 1904 \text{ pWOp}$$

Alternatively, the noise values can be read directly from nomograms A5-12, 13 and 15 using the DG_1, DG_2 and DP_2 values.

$$P(\omega) = 1800 + 56 + 15 = 1871 \text{ pWOp (small error due to accuracy of nomograms)}$$

All the nomograms and the constants in Tables A4-1 to A4-4 pertain to response deviations taken at ±10MHz, with a test-tone frequency of 2.4MHz. However, correction factors enabling noise calculations for any swept band and for any test-tone frequency are given in the last column of Tables A4-1 to A4-4.

Thus, using the same differential responses as before only this time assuming a swept bandwidth of ±6MHz:

From Figure 3-1(a),

$$DG_{1\,(6)} = \frac{8(f_3 - f_{-3}) - (f_6 - f_{-6})}{6} = \frac{8[1.55 - (-0.85)] - [3.8 - (-0.95)]}{6} = 2.41$$

$$DG_{2\,(6)} = \frac{16(f_3 + f_{-3}) - (f_6 + f_{-6})}{6} = \frac{16\ (1.55 - 0.85) - (3.8 - 0.95)}{6} = 1.39$$

$$DG_{3\,(6)} = \frac{2(f_6 - f_{-6}) - 4(f_3 - f_{-3})}{3} = \frac{2[3.8 - (-0.95)] - 4[1.55 - (-0.85)]}{3} \simeq 0$$

$$DG_{4\,(6)} = \frac{2(f_6 + f_{-6}) - 8(f_3 + f_{-3})}{3} = \frac{2(3.8 - 0.95) - 8(1.55 - 0.85)}{3} \simeq 0$$

Similarly from Figure 3-1(b),

$$DP1_{(6)} = \frac{8(f_3 - f_{-3}) - (f_6 - f_{-6})}{3} = \frac{(0.35 - 0.35) - (1.45 - 1.45)}{3} = 0$$

$$DP2_{(6)} = \frac{16(f_3 + f_{-3}) - (f_6 + f_{-6})}{6} = \frac{8(0.35 + 0.35) - (1.45 + 1.45)}{6} = 1.36$$

$$DP3_{(6)} = 0$$

$$DP4_{(6)} = 0$$

Hence using the correction factors shown in the last column of Table A4-3

$$P(\omega) = G1\,(\omega)\,DG1^2{}_{(6)} + G2\,(\omega)\,DG2^2{}_{(6)} + P2\,(\omega)\,DP2^2{}_{(6)}$$

$$= 115 \times 2.41^2 \times \frac{1}{\left(\frac{6}{10}\right)^2} + 3.07 \times 1.39^2 \times \frac{1}{\left(\frac{6}{10}\right)^4} + 0.935 \times 1.36^2 \times \frac{1}{\left(\frac{6}{10}\right)^4}$$

$$= 1854.8 + 45.8 + 13.3$$

$$= 1914 \text{ pWOp}$$

Example 2

Use of Nomograms to calculate differential distortion components. This form of calculation can be particularly useful where unknown values of AM/PM occur. Provided the linear/nonlinear section can be identified then the AM/PM can be derived with the aid of the nomograms.

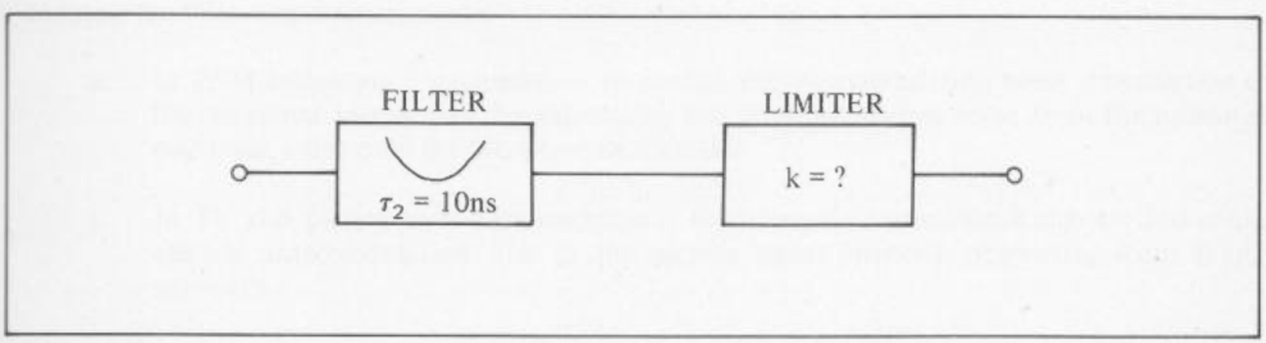


Figure 3-3 Linear/Nonlinear Section

Consider the linear/nonlinear section shown in Figure 3-3.

Any measured differential gain will have a slope component if there is any AM/PM present in the limiter. Assuming a measured differential gain value DG_1 of 1%, then using only DG_1 nomogram (say 1800 channel) the DG_1 noise value is noted. Using this same noise value obtain from the nomogram the equivalent delay + AM/PM value. Since the filter has a known parabolic delay, the AM/PM can be quickly calculated as follows:

Using the 1800 channel nomogram and the top slot,

DG_1 = 1%, gives a noise value of 115 pWOp.

From the same nomogram 115 pWOp gives a τ_2 (ns) k (°/dB) value of 18.

Now, since $\tau_2 = 10\text{ns}$ then $k = \frac{18}{10} = 1.8°/\text{dB}$

Note: In this case the α_3 (cubic amplitude) component is zero and is not considered. However, if some α_3 component was present, then the equivalent DG_1 value for α_3 has to be subtracted from the total DG_1 value.

eg, if $\alpha_3 = 0.2\text{dB}$; then using the 1800 channel nomogram, α_3 gives 19 pWOp which corresponds to a DG_1 value of 0.4%.

∴ the DG_1 due to parabolic delay + AM/PM is 1% −0.4% ie, 0.6%.

A DG_1 of 0.6% gives a noise value of 40 pWOp and corresponds to a τ_2 (ns) k (°/dB) value of 11.

ie, $k = \frac{11}{10} = 1.1°/\text{dB}$

SECTION IV – HP MLA MEASUREMENT CAPABILITIES FOR LINK OPTIMISATION

Sections I to III stated the distortions ecnountered in wideband microwave transmission equipment and described how they relate to each other.

The hp Microwave Link Analyzer (MLA) is intended to categorise these distortion sources. Measurements are made on a calibrated swept frequency basis, from which the results of classical link tests can be deduced. In particular, differential gain and differential phase measurements can be made, whereby the noise contribution for telephony loading or the waveform distortion for TV loading can be predicted.

The MLA comprises a transmitter and a receiver unit which interface with the radio equipment at BB or IF. Interface can also be made at an RF output terminal using the Down Converter 3730A, or at an RF input terminal using the Communication Sweep Oscillator 8605A. This BB, IF and RF capability allows the component sections of the link equipment to be analyzed wherever interface connectors are provided. Since the frequency modulated RF signal is of the same form as the IF signal, the distortion effects are also the same. Hence a separate analysis for RF distortion is not normally necessary.

The measurement technique of sweeping a test tone over the normal carrier band places important criteria on the choice of test-tone frequency. In general, the choice divides into two groups:

a. Low frequency (<1MHz) – for linearity measurements of baseband and modem circuits, and for group delay measurements of modem and IF.

b. High frequency (>2MHz) – for differential gain and phase measurements of IF circuits.

Low-frequency tones of 83.3, 250 and 500kHz are supplied as standard to serve different requirements in group delay and linearity measurements. In general, the 500kHz test tone would be used since this gives the best sensitivity over noise. However, to obtain the necessary resolution when measuring rapid group delay variations, such as ripple responses, the 250 or 83.3kHz tones should be used. The trade-off for increased resolution is increased noise on the display.

The high-frequency tones offered as standard are 2.4, 4.43, 5.6 and 8.2MHz. These tones are used to make the following measurements:

a. In FDM telephony transmission – to predict the intermodulation noise contribution of the measured carrier part by calculating the intermodulation noise from the measured responses, using only the MLA (see Section III).

b. In TV and picture sound transmission – to determine chrominance channel and sound channel intermodulation due to the picture signal crosstalk originating from IF/RF networks.

Other high-frequency tones are available as options eg, 3.58MHz is offered for the TVSC chrominance standard where NTSC is used.

MLA Measurements over FDM Telephony Links

Prediction of Intermodulation Noise from the IF/RF Responses

For predicting the intermodulation noise originating from carrier networks the 2.4MHz test-tone is recommended for measuring the DG/DP responses. The 2.4MHz is a high enough frequency to yield a low-noise display (especially in the case of DG measurements where the response amplitude is proportional to the test-tone frequency squared) while still being low enough to avoid averaging due to excessive spectrum spreading.

As discussed in Section III, two methods may be used to calculate intermodulation noise depending on the response type measured. For responses which can be reasonably approximated by low-order polynomials, equations (3-1), (3-2) and (3-3) can be used with the aid of Tables A4-1 to A4-4, or alternatively, with the nomograms given in Figures A5-2 to A5-21. In these equations and nomograms, the numerical constants are chosen to correspond to a 2.4MHz test-tone frequency and responses measured at ±10MHz from the carrier centre frequency. In Tables A4-1 to A4-4 correction factors are given which allow the equations and nomograms to be used for other test-tone frequencies and carrier bandwidths. Where the measured responses are more complicated and cannot be approximated by low-order polynomials the use of a computer program is recommended. A program will be available as part of the hp Contributed Software Catalogue at a later date.

MLA Measurements over Television Links

When testing systems transmitting TV colour picture and possibly four sound channels, the MLA high-frequency test tones are used to substitute the subcarrier signals which are present under operational conditions, and the sweep width is chosen to equal the peak deviation produced by the low-frequency picture signal components (approximately 2.2MHz with the pre-emphasis characteristic given in CCIR Rec. 405-1). Note that with such a low sweep width, the MLA REDUCE BB FREQ lamp will be lit, since the sidebands produced by the high-frequency test tone will not sweep the total IF band. However, the warning can be ignored in this case, since this is the actual operating condition to be simulated.

MLA Measurements for TV Picture-to-Chrominance Channel Crosstalk

The unwanted intermodulation of the 4.43MHz colour subcarrier caused by the picture signal low-frequency components can be evaluated using the MLA 4.43MHz test tone. This test tone is specifically chosen to equal the European standard colour subcarrier frequency. Consequently, DG and DP measurements performed using the sweep width as explained above, will bear a direct relationship with the standard television DG and DP measurements as specified by the CCIR recommendations. A similar argument applies for the North American (NTSC) subcarrier frequency of 3.58MHz (available as an option on the MLA).

MLA Measurements for TV Picture-to-Sound Channel Crosstalk

The television signal may carry as many as four sound subcarriers as well as the picture information. These subcarriers are usually transmitted in the range 7 to 9MHz and are frequency modulated by the sound information. Therefore DP measurements made using the MLA with an 8.2MHz test tone will indicate the unwanted intermodulation of one of the sound subcarriers, ie, picture-to-sound crosstalk.

The crosstalk noise level may be derived by relating the measured DP value to the phase deviation produced by a nominal 800Hz sound channel test tone. Measurements of DG are normally unnecessary for assessing sound channel performance since limiting before subcarrier demodulation will suppress any DG variations, unless AM/PM conversion occurs in the subcarrier limiter.

Under operational conditions, crosstalk is dependent on the type of picture signal. Normally frame sync pulses are responsible for most of the crosstalk experienced. Crosstalk values measured with a 70 or 50Hz sweep signal will therefore be characteristic of operational conditions since these frequencies are close to the picture repetition rate.

Table 4-1 is a summary of factors governing the choice of test-tone frequency.

Sweeping the Correct Carrier Bandwidth

With the introduction of high-frequency test tones, the traditional assumption that the test-tone deviation is insignificant compared to the sweep deviation is no longer justified. Consider, for example, a carrier swept ±15MHz while modulated with a 5.6MHz test tone. Then the total bandwidth explored, taking into account the test tone sidebands, will be ±15 ±5.6MHz=±20.6MHz instead of the desired ±15MHz. Thus, it would be possible to arrive at misleading results when switching from low to high-frequency test tones, due to the sidebands of the latter falling into the stop band of bandpass filters. There is also a danger of sweeping into the passband of neighbouring RF channels, making maintenance response tests in a multi-RF channel link hazardous.

This problem is overcome on the MLA by the 'Auto Sweep Reduction' facility. For test-tone frequencies above 500kHz, this facility automatically reduces the IF sweep width in proportion to the chosen test-tone frequency. Thus, the IF SWEEP WIDTH control setting will not show the IF carrier deviation, but the deviation produced by the sweep signal *plus* twice the test-tone frequency. The SWEEP WIDTH setting shows the actual bandwidth explored by the IF signal whether low or high-frequency test tones are used.

However, a certain amount of caution is required when using the sweep reduction facility since a change in the test-tone frequency will change two parameters simultaneously, ie, the test-tone frequency and the sweep width. The effects of test-tone frequency changes will not therefore show up clearly.

If the need arises to observe the dependence of the DG/DP responses on the test-tone frequency then the following method may be useful:

Using a low test-tone frequency, set the SWEEP WIDTH to cover the desired sweep range *plus twice* the highest test-tone frequency to be selected later. Position the sliding markers to cover the sweep range to be investigated. Observations should only be made between the sliding markers, and response parts falling outside these markers should be ignored. Switching over now to higher test-tone frequencies, the markers will be displaced outwards, indicating the action of the automatic sweep reduction.

If the basic shape of the response within the sliding markers changes when switching over to a higher test-tone frequency, then averaging due to spectrum spreading has taken place, and should be considered as a warning that the higher frequency test tone will not allow the display of the true response.

Table 4-1 Choice of Test-tone Frequency

MEASURED PARAMETER	MEASURED SECTION	RECOMMENDED TEST TONE	REASON FOR RECOMMENDATION
GROUP DELAY RESPONSE	Sections of Microwave Links interfacing BB, IF or RF.	83kHz 250kHz 500kHz	Group delay resolution decreases with increasing test-tone frequency. 500kHz is the maximum desirable, and even then only for slow variations in slope. Higher frequencies have better noise performance.
LINEARITY RESPONSE [DIFFERENTIAL GAIN DUE TO MODEM]	Modulators, Demodulators, Baseband Sections.	500kHz or lower	The linearity of these sections is independent of test-tone frequency. Frequencies higher than 500kHz may detect IF/RF nonlinearities, producing error.
DIFFERENTIAL GAIN & DIFFERENTIAL PHASE (For use in noise prediction etc, using Nomograms)	IF/RF Sections	2.4MHz	This test frequency is high enough to reveal deviations in practical circuits, yet low enough to avoid spectrum averaging. Also it is a frequency that can be used for system capacities down to 600 channels.
DIFFERENTIAL GAIN & DIFFERENTIAL PHASE (For prediction of Chrominance to Luminance Crosstalk)	Sections of Microwave Links interfacing at BB, IF or RF.	4.43MHz (European standard) 3.58MHz (North American standard)	This is almost equal to the TV colour sub-carrier frequency and so results will relate directly to standard television measurements.
DIFFERENTIAL PHASE (For prediction of picture to sound crosstalk)		8.2MHz (Reduce BB frequency lamp will light but should be ignored in this case).	This is the region where sound sub-carriers are transmitted. Since the sound is frequency modulated, differential phase measurements will give an indication of distortion.

APPENDIX A1 – DIFFERENTIAL GAIN AND DIFFERENTIAL PHASE DISTORTION IN FM SYSTEMS

DIFFERENTIAL GAIN AND DIFFERENTIAL PHASE DISTORTION IN FM SYSTEMS

Definition of Differential Gain and Differential Phase

Differential gain and differential phase are parameters defined for a transmission system with base-band input and output ports. The input signal is composed of a sinusoidal test tone with frequency ω_m and amplitude V_m superimposed on a slowly varying sweep signal V_s. Corresponding to the theoretical condition of the differential measurement, the limiting case of $V_m \longrightarrow 0$ is considered:

$$v_{in} = V_s + V_m \cos \omega_m t \quad \ldots \ldots \quad (1)$$

At the output port of the system the test-tone component v_{out} is measured. Let A_o and ϕ_o be the gain and the phase measured with the sweep signal V_s set to zero amplitude.

$$v_{out} = A_o V_m \cos [\omega_m t - \phi_o] \quad \ldots \ldots \quad (2)$$

Varying the sweep signal amplitude, the gain and the phase are found to be dependent on V_s due to the system nonlinearities; therefore at some amplitude $V_s = x$,

$$v_{out} = A(x) V_m \cos [\omega_m t - \phi(x)] \quad \ldots \ldots \quad (3)$$

Using the characteristics $A(x)$ and $\phi(x)$, the differential gain and the differential phase are defined by the following expressions:

$$DG(x) = \frac{A(x)}{A_o} \quad \ldots \ldots \quad (4)$$

$$DP(x) = \phi(x) - \phi_o \quad \ldots \ldots \quad (5)$$

In some calculations, notations might be simplified by introducing the complex differential response as follows:

$$D(x) = \frac{A(x)}{A_o} e^{-j[\phi(x) - \phi_o]} \quad \ldots \ldots \quad (6)$$

Differential gain and differential phase are related to the complex differential response by the expressions:

$$DG(x) = |D(x)| \qquad DP(x) = \text{Arc } D(x) \quad \ldots \ldots \quad (7)$$

For a system with no distortion, the differential gain is unity and the differential phase is zero. Any difference from these values will show up as a source of intermodulation distortion. In practice differential gain is normally expressed as a percentage.

In this definition (x) is some value of the sweep signal amplitude. However, it could also be defined as the deviation from the centre frequency of the carrier by the sweep signal.

Complex Representation of Modulated Signals

To calculate the differential gain and differential phase contributions of an FM system, a carrier phase modulated by the test-tone frequency is assumed at the input of the system.

$$v_{ci} = V_{ci}\, e^{j[\omega_c t + \Delta p \cos \omega_m t]} \quad \ldots \quad (8)$$

The method for measuring differential gain and phase is to sweep the carrier (ω_c) around the band centre frequency (ω_o). As explained earlier, (x) can be defined as the deviation from the centre frequency of the carrier by the sweep signal.

$$\therefore \quad \omega_c = \omega_o + x \quad \ldots \quad (9)$$

Δp is the phase deviation and has a differential value; therefore a first order approximation for the exponential $e^{j\Delta p \cos \omega_m t}$ is valid.

$$v_{ci} = V_{ci}\, e^{j\omega_c t}\, [1 + j\Delta p \cos \omega_m t] \quad \ldots \quad (10)$$

Expressing the cosine function with exponentials:

$$v_{ci} = V_{ci} \left[e^{j\omega_c t} + \frac{j}{2}\, \Delta p e^{j(\omega_c + \omega_m)t} + \frac{j}{2}\, \Delta p e^{j(\omega_c - \omega_m)t} \right] \quad \ldots \quad (11)$$

In equation (11), the phase modulated signal is expressed by the carrier and two sideband components. The transmission path will change the relative amplitudes and phases of the sideband components, resulting in *phase modulation distortion* and also in *phase-to-amplitude modulation conversion*. In order to evaluate these effects a carrier with both phase and amplitude modulations has to be considered.

Denote by p and m the phase and amplitude modulation indexes, and by ϑ_p and ϑ_m the modulation phases. With these notations a carrier having both PM and AM modulations can be written as:

$$v_c = V_c\, [1 + m \cos(\omega_m t - \vartheta_m)]\, e^{j[\omega_c t + p \cos(\omega_m t - \vartheta_p)]} \quad \ldots \quad (12)$$

With the approximation valid for $m \times p \ll 1$,

$$v_c = V_c\, [1 + m \cos(\omega_m t - \vartheta_m) + j p \cos(\omega_m t - \vartheta_p)]\, e^{j\omega_c t} \quad \ldots \quad (13)$$

To express the modulation with two sideband components, we introduce the complex modulation amplitudes for AM and PM as follows:

$$M = m e^{-j\vartheta_m} \qquad P = p e^{-j\vartheta_p} \quad \ldots \quad (14)$$

Using the complex amplitudes M and P the modulated carrier can be expressed as follows:

$$v_c = V_c \left[e^{j\omega_c t} + \frac{M + jP}{2}\, e^{j(\omega_c + \omega_m)t} + \frac{M^* + jP^*}{2}\, e^{j(\omega_c - \omega_m)t} \right] \quad \ldots \quad (15)$$

Where * denotes the complex conjugate

Equation (15), gives a unique representation of a carrier modulated by a single test-tone frequency. Denote by U_u and U_l, the relative complex amplitudes of the upper and lower sideband components respectively, normalised to the carrier amplitude V_c.

$$v_c = V_c \left[e^{j\omega_c t} + U_u e^{j(\omega_c + \omega_m)t} + U_l e^{j(\omega_c - \omega_m)t} \right] \qquad (16)$$

Given U_u and U_l, the complex modulation amplitudes P and M can be determined by comparing equations (15) and (16).

$$P = -j[U_u - U_l^*] \qquad M = U_u + U_l^* \qquad (17)$$

Equations (16) and (17) give the basic relationships for calculating the differential phase and gain contributions of an FM system.

Single Linear Network in the FM Path

First consider the case of a single linear network in the FM path, with gain characteristic $A(\omega)$ and phase characteristic $\psi(\omega)$. An input signal with frequency ω and phasor v_1 will produce an output v_2 given by:

$$v_2 = A(\omega)e^{-j\psi(\omega)} v_1 \qquad (18)$$

In further calculations the relative amplitude and phase characteristics $\alpha(\omega)$ and $\varphi(\omega)$ normalised to the complex gain at the nominal carrier frequency ω_o will be used as follows:

$$e^{\alpha(\omega) - j\varphi(\omega)} = \frac{A(\omega)e^{-j\psi(\omega)}}{A(\omega_o)e^{-j\psi(\omega_o)}} \qquad (19)$$

The differential characteristics are measured with the input signal given by equation (11). The output spectrum is calculated by multiplying each component by the corresponding complex gain factor. Using the notation introduced above:

$$v_2 = A(\omega_c)e^{-j\psi(\omega_c)} V_{ci} \left[e^{j\omega_c t} + j\frac{\Delta p}{2} \frac{e^{\alpha(\omega_c + \omega_m) - j\varphi(\omega_c + \omega_m)}}{e^{\alpha(\omega_c) - j\varphi(\omega_c)}} e^{j(\omega_c + \omega_m)t} \right.$$

$$\left. + j\frac{\Delta p}{2} \frac{e^{\alpha(\omega_c + \omega_m) - j\varphi(\omega_c + \omega_m)}}{e^{\alpha(\omega_c) - j\varphi(\omega_c)}} e^{j(\omega_c - \omega_m)t} \right] \qquad (20)$$

Where * denotes the complex conjugate

In the general case the output has both phase and amplitude modulations. The complex modulation amplitudes P and M can be determined using equation (17).

$$P_2 = \left[\frac{e^{\alpha(\omega_c + \omega_m) - j\varphi(\omega_c + \omega_m)}}{2e^{\alpha(\omega_c) - j\varphi(\omega_c)}} + \frac{e^{\alpha(\omega_c - \omega_m) + j\varphi(\omega_c - \omega_m)}}{2e^{\alpha(\omega_c) + j\varphi(\omega_c)}} \right] \Delta p \quad \cdots \quad (21)$$

$$M_2 = \left[\frac{je^{\alpha(\omega_c - \omega_m) - j\varphi(\omega_c + \omega_m)}}{2e^{\alpha(\omega_c) - j\varphi(\omega_c)}} - \frac{je^{\alpha(\omega_c - \omega_m) + j\varphi(\omega_c - \omega_m)}}{2e^{\alpha(\omega_c) + j\varphi(\omega_c)}} \right] \Delta p \quad \cdots \quad (22)$$

For evaluating the differential response contribution, an ideal PM demodulator is assumed at the output of the linear network, suppressing the amplitude modulation and reproducing the phase modulation component. The complex gain of the demodulated test-tone frequency component is given by the bracketed term in equation (21). Substitution of the sweep variable (x) from equation 9 into this term will yield the complex differential response as follows:

$$D(x) = \frac{e^{\alpha(\omega_o + x + \omega_m) - j\varphi(\omega_o + x + \omega_m)}}{2e^{\alpha(\omega_o + x) - j\varphi(\omega_o + x)}} + \frac{e^{\alpha(\omega_o + x - \omega_m) - j\varphi(\omega_o + x - \omega_m)}}{2e^{\alpha(\omega_o + x) + j\varphi(\omega_o + x)}} \quad \cdots \quad (23)$$

From equation (23), the differential gain and differential phase characteristics can be calculated by using the relationship given in equation (7).

