NATIONAL RADIO ASTRONOMY OBSERVATORY

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EQUALIZATION AND SIGNAL-TO-NOISE PROBLEMS IN THE IF TRANSMISSION SYSTEM

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I. Introduction

Some aspects of the design of the VLA IF system have been covered in two previous studies: VLA Electronics Memorandum No. 5, Modulation Systems for Transmission Through Dispersive Cables, and No. 6, Prototype IF Transmission System, by S. Weinreb. In the present memorandum the advantages of double sideband modulation for the IF system are discussed. A block diagram of the proposed DSB system is presented and the functions of the various components are described. Next a calculation is made of the signal-to-noise ratio as a function of the number of repeaters in a 21km link. In addition, the results of a computer computation of the intermodulation noise spectrum are presented. These are then used to derive the signal-to-noise ratio in each individual channel, for both pre-equalization and post-equalization of the transmission cables. Consideration is given to the effects of the non-linear phase shift of the cables (the treatment is similar to the one presented in VLA Electronics Memorandum No. 5). We also suggest a system for measuring the non-linear phase of the cables and amplifiers. Finally, the design of the cable equalizers is discussed in detail. Results are presented which include computer design curves and measurements on several prototype equalizers.

II. A Double Sideband IF System

The IF transmission link basically consists of an IF transmitter, an IF receiver, a cable system and several repeaters which connect the telescopes to the central control building. A block diagram of the principle components of the system is shown in Figure 1. The number of repeaters in the link is determined by signal-to-noise considerations. The number of telescopes on each IF link is determined by the type of modulation used and the bandwidth of the repeaters. An octave bandwidth is chosen since it eliminates square intermodulation components ($\approx S^2$) from the spectrum. If we consider a system operating from 1 to to 2 Gc with a signal bandwidth of 50Mc, we can have 6 channels of DSB or 12 channels of SSB (Allowing 40% of the spectrum for guard bands). For the array configuration presented in the VLA proposal a DSB system will have 3 telescopes on each cable, hence requiring 4 cables of lengths: 21km, 14.7km, 8.4km, and 2.km, for a total of 46.2km. Alternatively, a SSB system would have 6 telescopes per cable and require 2 cables whose total length is 29.4km. Hence the DSB system requires about 1.5 times the amount of cable and repeaters. However, there are a number of advantages to the DSB type of modulation which make it worth considering:

1. A 3 db increase in signal-to-noise ratio. (This is due to the fact that when the two sidebands are added the signal adds coherently while the noise adds incoherently.)
2. Cancellation of linear variations in attentuation ($\approx\omega$) and square variations in phase ($\approx\omega^2$) which are introduced in the process of transmission.(See VLA Electronics Memorandum No. 5).
3. Reduced cross talk between the two channels originating from an individual antenna. (In the SSB system these are located on the upper and lower sidebands of the same carrier).
4. Simpler transmitter and receiver design. (A detailed description of an SSB transmission system is given in VLA Electronics Memorandum No. 6).

In Figures 2 - 4 we show more detailed block diagrams of the IF system components. The repeater (Fig. 2) must be capable of giving a gain exactly equal to the cable loss, as well as equalizing the attenuation variation and non-linear phase shift produced by the cable. The equalizers are discussed in Section VII. The attenuator is necessary in order to set the level of the repeater output for optimum signal-to-noise ratio. The amplifiers have a gain of 25 db in the 1- 2 Gc band, a noise figure of +6 db and a tone intercept level of +20 dbm (Avantek AMP-1000). The total gain of the repeater will be less than 50 db as there is some loss in the equalizers (3 - 6 db).

The IF transmitter is shown in Figure 3. The local oscillator can be a standard signal source and does not need to be phase locked. The modulator must have a 50Mc bandwidth and operate between 1 and 2 Gc. It may be possible to use simply a mixer and a directional coupler. The output of the IF transmitter is fed into a repeater mounted on the telescope to produce the proper level and equalization required to couple into the IF link at that point.