Linear Network Followed by an AM/PM Converter

In FM systems, modulation distortion is originated by two different mechanisms. The kind of distortion produced by linear networks has been discussed in the preceding paragraphs. There is however, a second type of distortion resulting from the coupled effects of linear and nonlinear circuits in the path of the phase modulated signal. This second type of distortion can be considered as a series of cascaded linear and nonlinear circuits, as shown in Figure 2-2. In the cascade, frequency dependence is assumed for the linear part only, whereas the nonlinear circuit is taken to be independent of frequency within the sidebands of the modulated carrier. In practice this assumption is usually justified.

A carrier signal with pure phase modulation will be transmitted by the nonlinear network without distortion. As the linear circuit is assumed to be independent of frequency, the phase modulation index will be unchanged and there will be no conversion to amplitude modulation. However, when the nonlinear network is exposed to amplitude modulation, two forms of distortion can be produced.

1. The amplitude modulation index may be changed resulting in AM compression (denoted by γ).

2. AM to PM conversion may occur whereby the AM may be partly converted into PM (denoted by k).

In practical circuits, both γ and k appear to be independent of the modulating frequency, which corresponds with the previous assumption. Using the parameters γ and k, the modulation transmission through the nonlinear circuit can be described in terms of the modulation amplitudes P and M. Denoting by P_2 and M_2 the input and by P_3 and M_3 the output modulation amplitudes, then:

$$P_3 = P_2 + k M_2 \quad \ldots \quad (24)$$

$$M_3 = \gamma M_2 \quad \ldots \quad (25)$$

The overall characteristics of the cascaded linear and nonlinear circuits in Figure 2-1 can be determined by substituting equations (21) and (22) into equations (24) and (25). Then for the complex amplitudes P_3 and M_3:

$$P_3 = \left[\frac{1+jk}{2} \frac{e^{\alpha(\omega_c+\omega_m)-j\varphi(\omega_c+\omega_m)}}{e^{\alpha(\omega_c)-j\varphi(\omega_c)}} + \frac{1-jk}{2} \frac{e^{\alpha(\omega_c-\omega_m)+j\varphi(\omega_c-\omega_m)}}{e^{\alpha(\omega_c)+j\varphi(\omega_c)}} \right] \Delta p \quad \ldots \quad (26)$$

$$M_3 = \left[\frac{j\gamma}{2} \frac{e^{\alpha(\omega_c+\omega_m)-j\varphi(\omega_c+\omega_m)}}{e^{\alpha(\omega_c)-j\varphi(\omega_c)}} - \frac{j\gamma}{2} \frac{e^{\alpha(\omega_c-\omega_m)+j\varphi(\omega_c-\omega_m)}}{e^{\alpha(\omega_c)+j\varphi(\omega_c)}} \right] \Delta p \quad \ldots \quad (27)$$

The complex differential response is given by the coefficient of Δp in equation (26):

$$D(x) = \frac{1+jk}{2} \frac{e^{\alpha(\omega_o+x+\omega_m)-j\varphi(\omega_o+x+\omega_m)}}{e^{\alpha(\omega_o+x)-j\varphi(\omega_o+x)}} + \frac{1-jk}{2} \frac{e^{\alpha(\omega_o+x-\omega_m)-j\varphi(\omega_o+x-\omega_m)}}{e^{\alpha(\omega_o+x)+j\varphi(\omega_o+x)}} \quad \ldots \quad (28)$$

The amplitude modulation given by M_3 has no significance if the output is connected directly to the demodulator which is assumed to be insensitive to AM. If the cascade is representing only a part of the FM system and is followed by similar cascades of linear and nonlinear networks, then the amplitude modulation component has to be considered. This general case will be discussed in the last section.

Practical Expressions for the DG and DP Characteristics

The expression derived for the complex differential response as given in equation (28) can be simplified considerably by using some straight forward approximations. The differential characteristic is given by terms with differences from the normalized amplitude and phase responses in the exponentials. These differences can be approximately expressed using the first and second derivatives of the respective characteristics.

$$\alpha(\omega_o+x\pm\omega_m)-\alpha(\omega_o+x) \simeq \pm\alpha'(\omega_o+x)\,\omega_m+\alpha''(\omega_o+x)\,\frac{\omega_m^2}{2} \quad \ldots \quad (29)$$

$$\varphi(\omega_o+x\pm\omega_m)-\varphi(\omega_o+x) \simeq \pm\varphi'(\omega_o+x)\,\omega_m+\varphi''(\omega_o+x)\,\frac{\omega_m^2}{2} \quad \ldots \quad (30)$$

The approximations are valid when the test-tone frequency is not too high with respect to the curvature of the amplitude and phase responses. Using equations (29) and (30) the expression for $D(x)$ becomes:

$$D(x) = \frac{1+jk}{2} \exp\left\{[\alpha'(x) - j\tau'(x)]\,\omega_m + [\alpha''(x) - j\tau'(x)]\,\frac{\omega_m^2}{2}\right\} +$$

$$\frac{1-jk}{2} \exp\left\{[-\alpha'(x) - j\tau(x)]\,\omega_m + [\alpha''(x) + j\tau'(x)]\,\frac{\omega_m^2}{2}\right\} \quad \ldots \quad (31)$$

where simply x stands for the independent variable and $\tau(x)$ denotes the derivative of the phase, ie, the group delay characteristic.

For practical systems with small transmission deviations in the passband, the arguments of the exponential terms are much smaller than unity, and the relationship $\exp Z = 1 + Z + \frac{Z^2}{2}$ can be applied:

$$D(x) = 1 - j[\tau(x) - k\alpha'(x)]\,\omega_m + [\alpha''(x) + k\tau'(x)]\,\frac{\omega_m^2}{2} \quad \ldots \quad (32)$$

The differential gain and phase are calculated from equation (7). Since the contributions from the second and third terms in equation (32) are much smaller than unity, the absolute value and the phase of $D(x)$ will be given by the real and imaginary parts respectively:

$$DG(x) = 1 + [\alpha''(x) + k\tau'(x)]\,\frac{\omega_m^2}{2} \quad \ldots \quad (33)$$

$$DP(x) = [\tau(x) - k\alpha'(x)]\,\omega_m \quad \ldots \quad (34)$$

In the $DG(x)$ expression the term $\tau'^2(x)\,\frac{\omega_m^4}{8}$ should be added (see equation 36), as this is often significant. The $DG(x)$ term then becomes:

$$DG(x) = 1 + [\alpha''(x) + k\tau'(x)]\,\frac{\omega_m^2}{2} - \tau'^2(x)\,\frac{\omega_m^4}{8} \quad \ldots \quad (35)$$

Equations (34) and (35) are the practical DG and DP expressions for a linear network followed by an AM/PM converter.

In some cases the second order approximation in the Taylor expansions will not be valid. If terms up to fourth order in the Taylor expansions are taken, then the following equations for DG and DP are obtained. These additional terms were not considered when deriving the nomograms and tables for the noise due to the carrier amplitude and group delay terms. They do not affect the nomograms and tables for the noise due to the differential terms.

$$DG(x) = 1 + (\alpha'' + \alpha'^2 + k\tau' + k^2\alpha'^2)\frac{\omega m^2}{2}$$

$$+ \left(\frac{\alpha''''}{24} + \frac{\alpha''^2}{8} + \frac{\alpha'^2\alpha''}{4} + \frac{\alpha'\alpha'''}{6} + \frac{\alpha'^4}{24} - \frac{\tau'^2}{8}\right.$$

$$+ \frac{k\tau'''}{24} + \frac{k\alpha''\tau'}{24} - \frac{k\alpha'^2\tau'}{4} + \frac{k^2\alpha'\alpha'''}{6} + \frac{k^2\alpha'^2\alpha''}{4}$$

$$\left. - \frac{k^2\alpha'^4}{12} - \frac{k^3\alpha'^2\tau'}{4} - \frac{k^4\alpha'^4}{8}\right)\omega m^4 \quad \ldots\ldots\ldots \quad (36)$$

$$DP(x) = [\tau - k\alpha']\,\omega m + \left[\frac{\tau''}{6} - \frac{k\alpha'''}{6} + \frac{k(1+k^2)\alpha'^3}{3} + \frac{(1+k^2)\alpha'\tau'}{2}\right]\omega m^3 \quad \ldots\ldots\ldots \quad (37)$$

FM Path with Multiple AM/PM Converters

A general system configuration with multiple AM/PM converters is shown in Figure 2-2. Each of the linear transmission networks in the system will produce some spurious PM and AM components from the input PM according to equations (21) and (22).

The spurious components are sufficiently low so that superposition can be applied. Therefore, the direct DG and DP contributions of the linear networks, being independent of k in equations (33) and (34), will add at the output.

The AM component produced by a linear network will be affected by all the AM/PM converters placed after the linear network. However the AM will not be the same at the input of each converter since it will be reduced by the AM compression factors of any circuits placed between the distortion source and the actual AM/PM converter. Therefore $keff_r$, the effective conversion factor of the nonlinear circuit following the r-th linear network is given by the relationship:

$$keff_r = k_r + \gamma_r\,k_{r+1} + \gamma_r\,\gamma_{r+1}\,k_{r+2} + \ldots + \gamma_r \ldots \gamma_{n-1}\,k_n \quad \ldots\ldots\ldots \quad (38)$$

where k_r and γ_r stand for the conversion and compression factors of the r-th nonlinear network.

The AM produced by the linear networks is small, therefore each contribution can be calculated separately and added at the output.

Thus for the overall DG and DP characteristics of the system in Figure 2-2.

$$DG(x) = 1 + \sum_{r=1}^{n}\,[\alpha''_r(x) + keff_r\tau'_r(x)]\,\frac{\omega m^2}{2} - \tau'^2(x)\,\frac{\omega m^4}{8} \quad \ldots\ldots\ldots \quad (39)$$

$$DP(x) = \sum_{r=1}^{n}\,[\tau_r(x) - keff_r\alpha'_r(x)]\,\omega m \quad \ldots\ldots\ldots \quad (40)$$

APPENDIX A2 – GLOSSARY OF TERMS

GLOSSARY OF TERMS

AM Compression

The compression of the amplitude modulation on a carrier by a linear network. The ratio of the input to output AM index gives the compression factor of the network. This may be in the region of 30 – 40dB in the case of amplitude limiters.

AM to PM Conversion

Some nonlinear networks produce phase shifts dependent on the input signal amplitude. An amplitude modulated (AM) signal passing through such a network may consequently become phase modulated (PM). The amount of PM is dependent on the AM at the input and the conversion factor of the network. Typical AM to PM conversion factors range from 0.1 – 0.3^{o}/dB for good limiters, to 4 or 5^{o}/dB for some older travelling wave amplifiers.

Baseband

The frequency band containing information to be transmitted over the microwave radio (or cable) network. This may be a multichannel telephony, television or digital signal.

CCIR

The International Radio Consultative Committee. This is an international body whose recommendations on radio transmission procedures is generally accepted in Europe and many other parts of the world.

Closed Form Expression

An expression completely defining a function with a finite number of terms. The 3rd degree polynomial above would be a closed form expression if it completely defined f(n) ie, there are no neglected terms.

Differential Gain and Phase

These are defined in Section II, page 5.

First Order Theory

A theory whereby only the first (most important) term of a power series need be considered. For example, where:

$$f(n) = an + bn^2 + cn^3$$

The first order theory says,

$$f(n) = an$$

and all other terms may be neglected.

Frequency Division Multiplex (FDM)

The means by which the multichannel telephony signal is derived for an analogue transmission system. Individual telephone channels are formed into groups in the frequency domain and these groups are combined to form the baseband signal which may contain from 300 to 2700 channels.

Intermodulation Distortion

The interaction of component parts of a complex signal to produce distortion. This occurs when transmission parameters cause the lower frequency components to amplitude or frequency modulate the higher frequency components, producing unwanted sideband signals.

Intermodulation Noise

Wideband signals such as multichannel telephony consist of many 'random-like' components. The resultant intermodulation sidebands consequently produce a 'noise-like' spectrum.

Linear Distortion

Distortion of the form of amplitude and phase variations with frequency. This may change the envelope of a complex signal but will not produce harmonic or intermodulation components. Linear distortion is caused by linear networks.

Linear Network

A linear network is one that presents the same impedance at all amplitudes of the input signal. However, this impedance may be different at different frequencies.

Microwave Window

That part of the frequency spectrum suitable for wideband radio communications. This extends from around 1GHz up to 30GHz or more. The bottom end is mainly limited by atmospheric noise and the top end by atmospheric absorption and 'state of the art' hardware.

Nonlinear Distortion

Distortion of the form giving harmonic and/or intermodulation components. This may be caused by a nonlinear network AND/OR by a linear network followed by demodulation in the case of modulated transmission systems.

Nonlinear Network

A nonlinear network is one where the impedance is dependent on the amplitude of the input signal ie, it has a nonlinear voltage transfer characteristic.

Nyquist Principle

A principle on which the measurement of group delay is based. This is described in "Measurement of Phase Distortion" by H. Nyquist and S. Brand, BELL SYSTEM TECHNICAL JOURNAL No.9, Pages 522-549, 1930.

NTSC

The National Television System Committee. Their choice of television colour sub-carrier for use in their system is 3.58MHz.

PAL

Phase Alternate Line. This is a development of the NTSC television system and uses a colour sub-carrier of 4.43MHz.

Polynomial

A rational integral function which can be represented by a power series. The degree of the polynomial defines how many terms of the power series are present ie, 3rd degree polynomial in (n) may be:

$$f(n) = a + an + bn^2 + cn^3$$

ie, it contains dec, linear, quadratic and cubic terms.

Power Series

A series of the form $\sum a_n x^n$.

Stochastic Processes

This is a mathematical process based on random variables. Since the multichannel telephony signal is a random variable, mathematical representation of this signal and the distortion it undergoes must therefore be based on random or stochastic processes.

TVSC

The television sub-carrier containing the colour information for colour television. Other sub-carriers, placed above the video information, may carry sound information etc.

APPENDIX A3 – REFERENCES

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APPENDIX A4 – TABLES OF FIRST FACTOR NUMERICAL VALUES FOR SUBSTITUTION IN THE NOISE POWER EQUATIONS

These tables give the first factor numerical values for substitution in the noise power equations (Table 3-2). Each table is applicable to only one specific channel capacity (e.g., 1800). All the tables are based on a test-tone frequency of 2.4MHz and a measurement bandwidth of ±10MHz. However, correction factors are given with each table which allow calculations for any test-tone frequency and any bandwidth.

Table A4-1 First Factor Numerical Values in the Noise Power Equations of Table 3-2

Channel number	960	Test tone dev.	200 kHz rms
Baseband limits	60–4028 kHz	Multichannel dev.	1102 kHz rms

ω kHz	70	270	534	1248	2438	3886	Correction factor*
$G_1(\omega)$			0.073	0.73	3.39	7.94	$1/(m^2p^4)$
$\alpha_3(\omega)$		0.029	0.29	2.89	13.51	31.6	$1/(m^6)$
$\tau_{2k}(\omega)$					0.01	0.024	$1/(m^4)$
$P_1(\omega)$	0.051	0.783	2.84	8.05	9.98	9.2	$1/(m^2p^2)$
$\tau_1(\omega)$	0.038	0.59	2.12	6.02	7.46	6.87	$1/(m^2)$
$\alpha_{2k}(\omega)$	0.012	0.18	0.67	1.89	2.34	2.16	$1/(m^4)$
$G_2(\omega)$					0.087	0.22	$1/(m^4p^4)$
$\alpha_4(\omega)$			0.01	0.15	1.38	3.46	$1/(m^8)$
$\tau_{3k}(\omega)$							
$P_2(\omega)$				1.02	2.54	2.54	$1/(m^4p^2)$
$\tau_2(\omega)$			0.018	0.076	0.189	0.189	$1/(m^4)$
$\alpha_{3k}(\omega)$			0.012	0.053	0.13	0.13	$1/(m^6)$
$\tau_3(\omega)$				0.016	0.04	0.14	$1/(m^6)$
$\tau_{13}(\omega)$		0.058	0.203	0.59	1.07	1.94	$1/(m^4)$

Numbers not shown are less than 0.01

$$* \; m = \frac{x}{10} \quad ; \quad p = \frac{y}{2.4}$$

where, x = required sweep width (MHz)
y = required test-tone frequency (MHz)

Table A4-2 First Factor Numerical Values in the Noise Power Equations of Table 3-2

Channel number	1260	Test tone dev.	140 kHz rms
Baseband limits	60–5636 kHz	Multichannel dev.	884 kHz rms

ω kHz	70	270	534	1248	2438	3886	5340	Correction factor*
$G_1(\omega)$			0.036	0.72	4.23	10.46	18.85	$1/(m^2p^4)$
$\alpha_3(\omega)$		0.037	0.14	2.86	16.82	41.7	75	$1/(m^6)$
$\tau_{2k}(\omega)$					0.013	0.032	0.06	$1/(m^4)$
$P_1(\omega)$	0.032	0.506	1.98	7.15	12.26	12.16	11.54	$1/(m^2p^2)$
$\tau_1(\omega)$	0.024	0.38	1.48	5.34	9.15	9.08	8.62	$1/(m^2)$
$\alpha_{2k}(\omega)$		0.12	0.47	2.35	2.88	2.86	2.71	$1/(m^4)$
$G_2(\omega)$					0.049	0.19	0.34	$1/(m^4p^4)$
$\alpha_4(\omega)$				0.079	0.78	3.09	5.42	$1/(m^8)$
$\tau_{3k}(\omega)$								
$P_2(\omega)$				0.047	0.138	0.224	0.21	$1/(m^4p^2)$
$\tau_2(\omega)$				0.035	0.103	0.17	0.16	$1/(m^4)$
$\alpha_{3k}(\omega)$				0.024	0.072	0.12	0.11	$1/(m^6)$
$\tau_3(\omega)$			0.016	0.053	0.12	0.25	0.63	$1/(m^6)$
$\tau_{13}(\omega)$		0.07	0.29	0.99	1.98	2.99	4.63	$1/(m^4)$

Numbers not shown are less than 0.01

$$* \; m = \frac{x}{10} \quad ; \quad p = \frac{y}{2.4}$$

where, x = required sweep width (MHz)
y = required test-tone frequency (MHz)

Table A4-3 First Factor Numerical Values in the Noise Power Equations of Table 3-2

Channel number	1800	Test tone dev.	140 kHz rms
Baseband limits	316–8204 kHz	Multichannel dev.	1056 kHz rms

ω kHz	534	1248	2438	3886	5340	7600	8002	Correction factor*
$G_1(\omega)$	0.112	1.66	10.18	33.3	60.9	115	127	$1/(m^2p^4)$
$\alpha_3(\omega)$	0.44	6.63	40.5	133	243	456	508	$1/(m^6)$
$\tau_{2k}(\omega)$			0.031	0.1	0.18	0.34	0.38	$1/(m^4)$
$P_1(\omega)$	2.9	13.21	29.81	38.2	38.05	37.74	37.0	$1/(m^2p^2)$
$\tau_1(\omega)$	2.16	9.86	22.3	28.4	27.9	25.9	25.4	$1/(m^2)$
$\alpha_{2k}(\omega)$	0.68	3.1	7	8.94	8.78	8.16	8.15	$1/(m^4)$
$G_2(\omega)$		0.014	0.11	0.6	1.54	3.07	3.21	$1/(m^4p^4)$
$\alpha_4(\omega)$	0.012	0.22	1.81	9.56	24.5	48.9	51.2	$1/(m^8)$
$\tau_{3k}(\omega)$					0.01	0.021	0.022	$1/(m^6)$
$P_2(\omega)$		0.106	0.333	0.676	0.931	0.935	0.91	$1/(m^4p^2)$
$\tau_2(\omega)$	0.014	0.079	0.25	0.5	0.7	0.7	0.68	$1/(m^4)$
$\alpha_{3k}(\omega)$		0.055	0.17	0.35	0.49	0.49	0.47	$1/(m^6)$
$\tau_3(\omega)$	0.1	0.45	1.01	1.78	3.06	7.8	9.2	$1/(m^6)$
$\tau_{13}(\omega)$	0.88	3.88	8.92	13.8	18.2	28.3	31.4	$1/(m^4)$

Numbers not shown are less than 0.01

$$* \ m = \frac{x}{10} \quad ; \quad p = \frac{y}{2.4}$$

where, x = required sweep width (MHz)
y = required test-tone frequency (MHz)

Table A4-4 First Factor Numerical Values in the Noise Power Equations of Table 3-2

Channel number	2700	Test tone dev.	140 kHz rms
Baseband limits	316–12388 kHz	Multichannel dev.	1294 kHz rms

ω kHz	534	1248	2438	3886	5340	7600	11700	Correction factor*
$G_1(\omega)$	0.39	2	22.5	94.2	210	419	936	$1/(m^2p^4)$
$\alpha_3(\omega)$	1.54	7.95	89.5	375	836	1706	3726	$1/(m^6)$
$\tau_{2k}(\omega)$			0.067	0.28	0.63	1.29	2.81	$1/(m^4)$
$P_1(\omega)$	4.41	22.57	64.57	106.1	126.3	129.7	119.4	$1/(m^2p^2)$
$\tau_1(\omega)$	3.29	17	48.2	79.2	94.3	96.9	86.2	$1/(m^2)$
$\alpha_{2k}(\omega)$	1.04	5.34	15.2	24.9	29.7	30.5	28	$1/(m^4)$
$G_2(\omega)$		0.019	0.29	1.71	5.18	15.3	36.4	$1/(m^4p^4)$
$\alpha_4(\omega)$	0.06	0.31	4.55	27.3	82.6	243	580	$1/(m^8)$
$\tau_{3k}(\omega)$				0.012	0.035	0.103	0.25	$1/(m^6)$
$P_2(\omega)$	0.043	0.221	0.821	1.88	3.05	4.59	4.54	$1/(m^4p^2)$
$\tau_2(\omega)$	0.032	0.165	0.61	1.4	2.28	3.43	3.39	$1/(m^4)$
$\alpha_{3k}(\omega)$	0.022	0.12	0.42	0.98	1.59	2.39	2.37	$1/(m^6)$
$\tau_3(\omega)$	0.83	4.27	11.11	19.12	27.88	48.22	150	$1/(m^6)$
$\tau_{13}(\omega)$	3.06	15.83	43	73.5	99	134	230	$1/(m^4)$

Numbers not shown are less than 0.01

$$* \ m = \frac{x}{10} \quad ; \quad p = \frac{y}{2.4}$$

where, x = required sweep width (MHz)
y = required test-tone frequency (MHz)

APPENDIX A5 – NOMOGRAMS FOR THE NOISE POWER EQUATIONS

As an aid to the rapid determination of intermodulation noise, equations (3-1), (3-2) and (3-3) in Table 3-2 have been plotted as a series of nomograms. These nomograms will give the pWOp value of the intermodulation noise power as a function of the response deviations at ±10MHz. Log-log co-ordinate systems are used throughout and the quadratic dependence of noise power on the response deviation results in straight line plots. In addition to the logarithmic pWOp scale, a linear dB scale is provided which enables direct 'noise power ratio' (NPR) readings if required. The abscissa scale is calibrated in practical deviation units (ns, dB, %, deg.), corresponding to equations (3-1), (3-2) and (3-3), and can be directly read from the MLA display. Each nomogram is accompanied by a simple block diagram and response shapes, as a quick reminder of the field of application and of the notation.

The nomograms are intended to evaluate the noise originating from different kinds of distortion for a specific channel number. As an example consider N = 1800 channels, which is representative of modern high capacity microwave links. Nomograms shown in Figures A5-12, 13, 14 and 15 are suitable for determining noise originating from the equivalent characteristics. In each of these figures, terms shown in one row of equations (3-1), (3-2) and (3-3) are plotted – Figure A5-12 presents the terms of the first row, Figure A5-13 the terms in the second row etc. Rapid comparison of noise contributions originating from equivalent characteristics in thus possible.

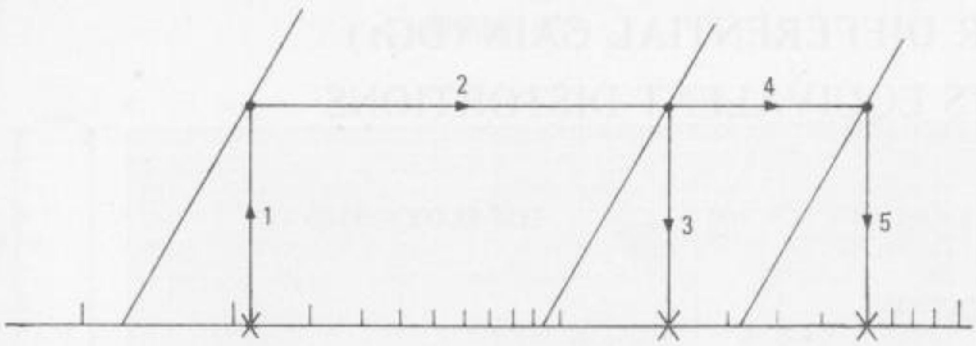


Figure A5-1 Method for Calculating Equivalent Deviations Using "EQUIVALENT DISTORTIONS" Nomograms

Step 1 – Enter vertically upward at selected kind of deviation value, and find the intersection with corresponding plot.

Step 2 – From this first intersection, proceed horizontally and find intersection with the equivalent plot pertaining to the same slot frequency.

Step 3 – From this second intersection, proceed vertically downward and find intersection with abscissa scale. This will give the equivalent deviation.

Step 4 – If needed, from the second intersection proceed horizontally again and find the intersection with the next equivalent plot pertaining to the same slot frequency.

Step 5 – From this third intersection, proceed vertically downwards and again find the intersection with the abscissa scale. This will give the other equivalent deviation.

Numerical Example: For N = 1800, the equivalent DG_1 distortions are;

Linear differential gain	DG_1 = 0.4%
Cubic gain	α_3 = 0.2dB
Parabolic delay +AM/PM	τ_{2k}k = 7(ns. °/dB)

(see Figure A5-12)

LINEAR DIFFERENTIAL GAIN (DG_1) AND ITS EQUIVALENT DISTORTIONS

CHANNEL CAPACITY N = 960 **TOP SLOT = 3886 kHz**

CONFIGURATION:

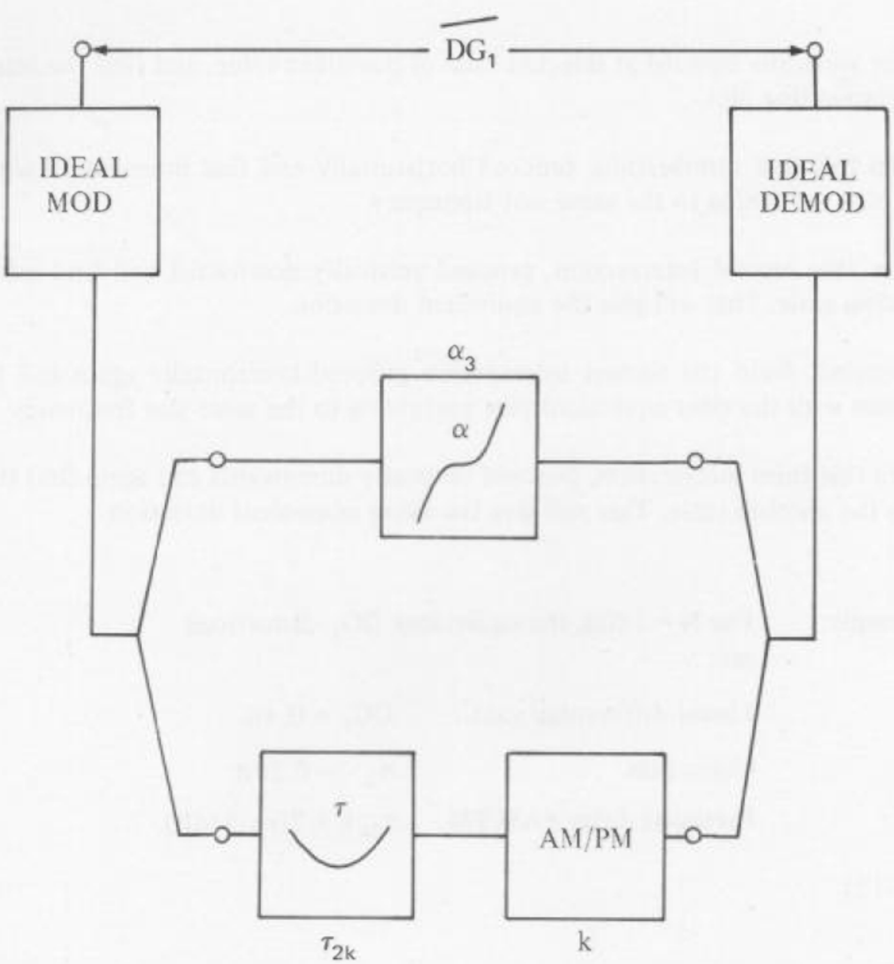
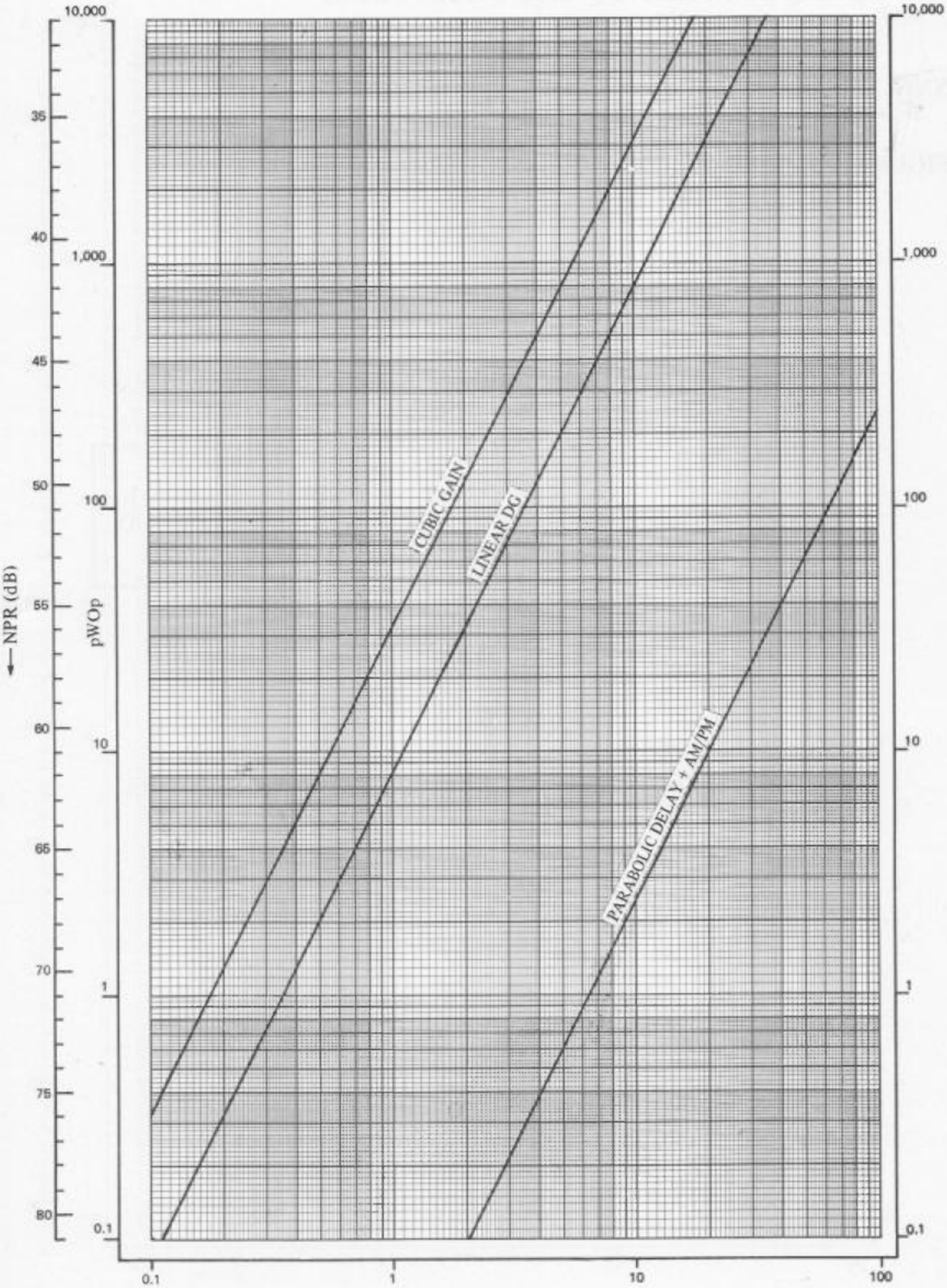


FIGURE A5-2



DG_1 (%) α_3 (dB) [τ_{2k} (ns) k(°/dB)]

PARABOLIC DIFFERENTIAL GAIN (DG_2) AND ITS EQUIVALENT DISTORTIONS

CHANNEL CAPACITY N = 960 **TOP SLOT = 3886 kHz**

CONFIGURATION:

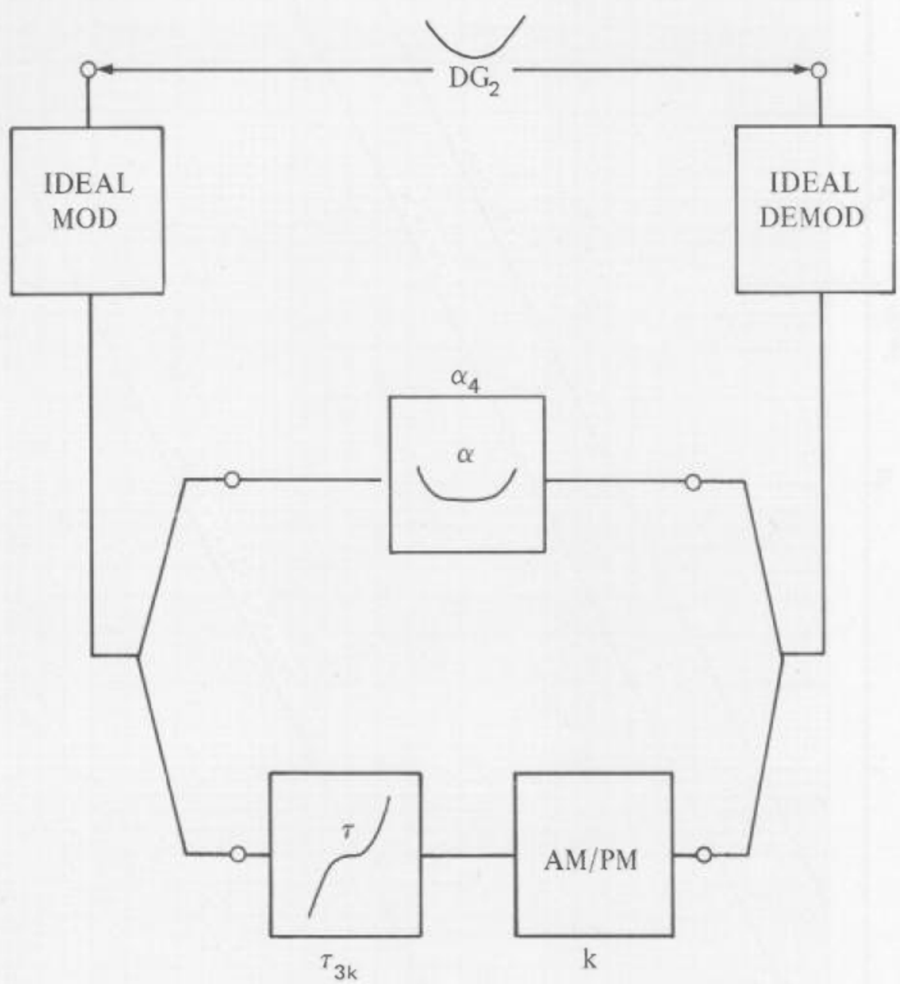
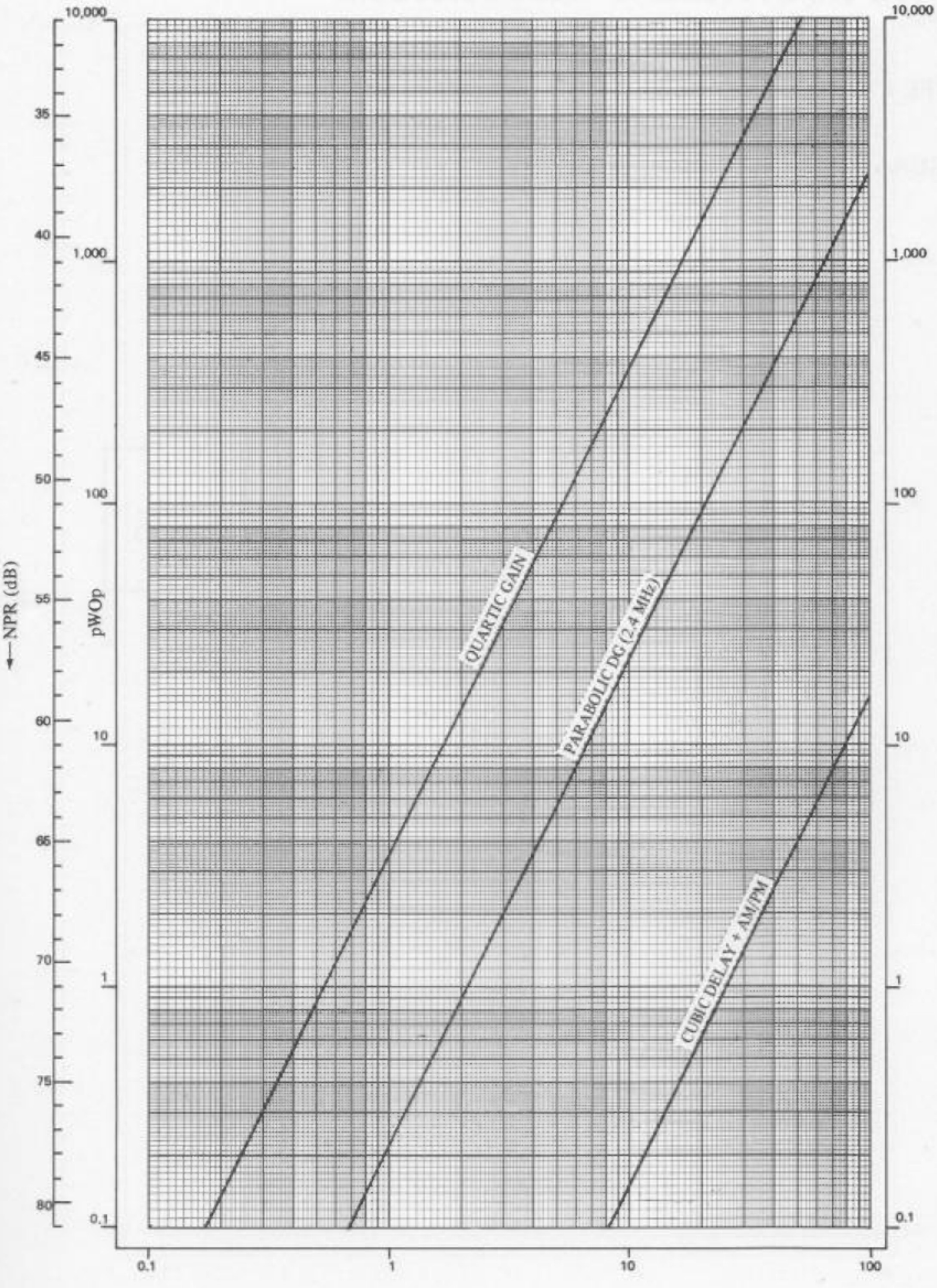


FIGURE A5-3



DG_2 (%) α_4 (dB) [τ_{3k} (ns) k(°/dB)]

LINEAR DIFFERENTIAL PHASE (DP_1) AND ITS EQUIVALENT DISTORTIONS

CHANNEL CAPACITY N = 960 **TOP SLOT = 3886 kHz**

CONFIGURATION:

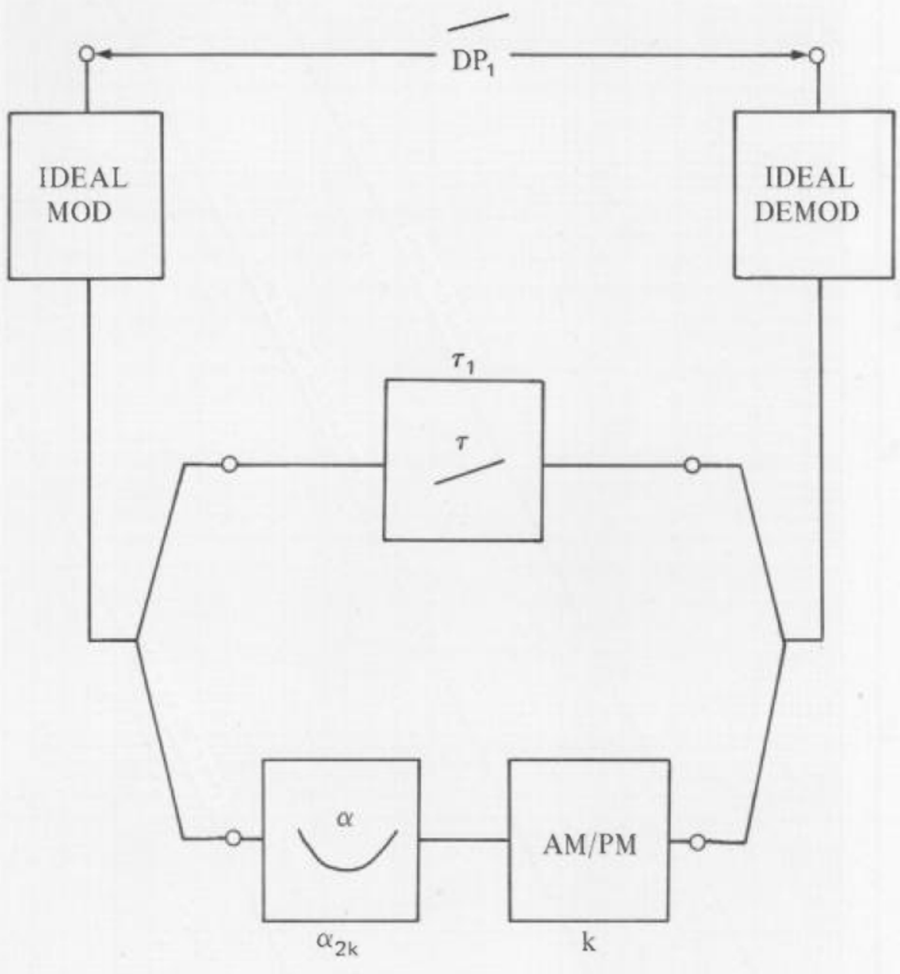
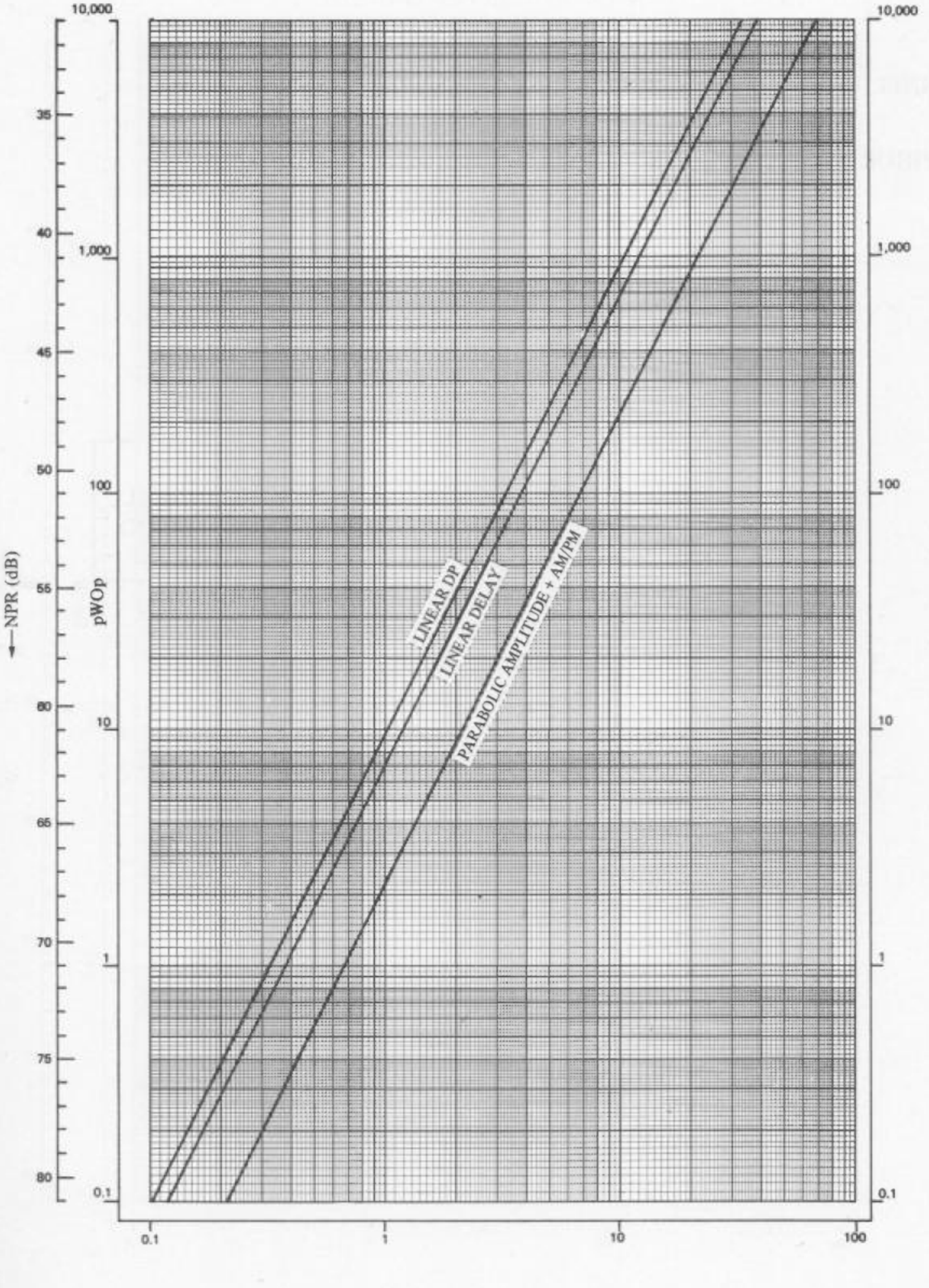


FIGURE A5-4



DP_1 (°) τ_1 (ns) [α_{2k} (dB) k(°/dB)]

PARABOLIC DIFFERENTIAL PHASE (DP_2) AND ITS EQUIVALENT DISTORTIONS

CHANNEL CAPACITY N = 960 **TOP SLOT = 3886 kHz**

CONFIGURATION:

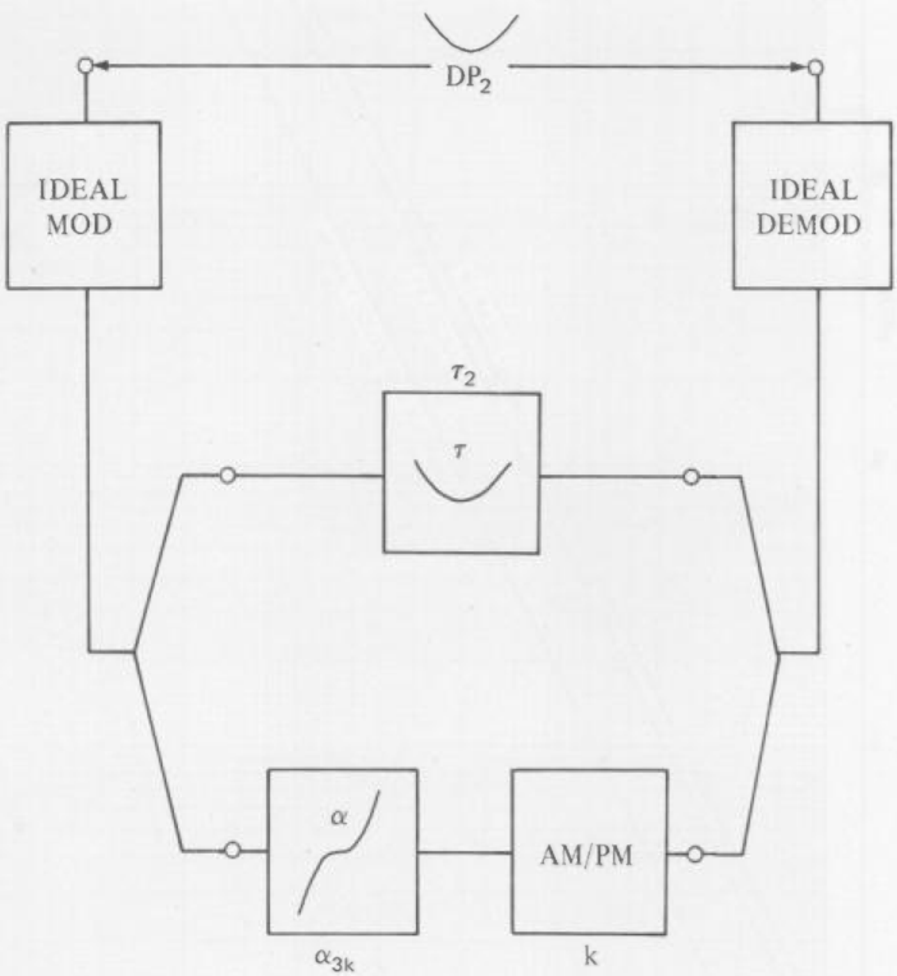
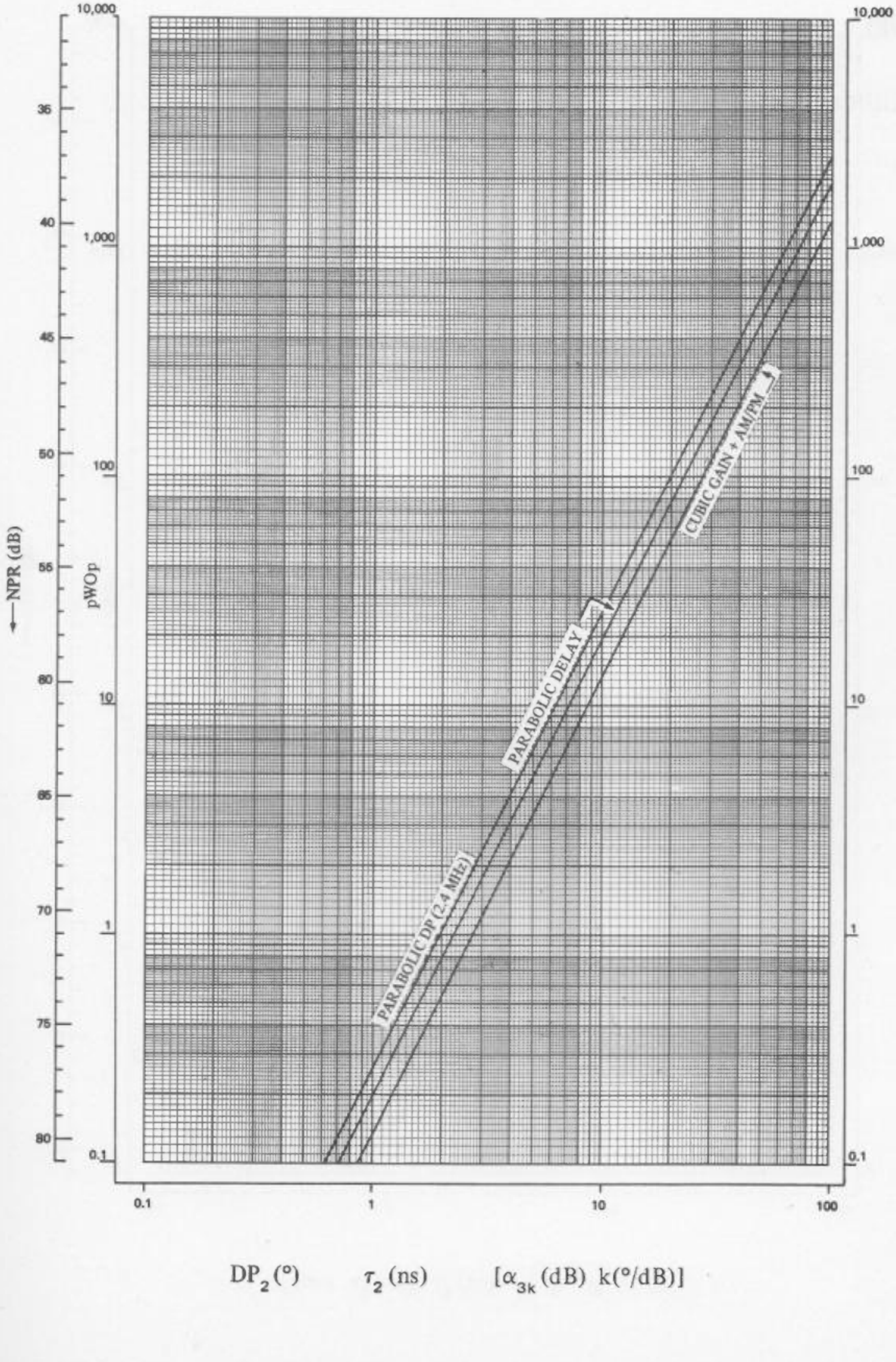


FIGURE A5-5



←NPR (dB)
pWOp
35
40
45
50
55
60
65
70
75
80
10,000
1,000
100
10
1
0.1
0.1
1
10
100
PARABOLIC DP (2.4 MHz)
PARABOLIC DELAY
CUBIC GAIN + AM/PM
DP₂ (°)   τ₂ (ns)   [α₃ₖ (dB) k(°/dB)]

GROUP DELAY DISTORTIONS (GD)

CHANNEL CAPACITY N = 960 **TOP SLOT = 3886 kHz**

CONFIGURATION:

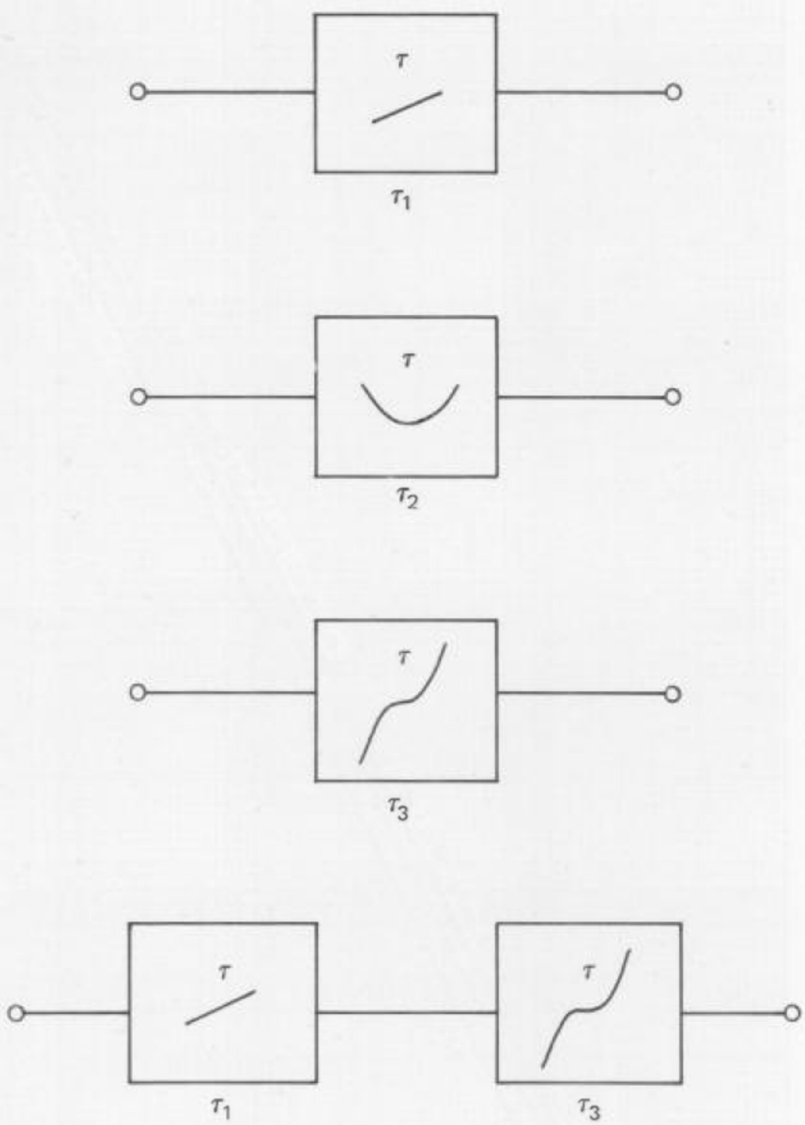
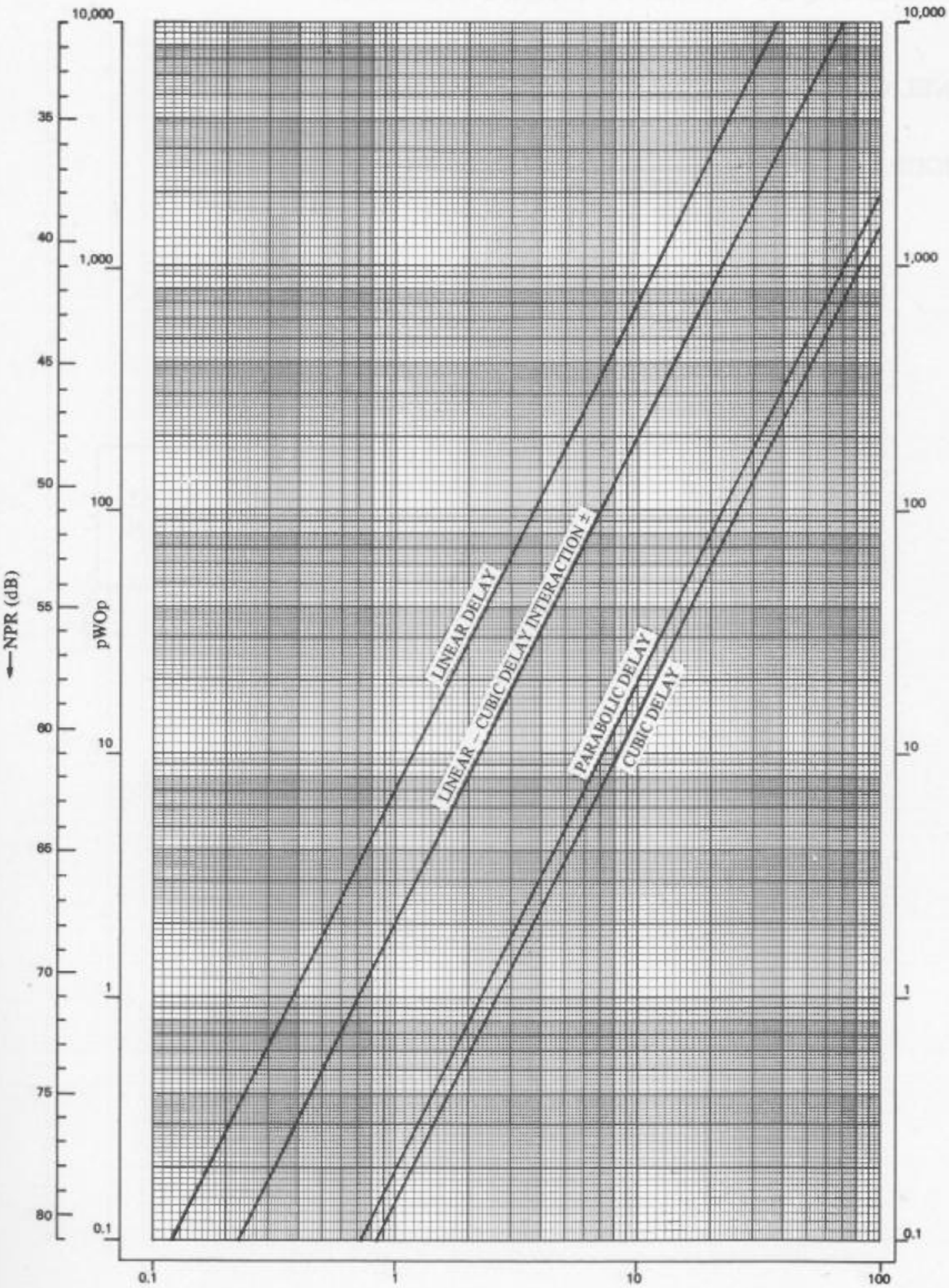


FIGURE A5-6



τ_1 (ns), τ_2 (ns), τ_3 (ns), $\sqrt{\tau_1 \quad \tau_3}$ (ns)

LINEAR DIFFERENTIAL GAIN (DG1) AND ITS EQUIVALENT DISTORTIONS

CHANNEL CAPACITY N = 1260 **TOP SLOT = 5340 kHz**

CONFIGURATION:

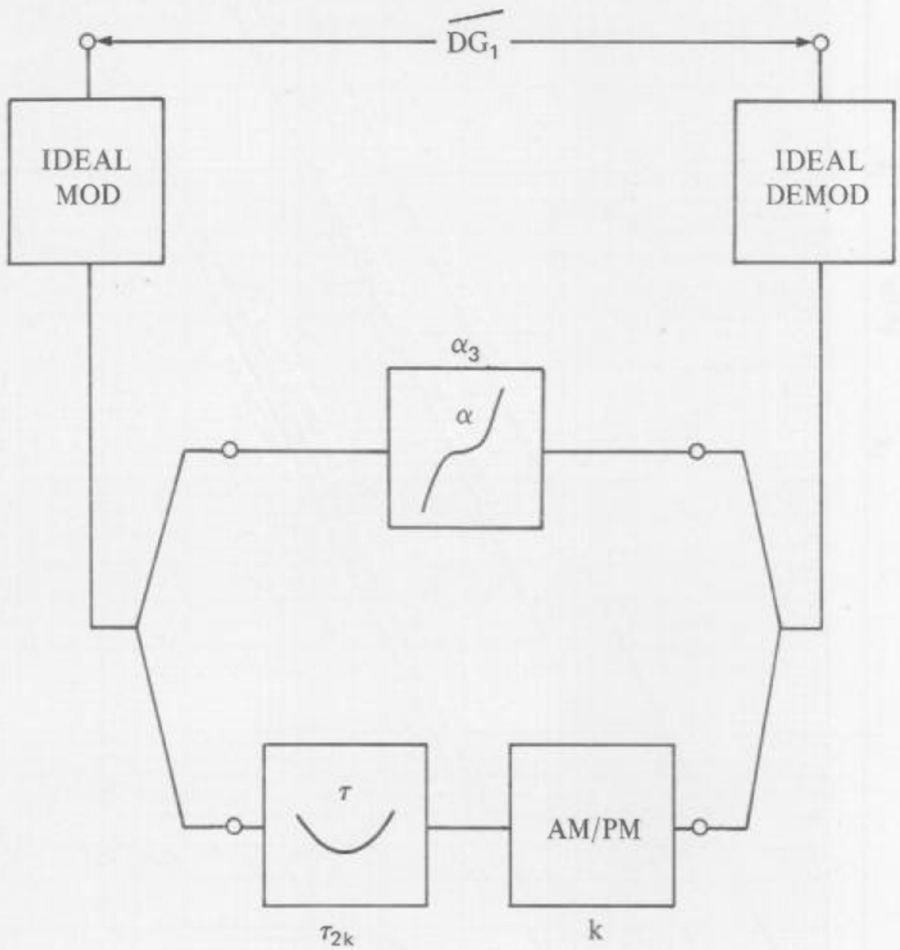
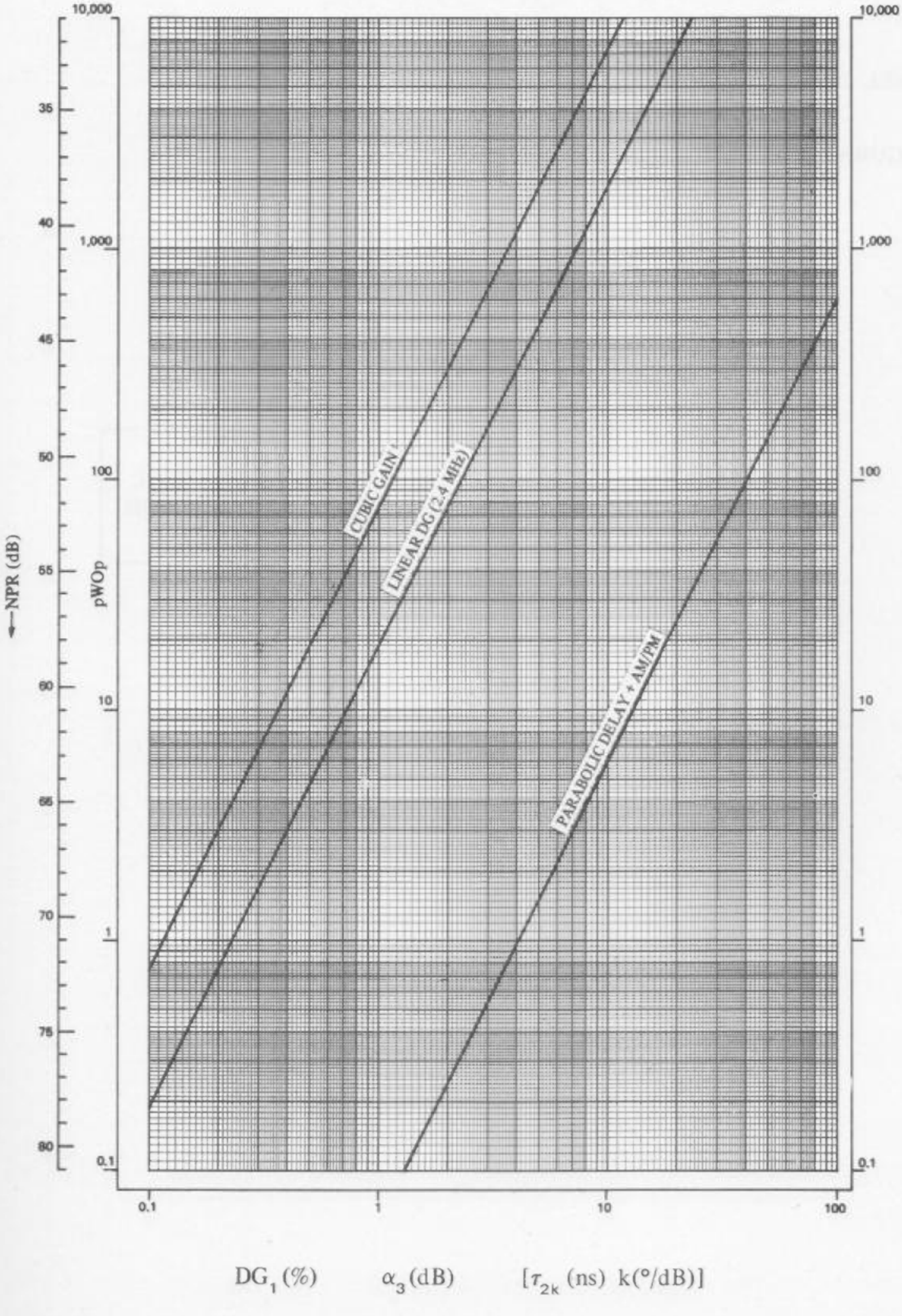


FIGURE A5-7



10,000
35
40
1,000
45
50
100
55
60
10
65
70
1
75
80
0.1
←NPR (dB)
pWOp
CUBIC GAIN
LINEAR DG (2.4 MHz)
PARABOLIC DELAY + AM/PM
0.1
1
10
100
DG₁ (%)   α₃ (dB)   [τ₂ₖ (ns) k(°/dB)]

PARABOLIC DIFFERENTIAL GAIN (DG_2) AND ITS EQUIVALENT DISTORTIONS

CHANNEL CAPACITY N = 1260 **TOP SLOT = 5340 kHz**

CONFIGURATION:

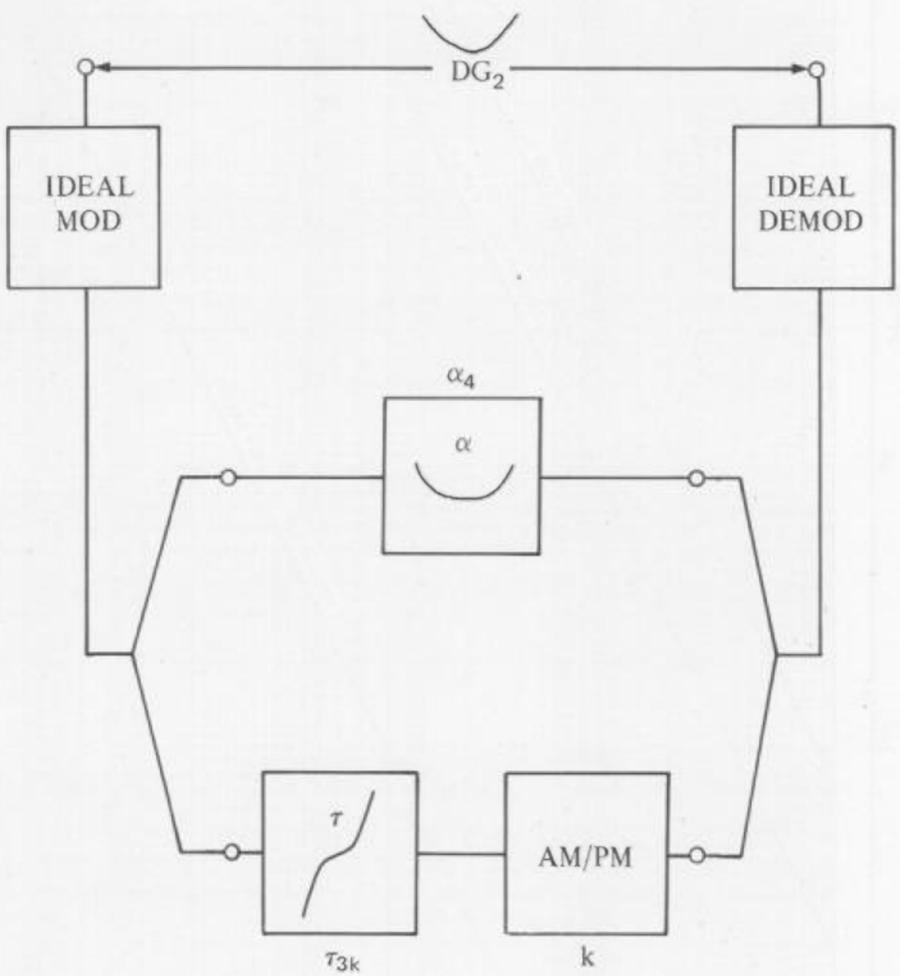
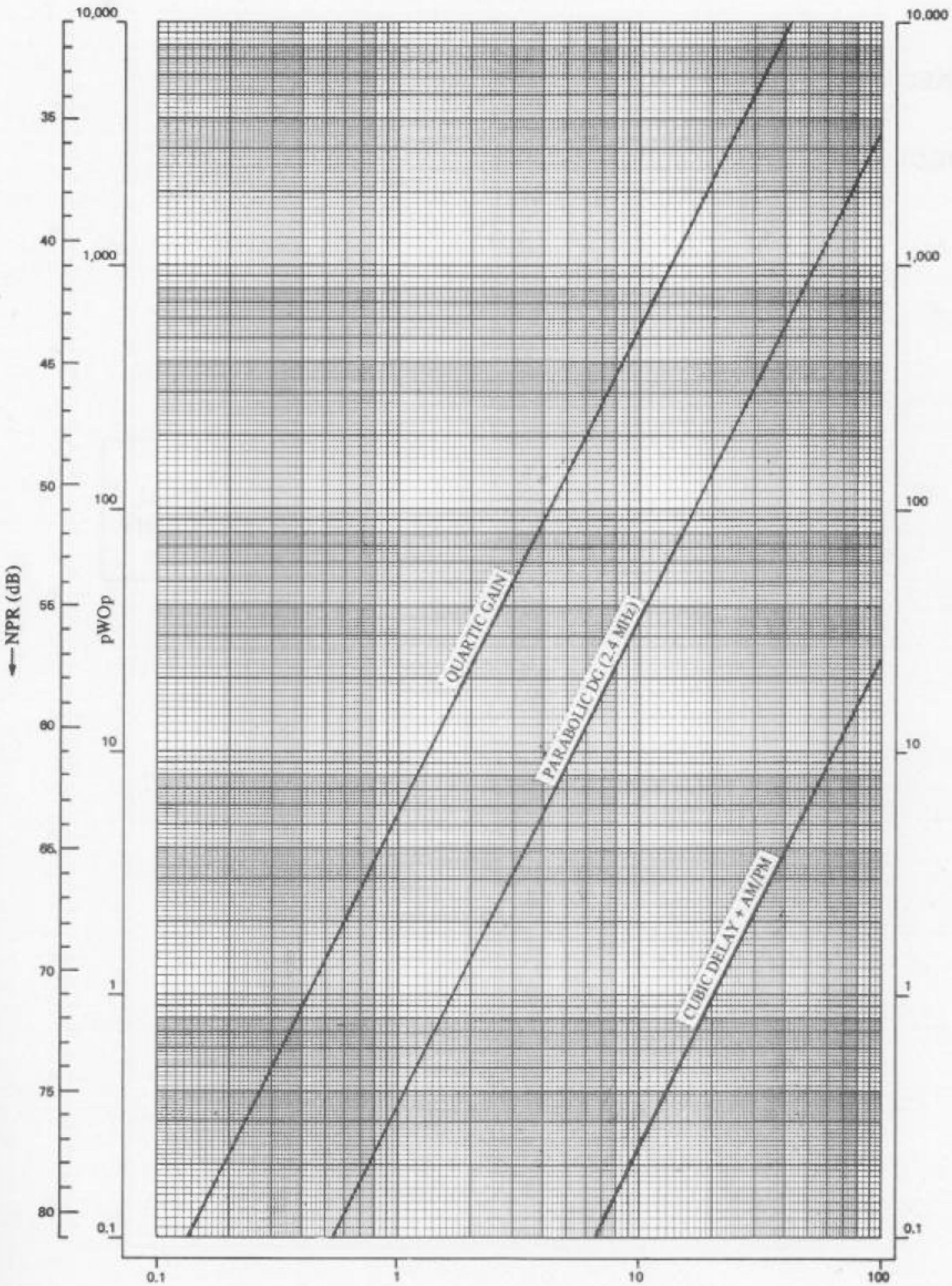


FIGURE A5-8



DG_2 (%) α_4 (dB) [τ_{3k} (ns) k(°/dB)]

LINEAR DIFFERENTIAL PHASE (DP_1) AND ITS EQUIVALENT DISTORTIONS

CHANNEL CAPACITY N = 1260 **TOP SLOT = 5340 kHz**

CONFIGURATION:

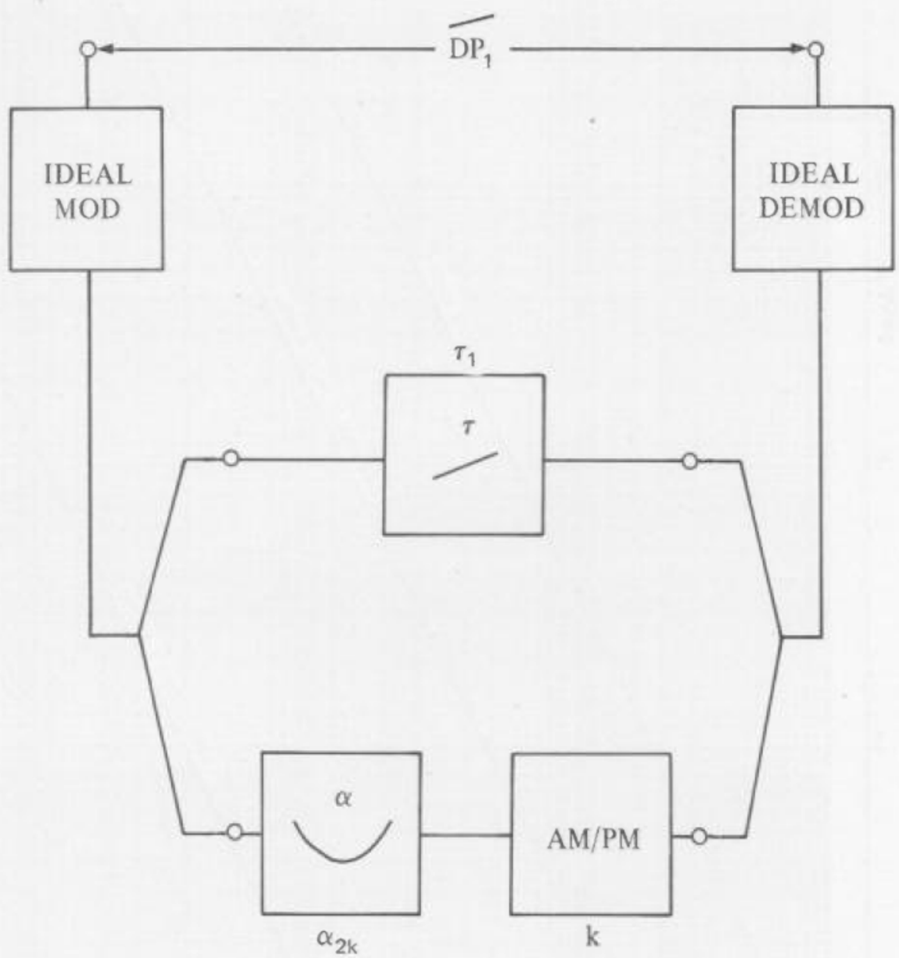
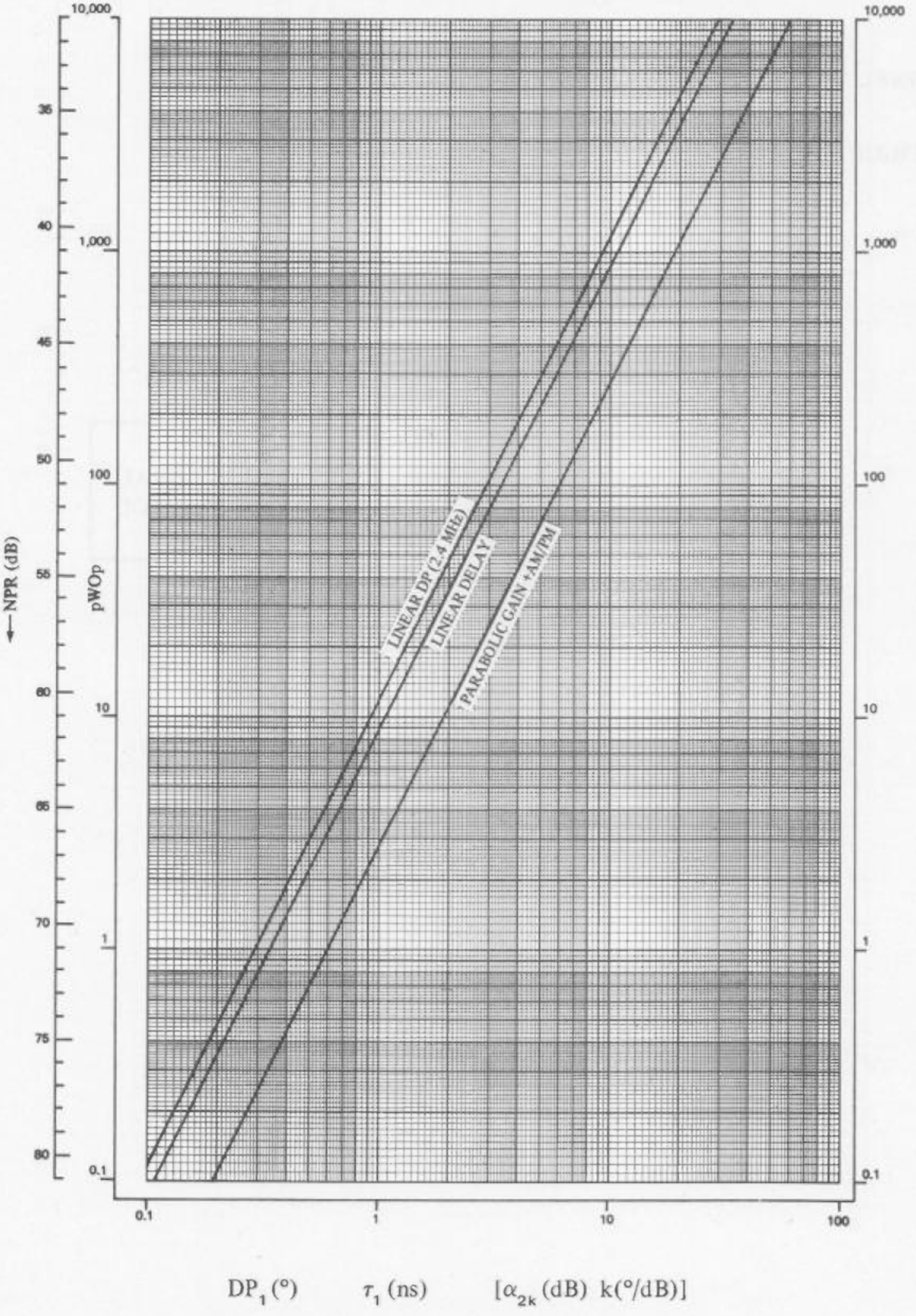


FIGURE A5-9



← NPR (dB)
pWOp
10,000
1,000
100
10
1
0.1
35
40
45
50
55
60
65
70
75
80
0.1
1
10
100
LINEAR DP (2.4 MHz)
LINEAR DELAY
PARABOLIC GAIN +AM/PM
DP₁ (°)   τ₁ (ns)   [α₂ₖ (dB) k(°/dB)]

PARABOLIC DIFFERENTIAL PHASE (DP2) AND ITS EQUIVALENT DISTORTIONS

CHANNEL CAPACITY N = 1260 **TOP SLOT = 5340 kHz**

CONFIGURATION:

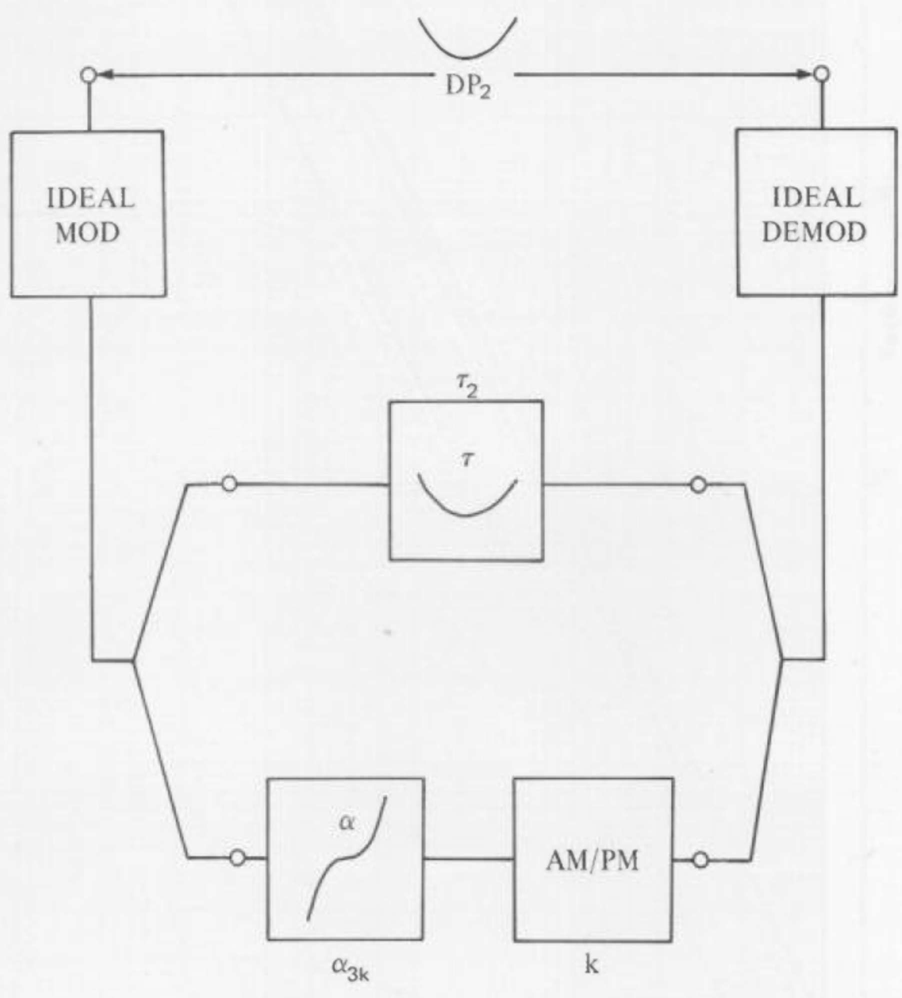
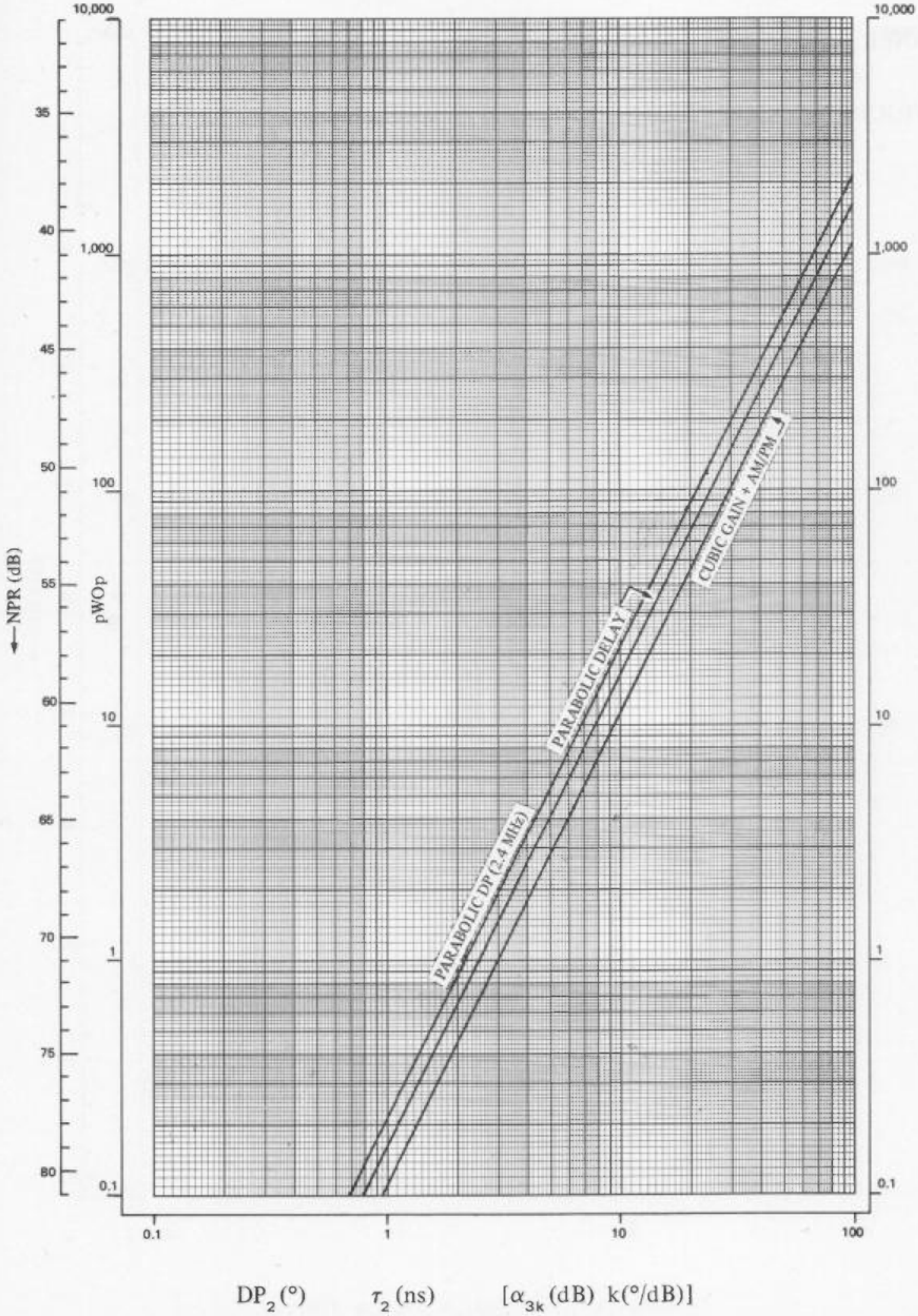


FIGURE A5-10



10,000
1,000
100
10
1
0.1
35
40
45
50
55
60
65
70
75
80
← NPR (dB)
pWOp
0.1
1
10
100
PARABOLIC DP (2.4 MHz)
PARABOLIC DELAY
CUBIC GAIN + AM/PM
DP$_2$ (°) τ_2 (ns) [α_{3k} (dB) k(°/dB)]

GROUP DELAY DISTORTIONS (GD)

CHANNEL CAPACITY N = 1260 **TOP SLOT = 5340 kHz**

CONFIGURATION:

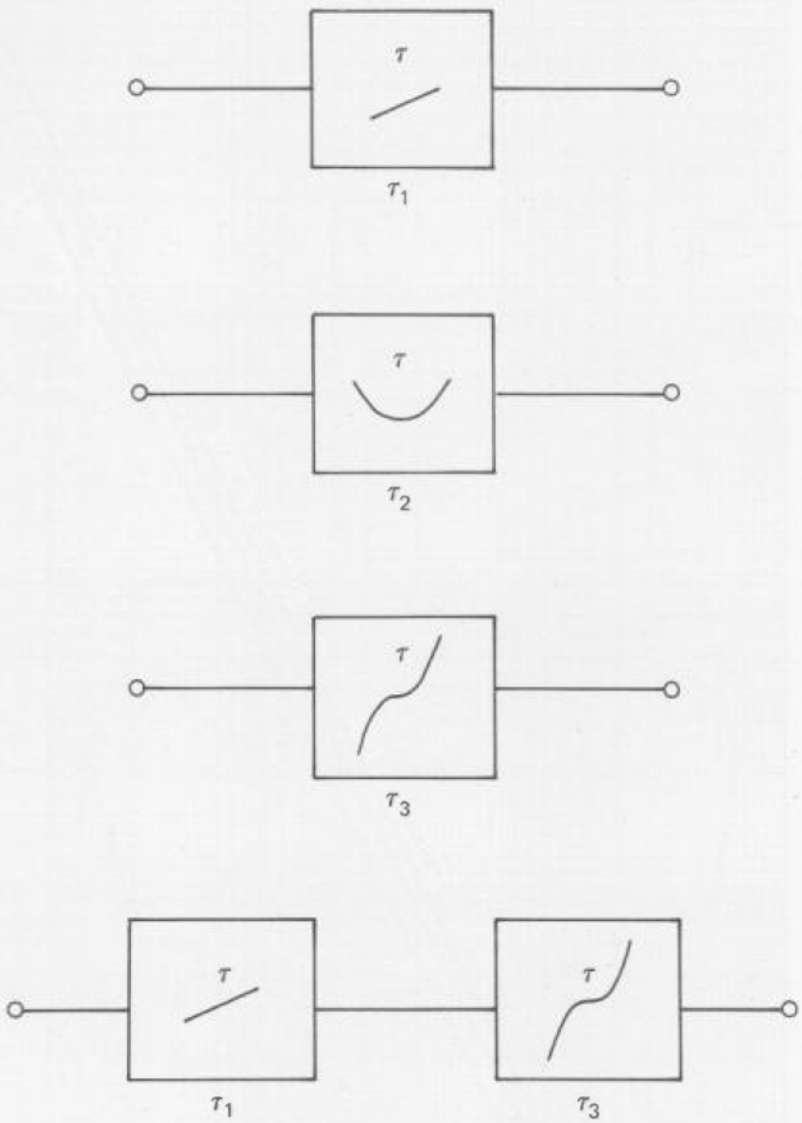
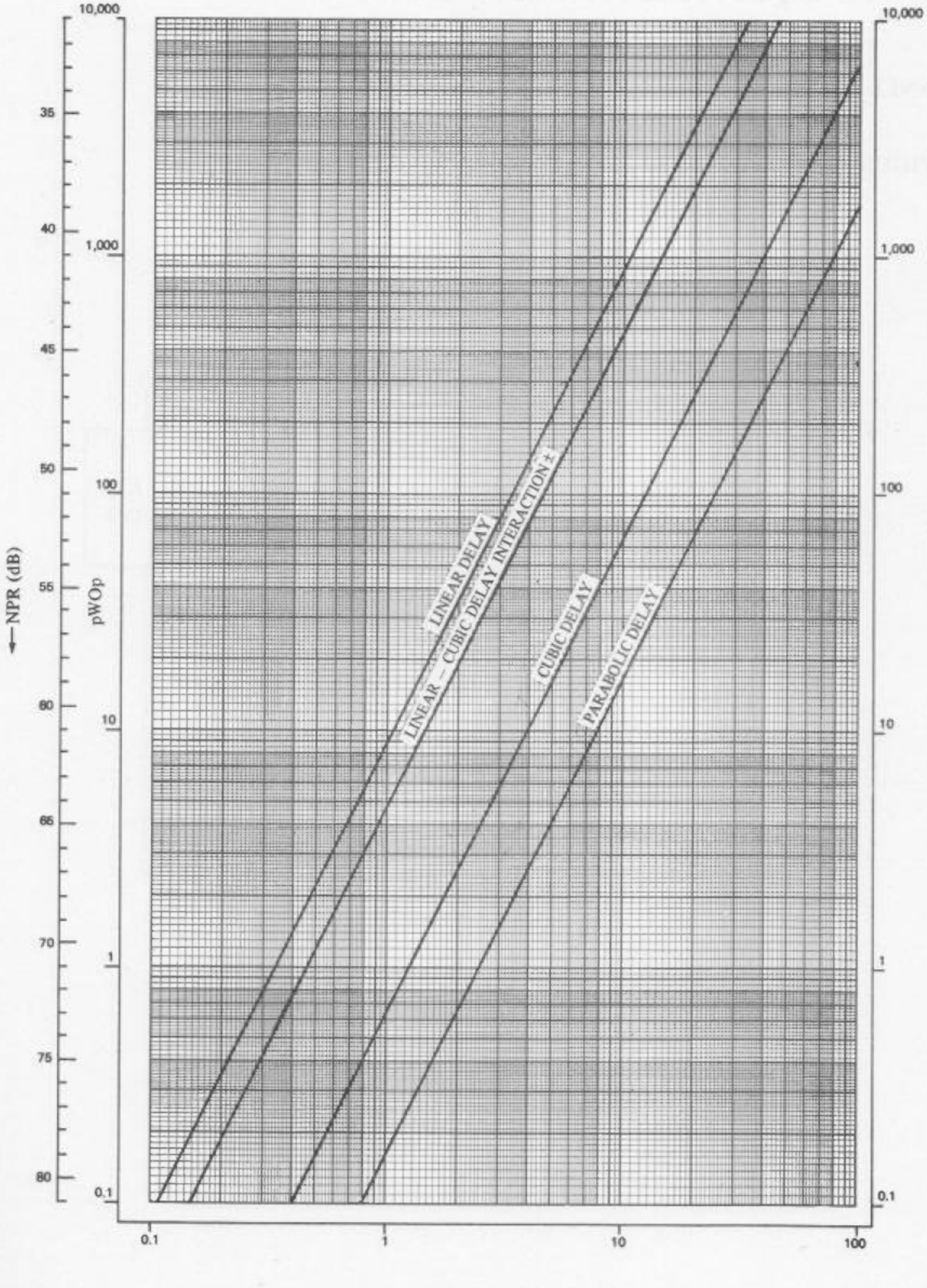


FIGURE A5-11



τ_1 (ns), τ_2 (ns), τ_3 (ns), $\sqrt{\tau_1 \cdot \tau_3}$ (ns)

LINEAR DIFFERENTIAL GAIN (DG1) AND ITS EQUIVALENT DISTORTIONS

CHANNEL CAPACITY N = 1800 **TOP SLOT = 7600 kHz**

CONFIGURATION:

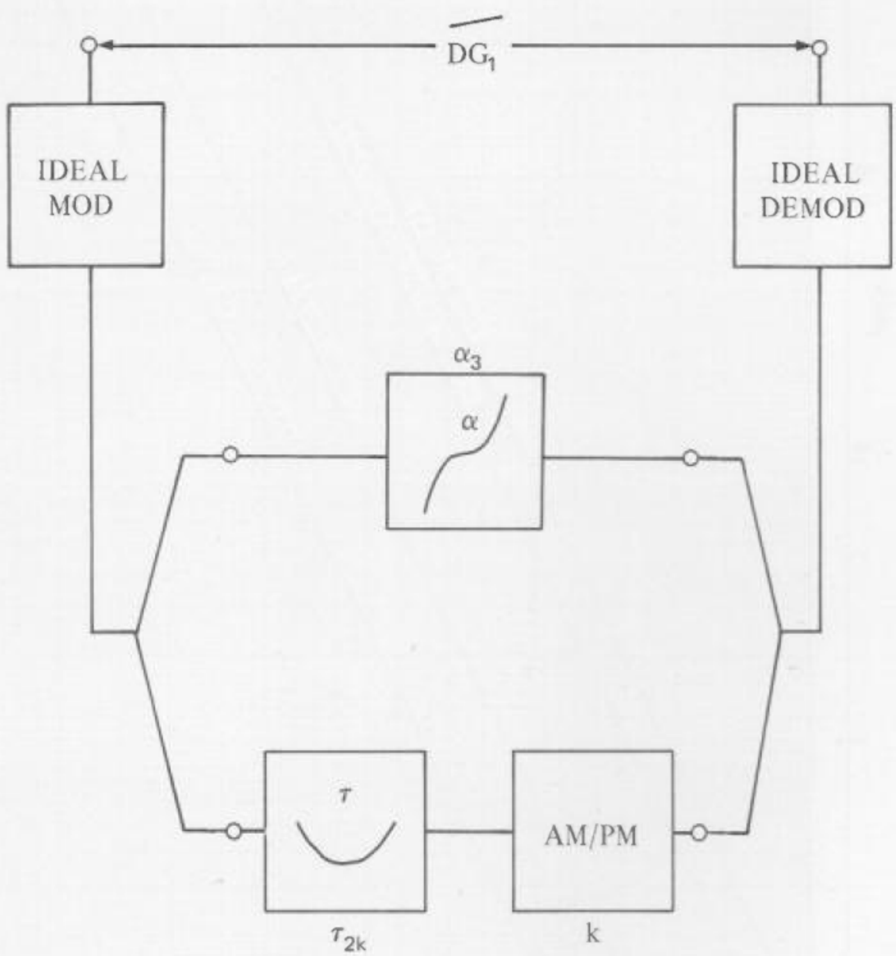
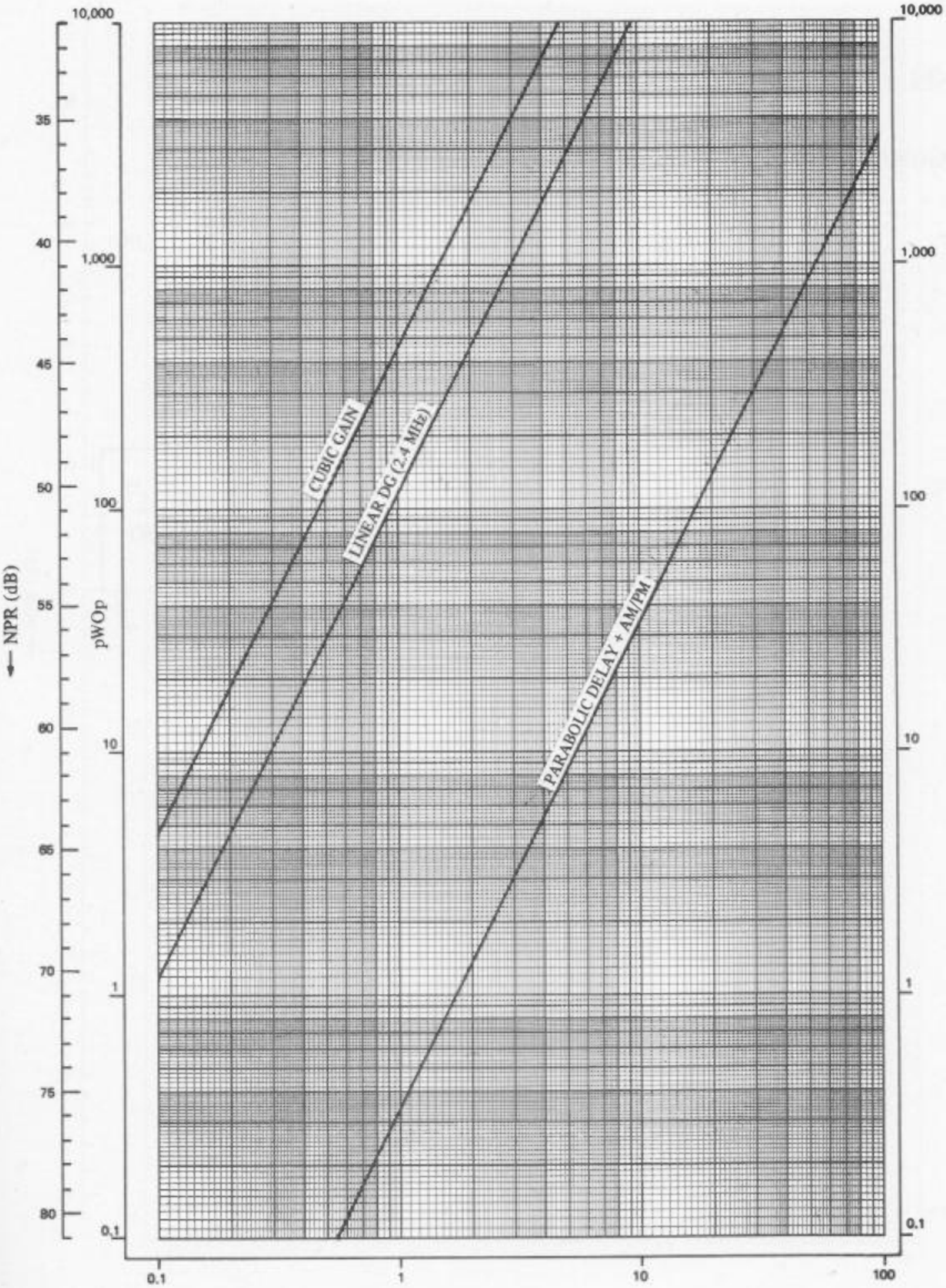


FIGURE A5-12



DG_1 (%) α_3 (dB)   [τ_{2k} (ns) k(°/dB)]

PARABOLIC DIFFERENTIAL GAIN (DG_2) AND ITS EQUIVALENT DISTORTIONS

CHANNEL CAPACITY N = 1800 **TOP SLOT = 7600 kHz**

CONFIGURATION:

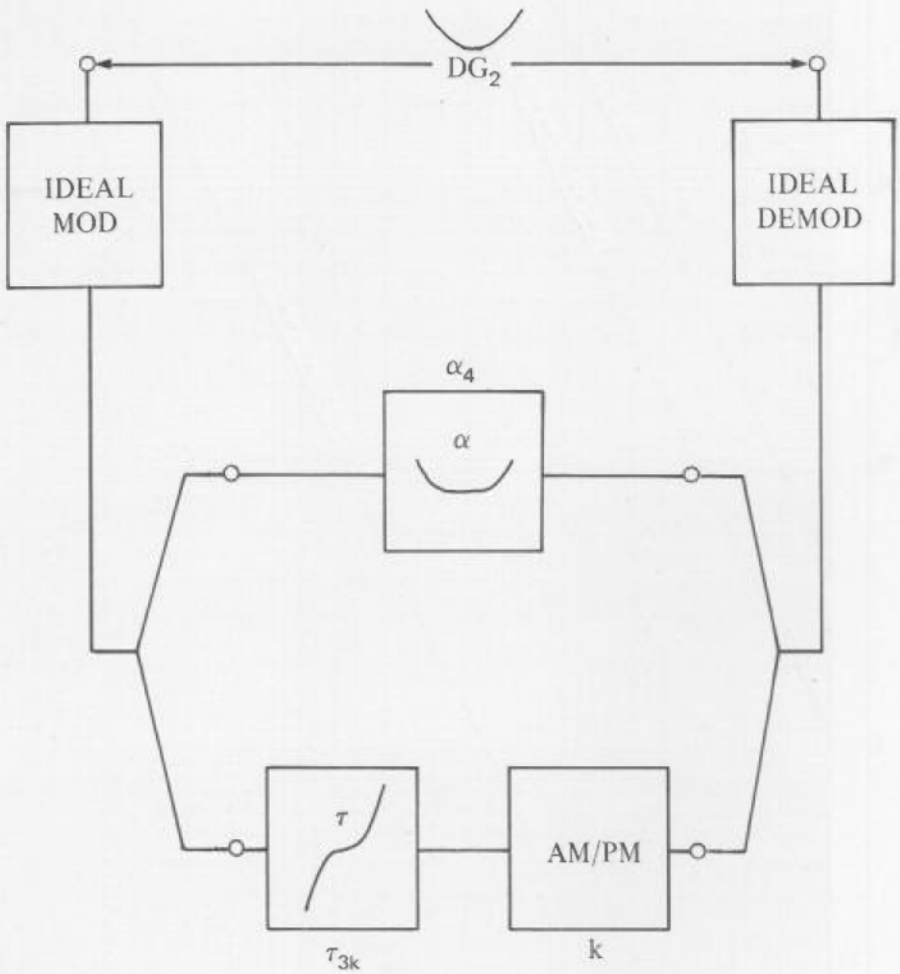
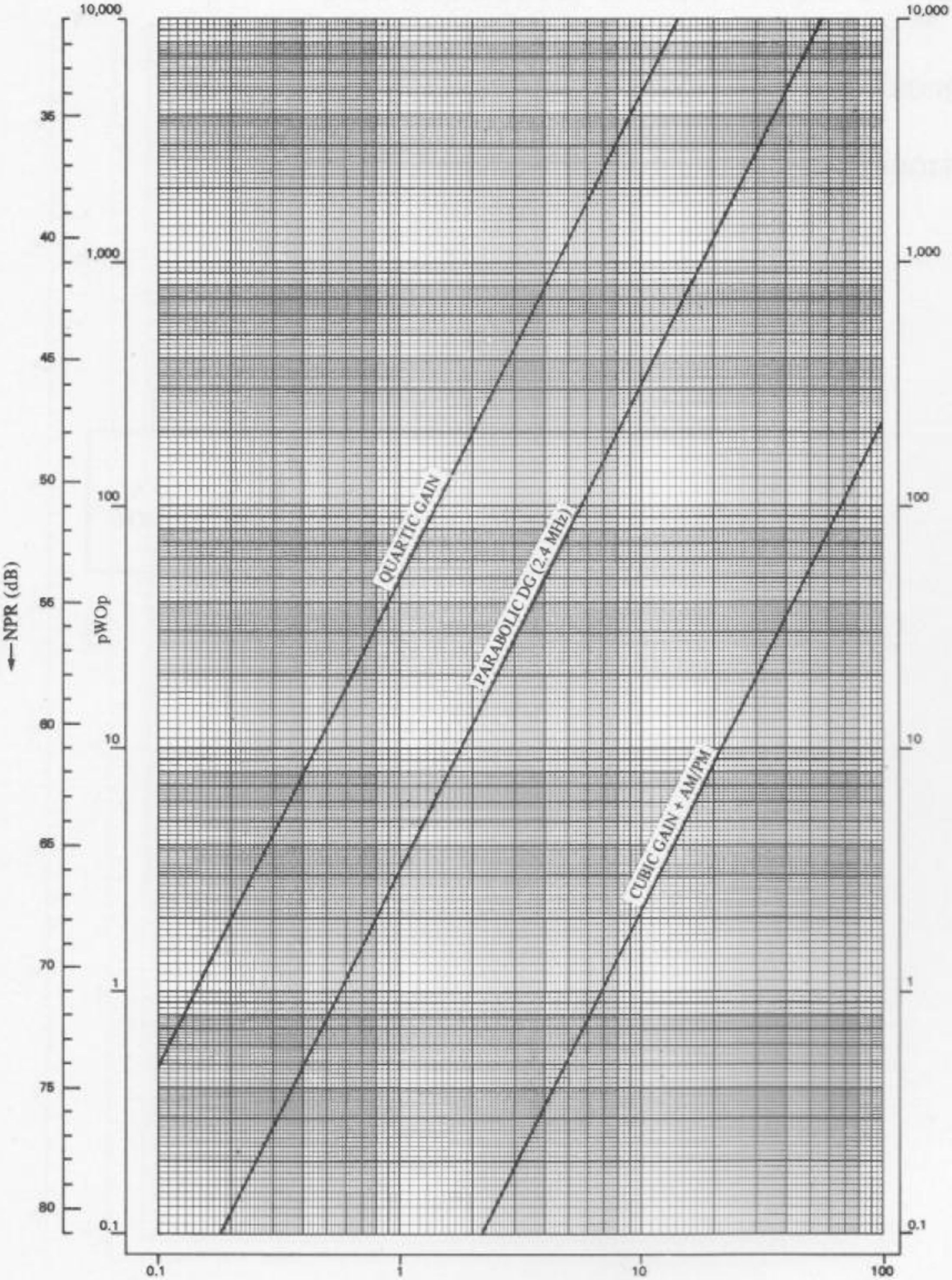


FIGURE A5-13



DG_2 (%) α_4 (dB) [τ_{3k} (ns) k(°/dB)]

LINEAR DIFFERENTIAL PHASE (DP_1) AND ITS EQUIVALENT DISTORTIONS

CHANNEL CAPACITY N = 1800 **TOP SLOT = 7600 kHz**

CONFIGURATION:

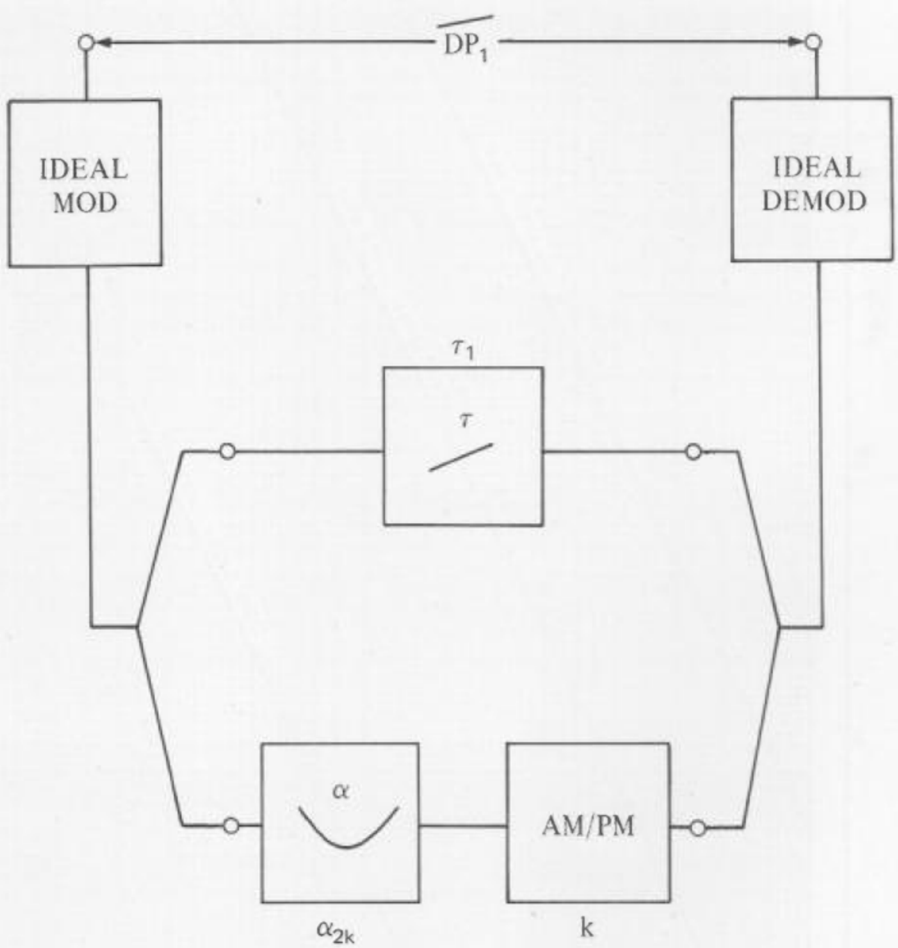
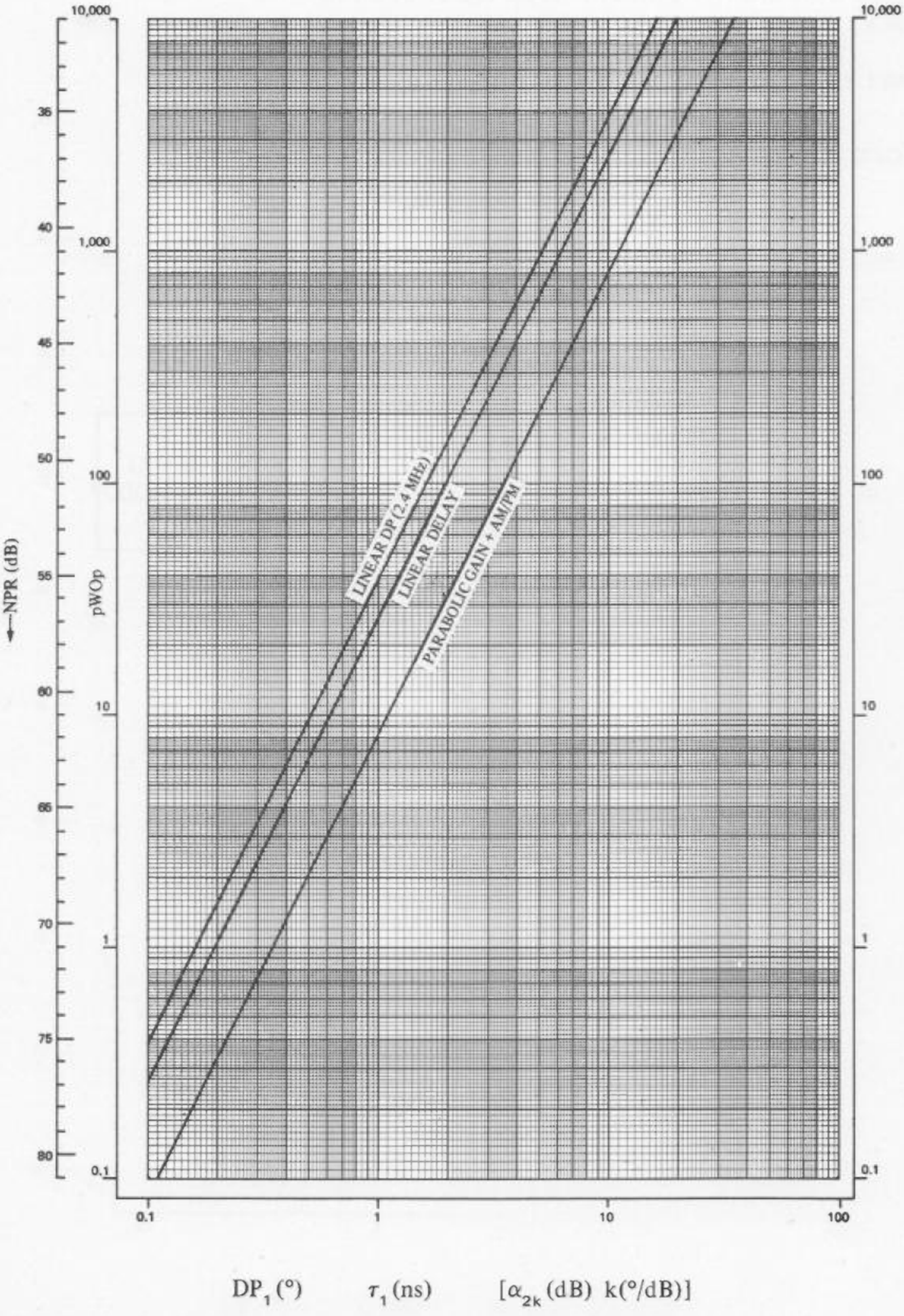


FIGURE A5-14



10,000
36
40
1,000
45
50
100
55
60
10
65
70
1
75
80
0.1
← NPR (dB)
pWOp
LINEAR DP (2.4 MHz)
LINEAR DELAY
PARABOLIC GAIN + AM/PM
0.1
1
10
100
DP1 (°)
τ1 (ns)
[α2k (dB) k(°/dB)]

PARABOLIC DIFFERENTIAL PHASE (DP_2) AND ITS EQUIVALENT DISTORTIONS

CHANNEL CAPACITY N = 1800 **TOP SLOT = 7600 kHz**

CONFIGURATION:

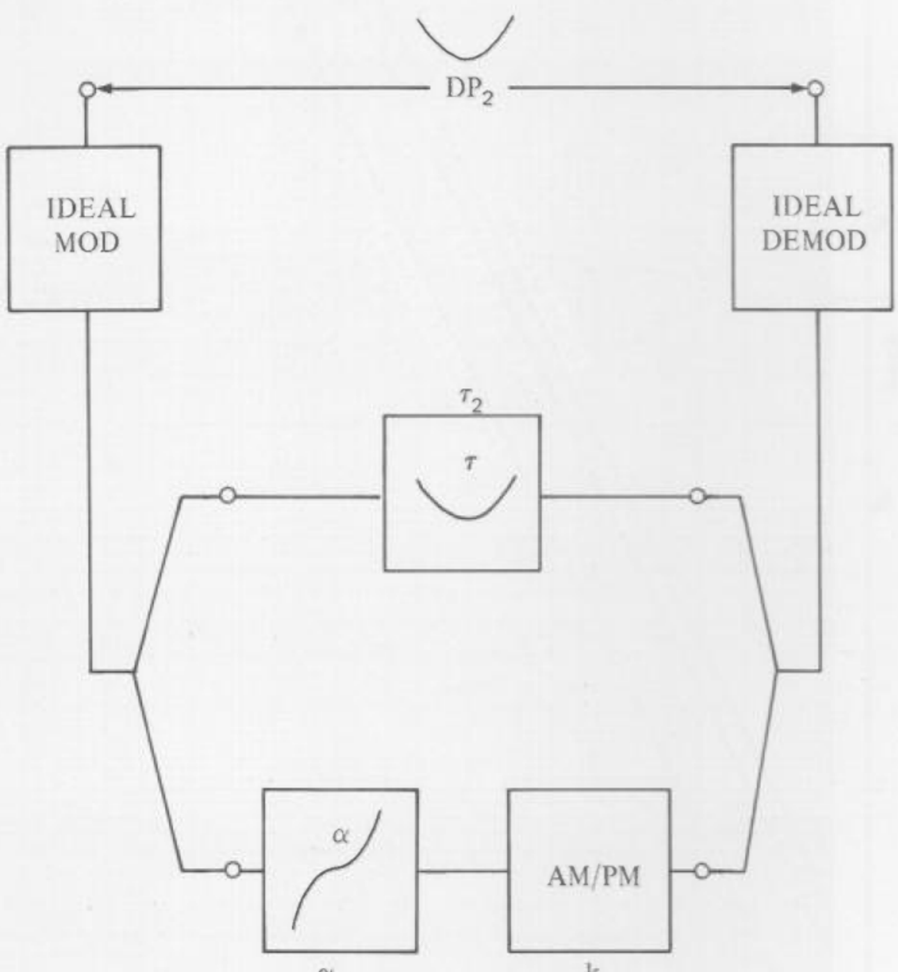
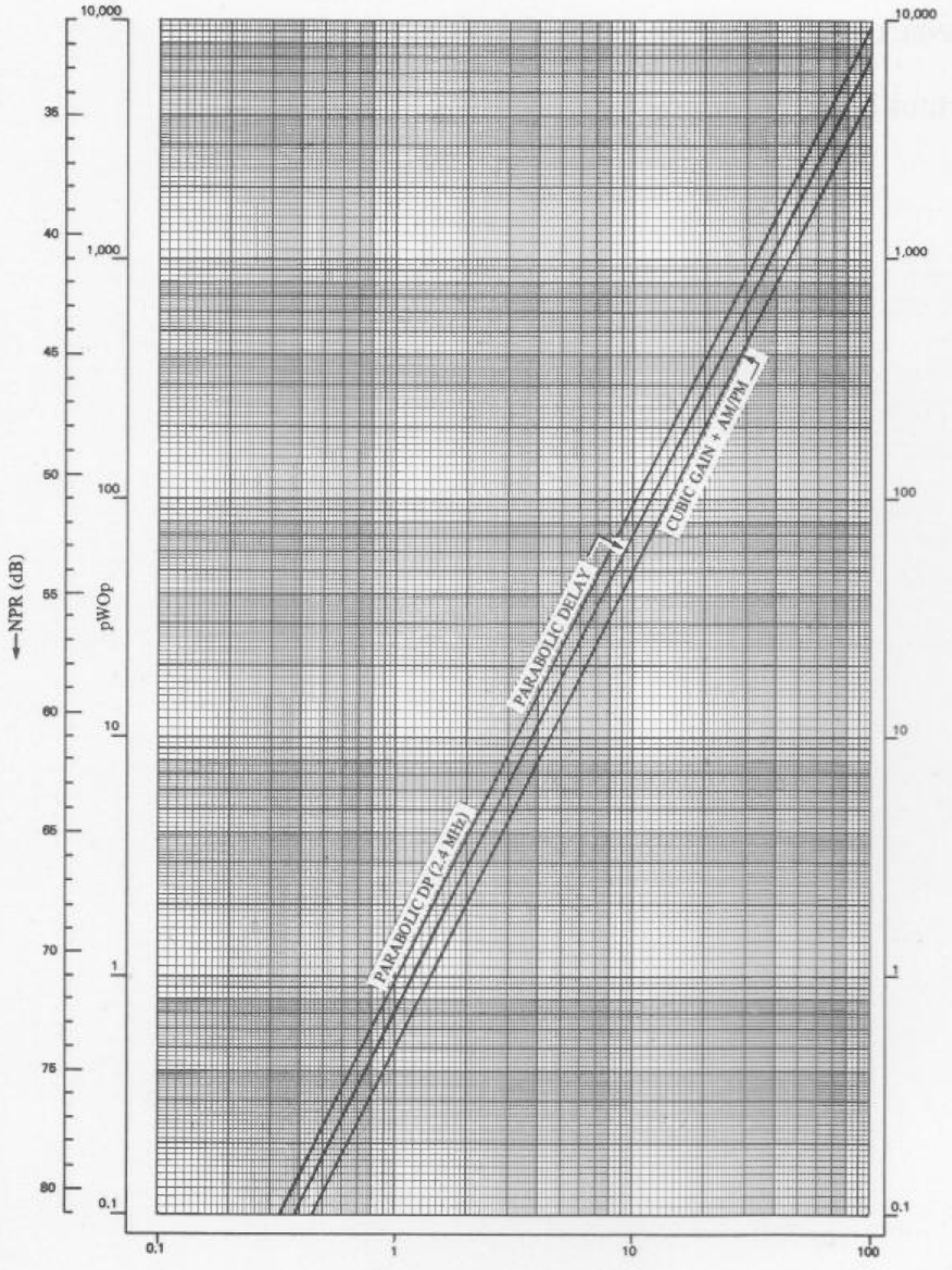


FIGURE A5-15



DP_2 (°) τ_2 (ns) [α_{3k} (dB) k(°/dB)]

GROUP DELAY DISTORTIONS (GD)

CHANNEL CAPACITY N = 1800 **TOP SLOT = 7600 kHz**

CONFIGURATION:

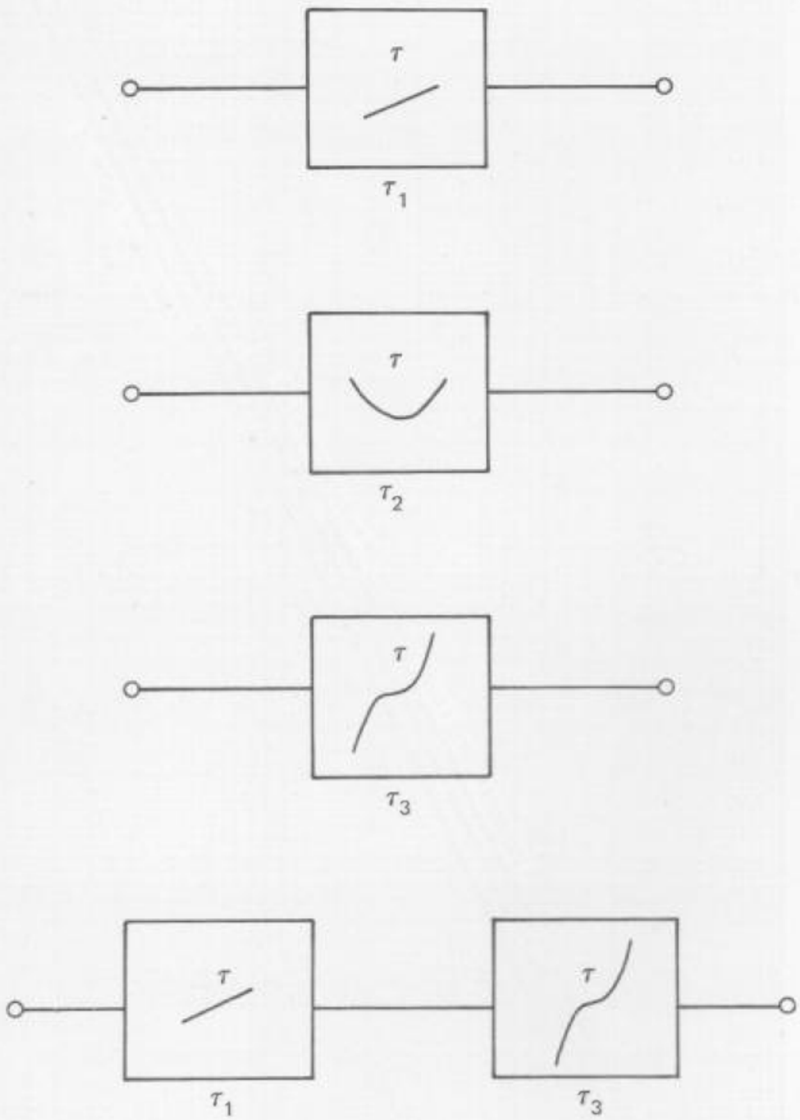
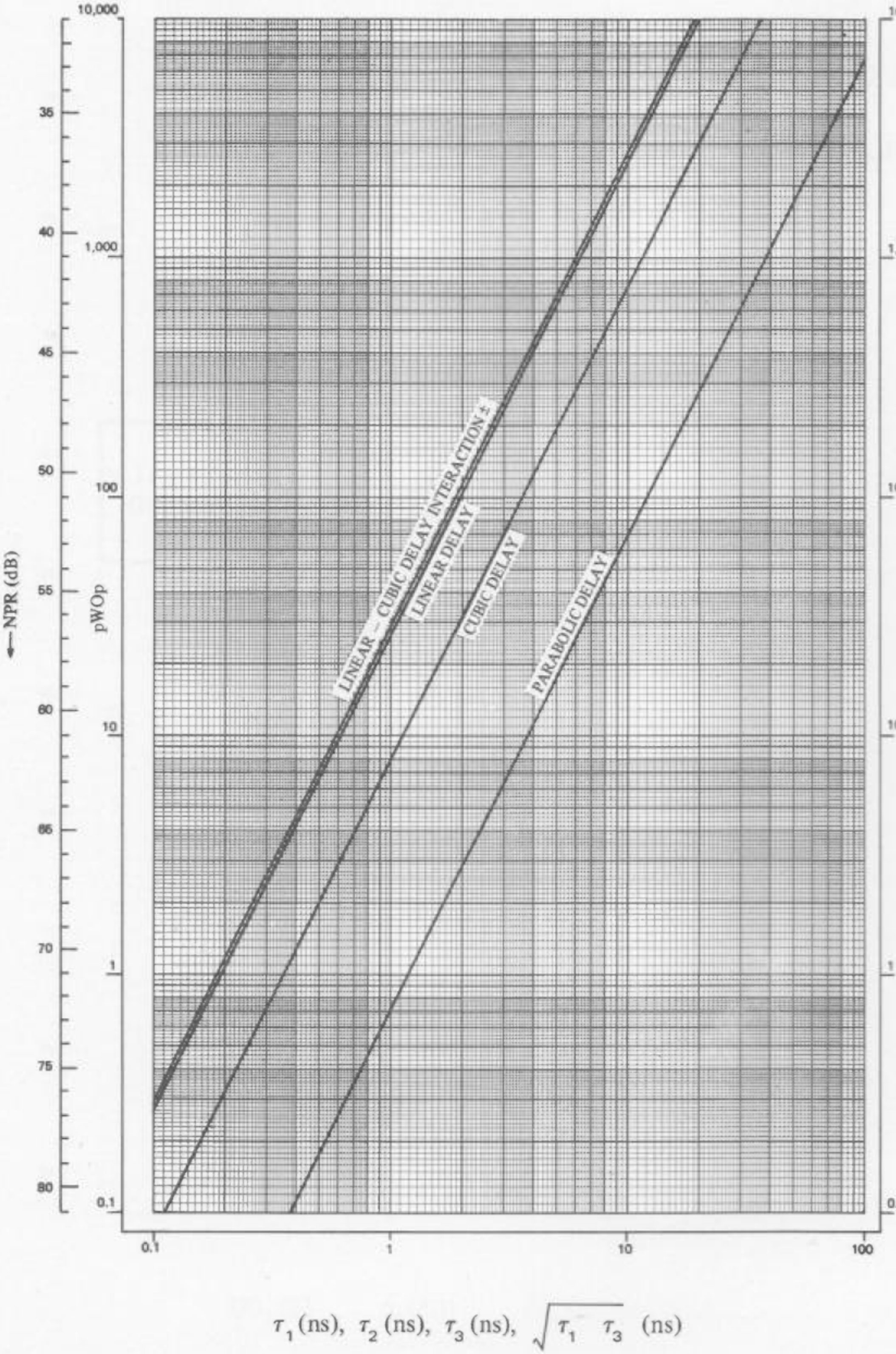


FIGURE A5-16



10,000
1,000
100
10
1
0.1
35
40
45
50
55
60
65
70
75
80
← NPR (dB)
pWOp
LINEAR – CUBIC DELAY INTERACTION ±
LINEAR DELAY
CUBIC DELAY
PARABOLIC DELAY
0.1
1
10
100
τ_1 (ns), τ_2 (ns), τ_3 (ns), $\sqrt{\tau_1 \quad \tau_3}$ (ns)

LINEAR DIFFERENTIAL GAIN (DG_1) AND ITS EQUIVALENT DISTORTIONS

CHANNEL CAPACITY N = 2700 **TOP SLOT = 11700 kHz**

CONFIGURATION:

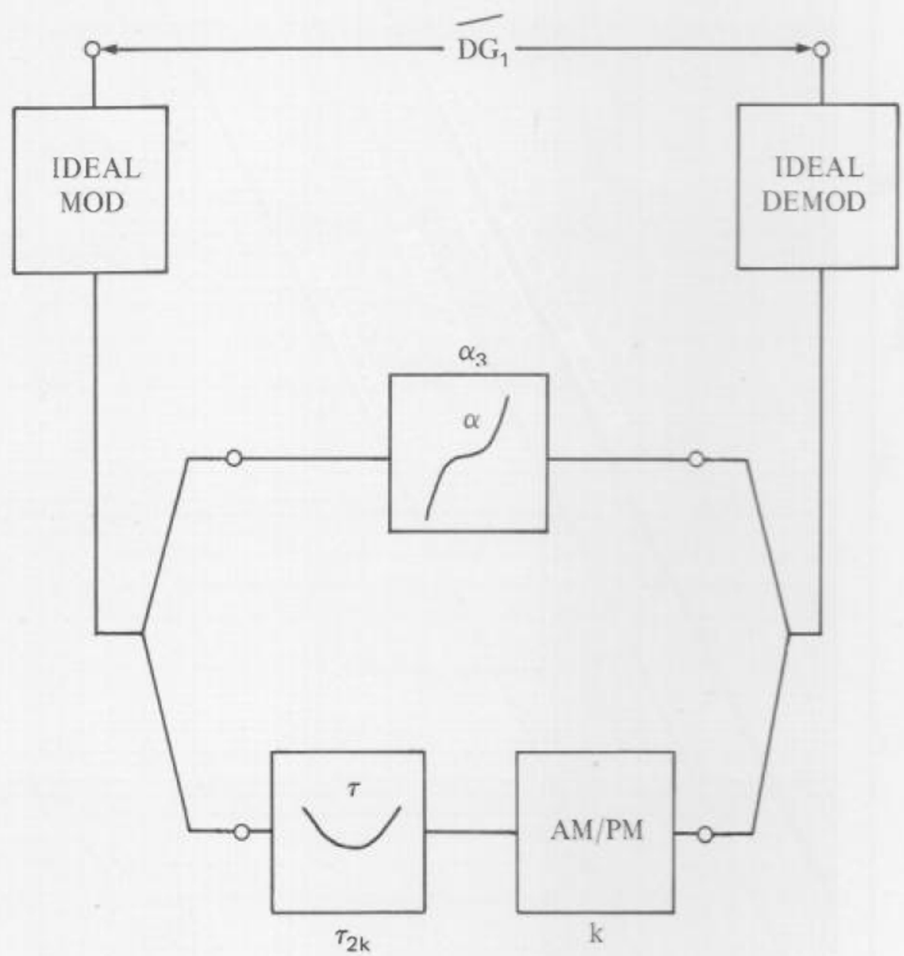
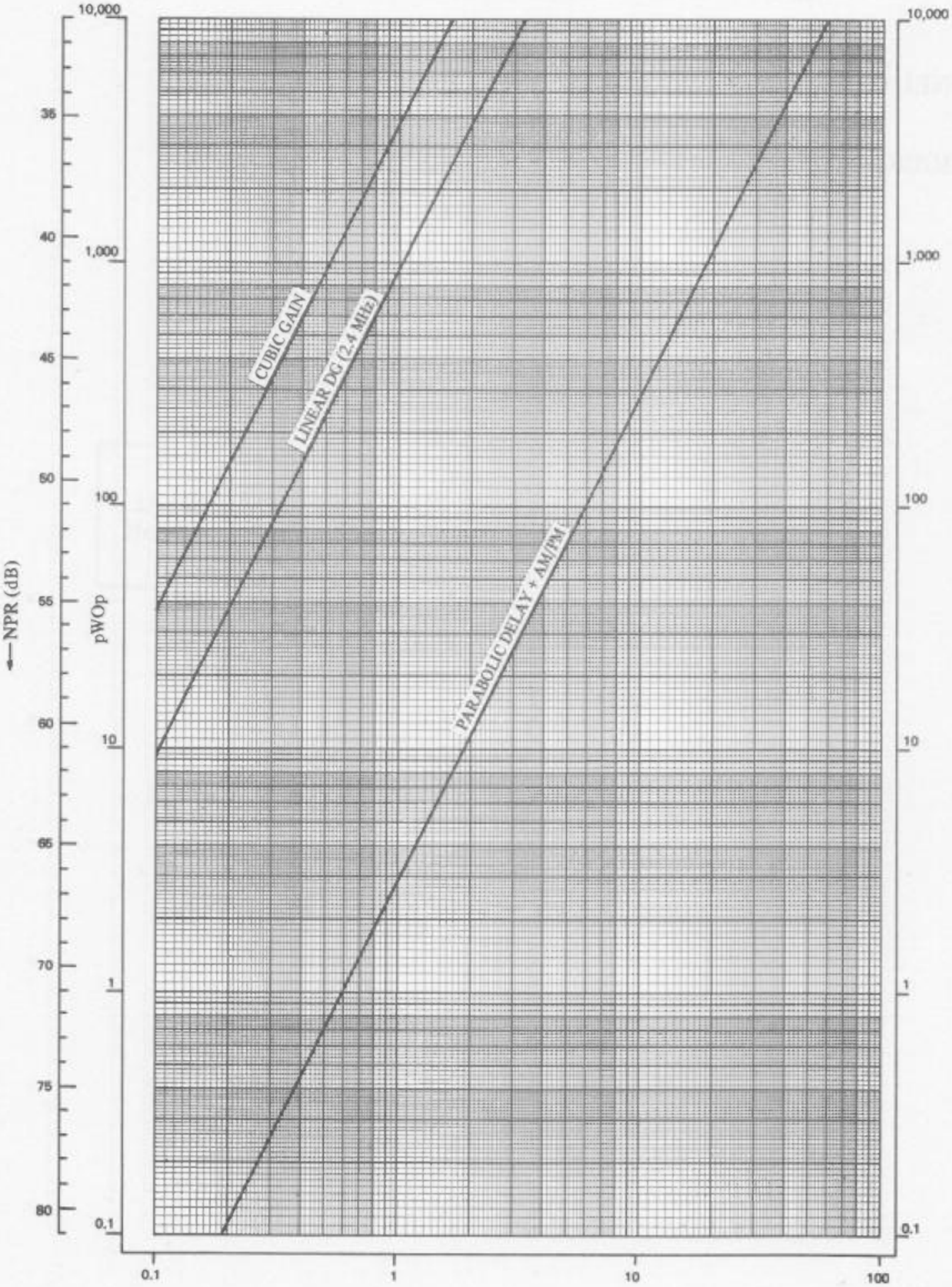


FIGURE A5-17



DG_1 (%) α_3 (dB) [τ_{2k} (ns) k(°/dB)]

PARABOLIC DIFFERENTIAL GAIN (DG_2) AND ITS EQUIVALENT DISTORTIONS

CHANNEL CAPACITY N = 2700 **TOP SLOT = 11700 kHz**

CONFIGURATION:

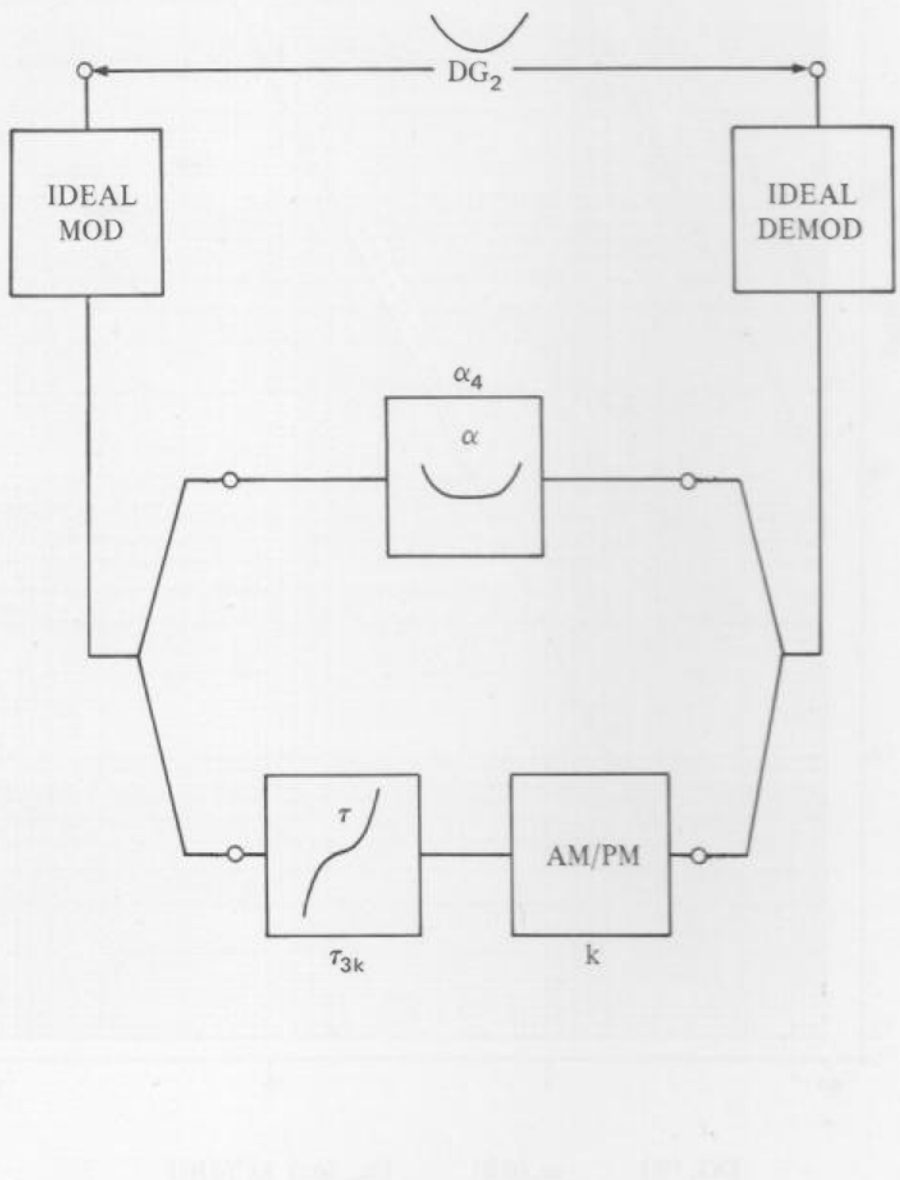
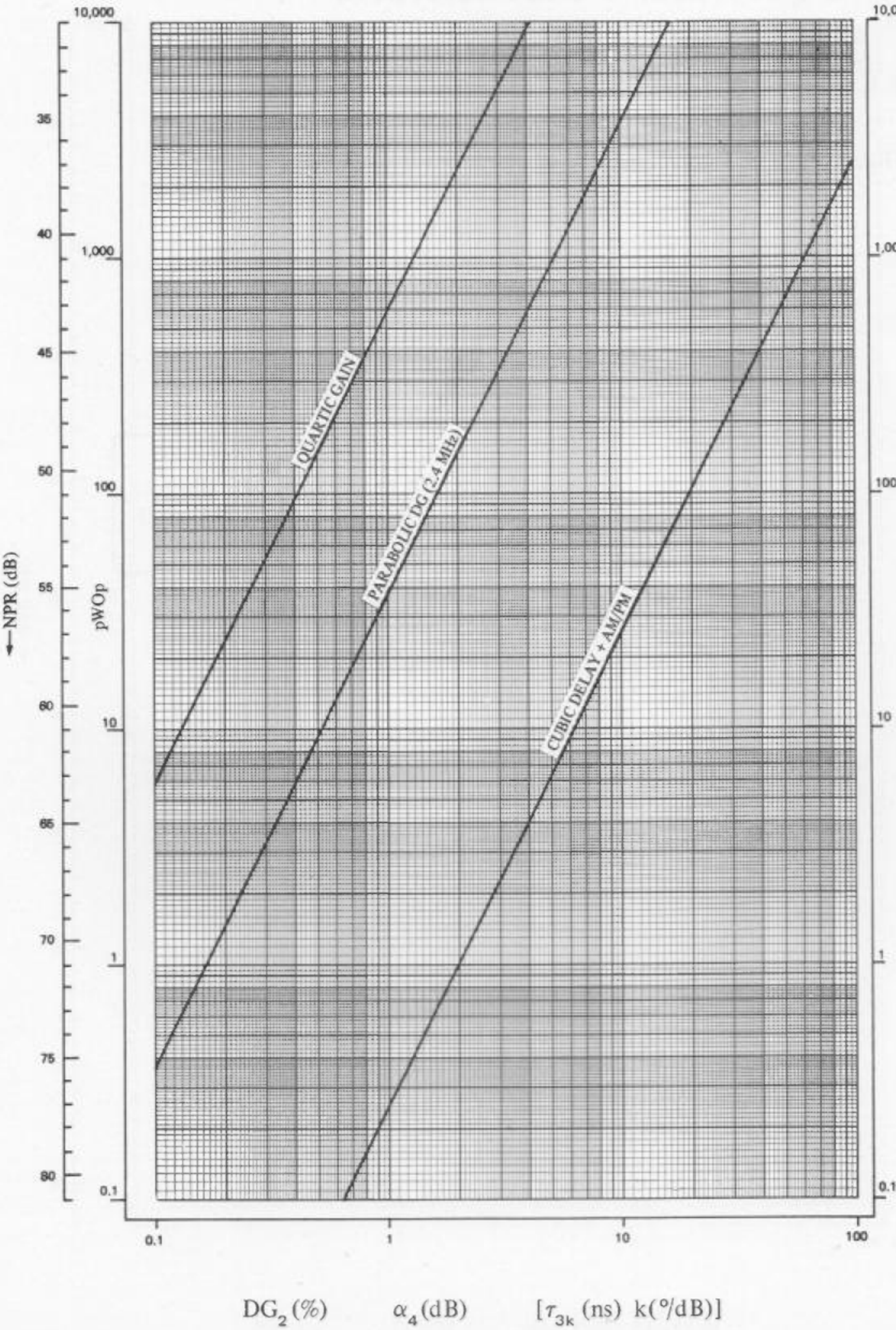


FIGURE A5-18



10,000
35
40
1,000
45
50
100
55
60
10
65
70
1
75
80
0.1
← NPR (dB)
pWOp
QUARTIC GAIN
PARABOLIC DG (2.4 MHz)
CUBIC DELAY + AM/PM
0.1
1
10
100
DG_2 (%) α_4 (dB) [τ_{3k} (ns) k(°/dB)]

LINEAR DIFFERENTIAL PHASE (DP_1) AND ITS EQUIVALENT DISTORTIONS

CHANNEL CAPACITY N = 2700 **TOP SLOT = 11700 kHz**

CONFIGURATION:

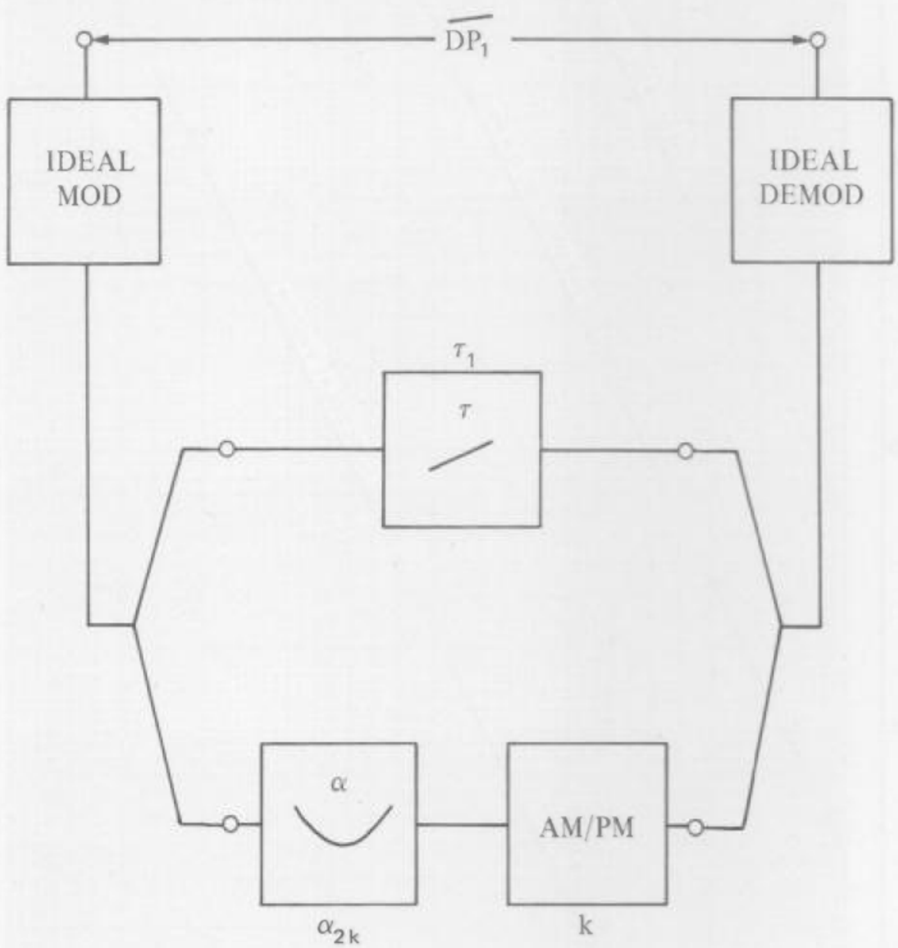
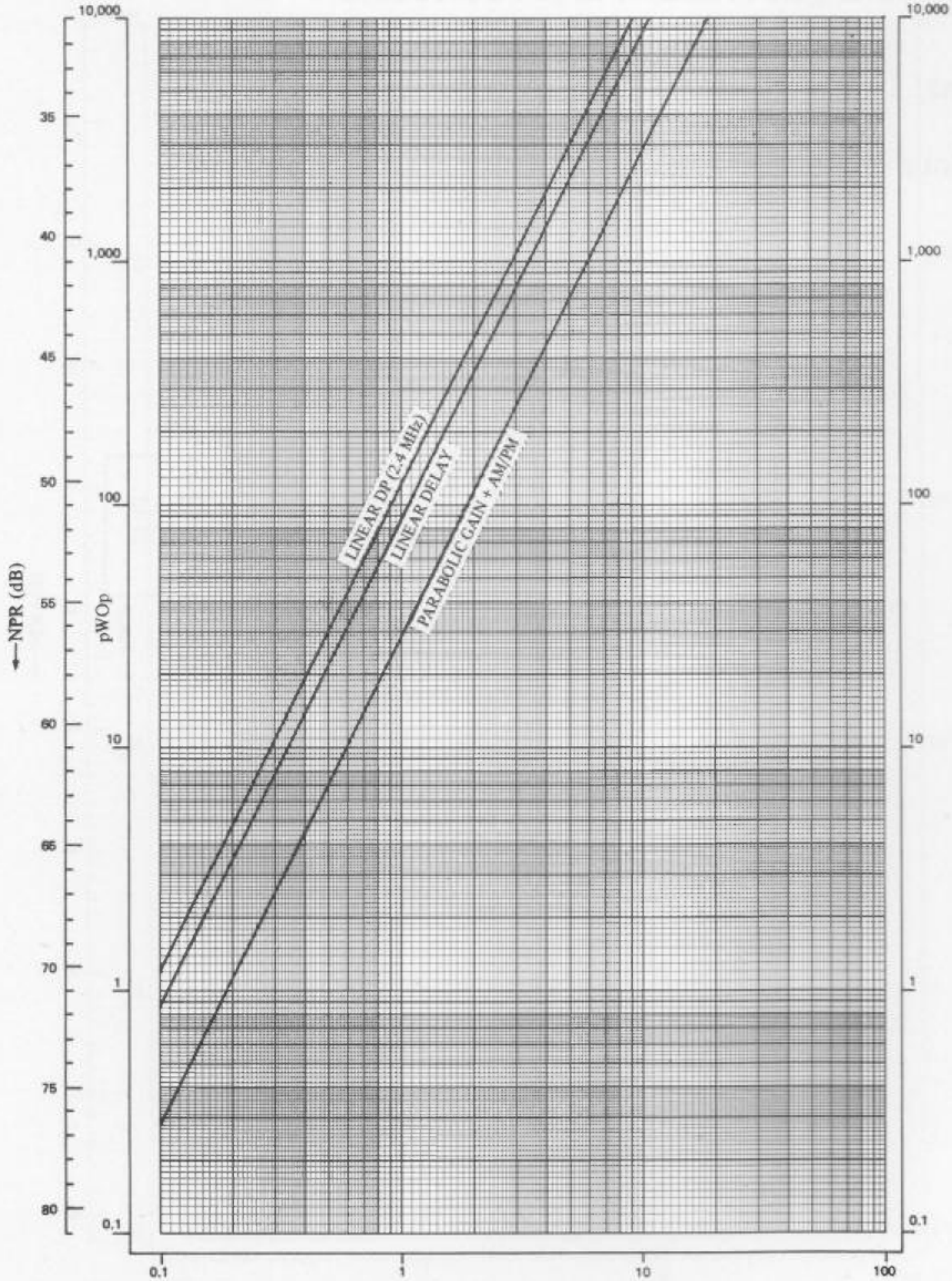


FIGURE A5-19



DP_1 (°) τ_1 (ns)   [α_{2k} (dB) k(°/dB)]

PARABOLIC DIFFERENTIAL PHASE (DP_2) AND ITS EQUIVALENT DISTORTIONS

CHANNEL CAPACITY N = 2700 TOP SLOT = 11700 kHz

CONFIGURATION:

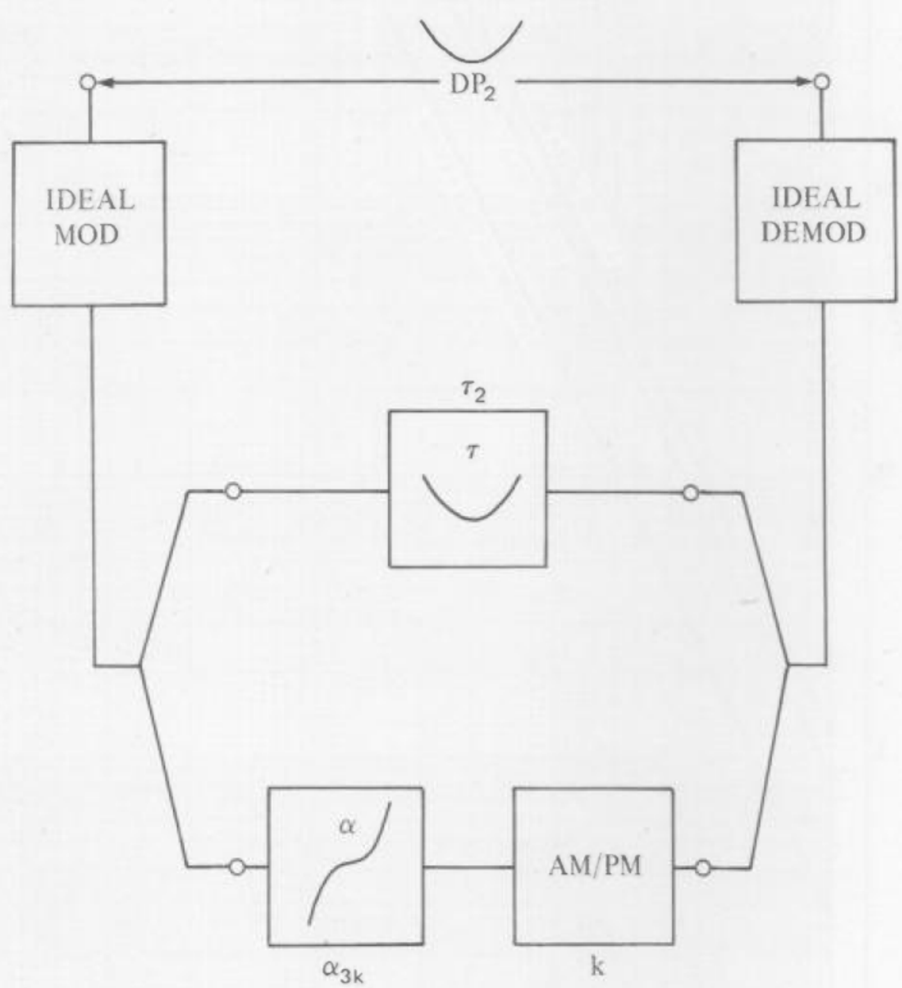
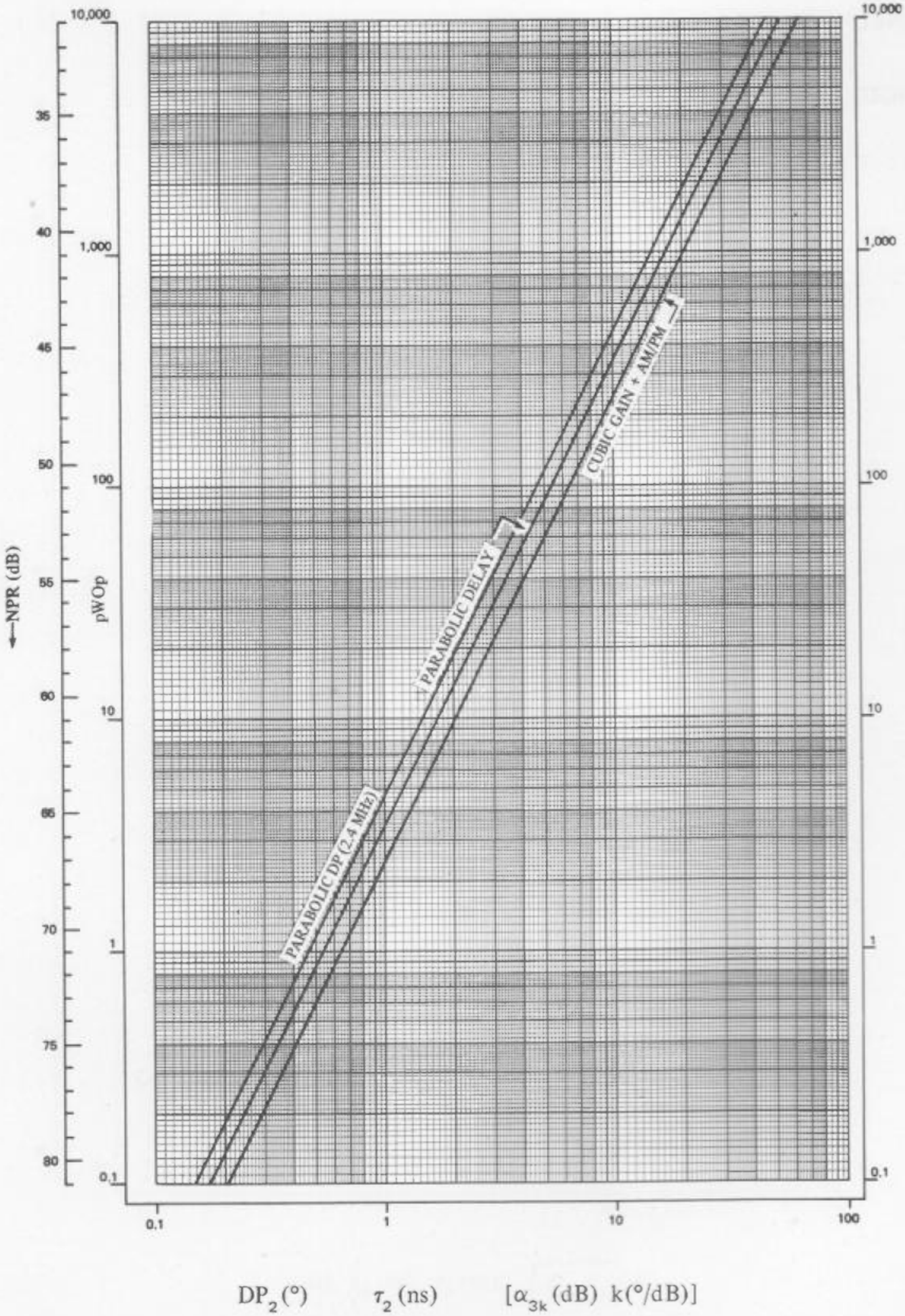


FIGURE A5-20



←NPR (dB)
35
40
45
50
55
60
65
70
75
80
pWOp
10,000
1,000
100
10
1
0.1
PARABOLIC DP (2.4 MHz)
PARABOLIC DELAY
CUBIC GAIN + AM/PM
0.1
1
10
100
DP_2 (°)
τ_2 (ns)
[α_{3k} (dB) k(°/dB)]

GROUP DELAY DISTORTIONS (GD)

CHANNEL CAPACITY N = 2700 **TOP SLOT = 11700 kHz**

CONFIGURATION:

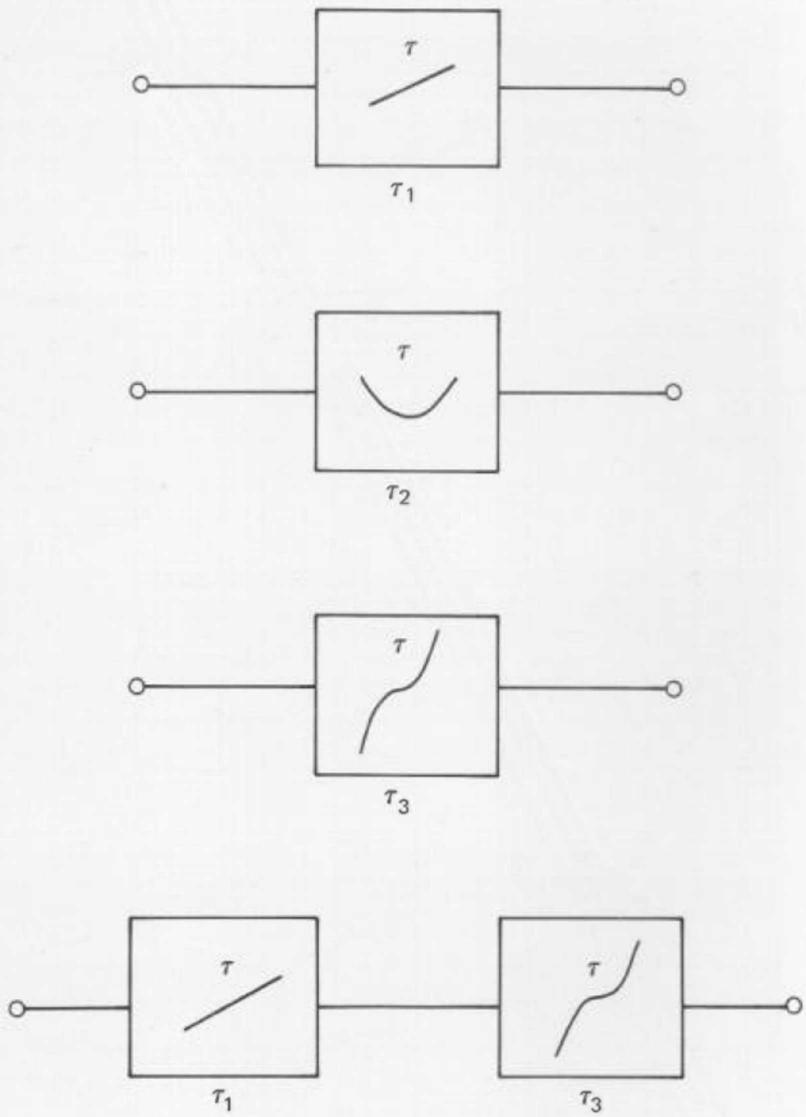
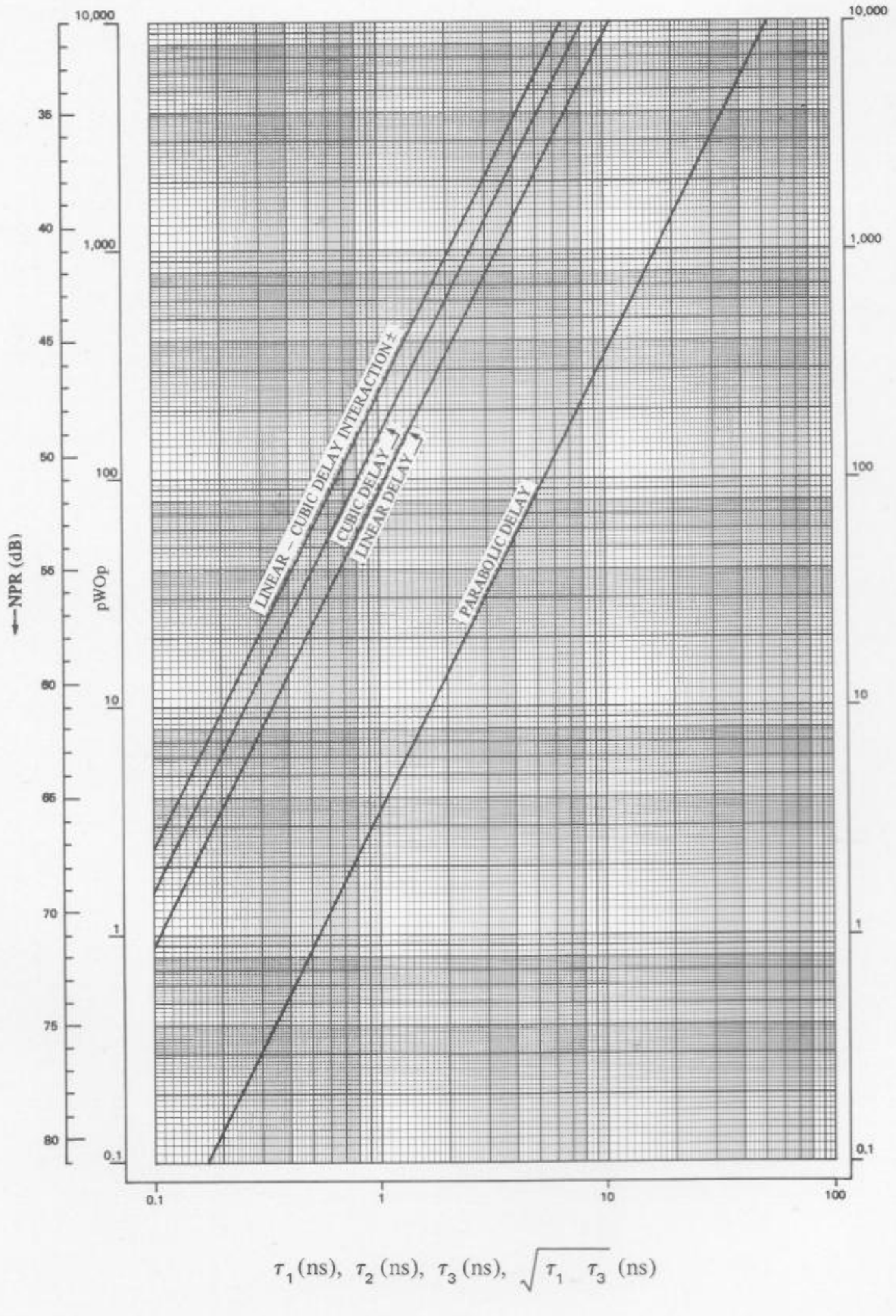


FIGURE A5-21


10,000
1,000
100
10
1
0.1
35
40
45
50
55
60
65
70
75
80
←NPR (dB)
pWOp
LINEAR – CUBIC DELAY INTERACTION ±
CUBIC DELAY
LINEAR DELAY
PARABOLIC DELAY
0.1
1
10
100
τ_1 (ns), τ_2 (ns), τ_3 (ns), $\sqrt{\tau_1 \quad \tau_3}$ (ns)

APPENDIX A6 – NOISE CONVERSION TABLE
[dBrncO, pWOp, S/N (dB), NPR (dB)]

Table A6-1 allows conversion between the various terms used for measuring noise related to a 0dBm test level point. Since these terms are sometimes derived in different ways, a brief explanation of each term is given below:

S/N (dBmO): This is the signal to noise related to a 0dBm test level point with no weighting.

$\therefore$0dBmO = 1mWO

-90dBmO = 1pWO

pWOp: This is the noise power in one voice channel given in pW with psophometric weighting. Psophometric weighting requires a 2.5dB noise addition to the unweighted level.

$\therefore$1pWOp = (-90 + 2.5) = -87.5dBmO (This relationship is used throughout the Application Note).

Note: In the US a factor of 1.5dB is often used, ie,

1pWOp = -88.5dBmO

dBrncO: This is the noise power in one voice channel given in dB above the reference noise related to zero test level with C-message weighting. C-message weighting requires a 1.5dB correction to the unweighted level.

$\therefore$0dBrncO = -90 + 1.5 = -88.5dBmO

and 1dBrncO = 1pWOp

NPR: This is the ratio of the reference noise in a voice channel to the noise remaining when the reference noise is removed from that channel, all other baseband channels being loaded with reference noise in both instances. Under standard CCIR loading conditions this yields,

NPR = (S/N - 17)dB

Table A6-1 Noise Conversion Table

dBrncO	pWOp	S/N (dB)	NPR (dB)	dBrncO	pWOp	S/N (dB)	NPR (dB)
-4	0.4	92	75	22	158	66	49
-3	0.5	91	74	23	200	65	48
-2	0.63	90	73	24	251	64	47
-1	0.79	89	72	25	316	63	46
0	1.0	88	71	26	398	62	45
1	1.3	87	70	27	501	61	44
2	1.6	86	69	28	631	60	43
3	2.0	85	68	29	794	59	42
4	2.5	84	67	30	1,000	58	41
5	3.2	83	66	31	1,259	57	40
6	4.0	82	65	32	1,585	56	39
7	5.0	81	64	33	1,995	55	38
8	6.3	80	63	34	2,512	54	37
9	7.9	79	62	35	3,162	53	36
10	10.0	78	61	36	3,981	52	35
11	12.6	77	60	37	5,012	51	34
12	15.8	76	59	38	6,310	50	33
13	20.0	75	58	39	7,943	49	32
14	25.1	74	57	40	10,000	48	31
15	31.6	73	56	41	12,589	47	30
16	39.8	72	55	42	15,850	46	29
17	50.1	71	54	43	19,953	45	28
18	63.1	70	53	44	25,120	44	27
19	79.4	69	52	45	31,620	43	26
20	100	68	51	46	39,810	42	25
21	126	67	50	47	50,120	41	24

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