The IF receiver (Fig. 4) is shown in two different versions. An AGC amplifier sets the input level and a trim equalizer corrects for the amplitude and phase variations produced by the repeaters in the link. A second amplifier and power splitter may be required to distribute the signals to the various channel filters. The channel filters are 100Mc wide and pass the carrier plus sidebands which are then video detected. The second 50 Mc low pass filter is used to reject any unwanted mixer products. An alternative scheme is also shown which uses a phase locked oscillator and mixer in place of the filter and video detector. This may provide some improvement in signal-to-noise ratio. Finally, the output of the IF receiver must be fed into a 50Mc modulator (possibly followed by a 50-100Mc filter) since the delay lines require the signal band to be located between 50 and 100Mc.

III. Signal-to-Noise Ratio for an Octave Bandwidth System

Most of the considerations involved in the calculation of the signal-to-noise ratio are given in the ITT Labs report. To obtain the S/N variation as a function of the number of repeaters we define the following quantities:

S_o = total signal power at repeater output (dbm)
N_T = total thermal noise power referred to repeater input (dbm)
N_{IM} = total intermodulation noise power referred to repeater input (dbm)
L = total loss in the cable (21km) at $2f_o$
R = number of repeaters
G = L/R = repeater gain
X = tone intercept level of repeater output amplifier (dbm) (1)
f_o = lowest operating frequency
$2f_o$ = highest operating frequency

In determining the noise level at the end of the IF transmission it is important to realize that the thermal noise power of each individual repeater adds an equal amount to the total noise, since the gain of each stage (repeater + cable) is one. Hence the noise at the end of the link is 10 log R greater than at the beginning. The intermodulation noise adds coherently if each amplifier has identical non-linearities and thus the IM noise is greater by 20 log R. The IM noise is assumed to be caused by cubic distortion ($\sim S^3$) in the amplifier. For the case of two pure tones the signal to IM noise is just twice the difference between the tone intercept level and the signal output level. This gives the following signal-to-noise ratios:

$$\left(S/N_T\right) = S_o - N_T - L/R - 10 \log R$$
$$\left(S/N_{IM}\right) = 2(X - S_o) - 20 \log R \tag{2}$$

It can be shown that the maximum S/N occurs when the thermal noise is twice the IM noise. Hence, one can derive the following relations:

$$\left(S/N_T\right) = \frac{2}{3}\left(-N_T - 20 \log R - L/R + X - 1.5\right)$$
$$\left(S/N_{IM}\right) = \left(S/N_T\right) + 3 \tag{3}$$
$$S_o = \frac{1}{3}\left(N_T - 10 \log R + L/R + 2X - 3\right)$$

The S/N as a function of the number of repeaters is plotted in Fig. 5. This is for a 21km link and thus represents a worst case value of S/N . The parameter on the curves is the cable loss at the highest operating frequency. Typical 1 - 5/8" cable losses at 2Gc vary from 33 db/km to 44 db/km. In addition, we have plotted the output signal level which is necessary to obtain these optimum S/N in Fig. 6. In this case we are considering the total signal and noise power in the octave bandwidth system. In the next section we will discuss the detailed spectrum of the signal and noise.

IV. Intermodulation Noise in an Octave Bandwidth System

In order to calculate the IM noise spectrum we assume a discrete input spectrum to the final amplifier in the repeater which consists of 100 pure tones equally spaced across the octave bandwidth. The spectrum is assumed to be either flat (post-equalization) or increasing in db as $f^{1/2}$ (pre-equalization).

$$S = \sum_{n=1}^{100} A_n \cos\left[\left(1 + \frac{n}{100}\right)\omega_o t + \phi_n\right] \tag{4}$$

The output of the amplifier contains only cubic and higher order distortion components in the octave band and the spectrum of these can be obtained by cubing the input time function:

$$S_o = S_i + kS_i^3 \tag{5}$$

$$S_i^3 = \left(\sum_{\ell=1}^{100} A_\ell \cos \left[(1 + \frac{\ell}{100})\ \omega_o t + \phi_\ell \right] \right) \left(\sum_{m=1}^{100} \text{same} \right) \left(\sum_{n=1}^{100} \text{same} \right) \tag{6}$$

which can be expanded as:

$$S_i^3 = \sum_{\ell=1}^{100} \sum_{m=1}^{100} \sum_{n=1}^{100} A_\ell\ A_m\ A_n \left(\cos \left[(1 + \frac{\ell}{100})\ \omega_o t + \phi_\ell \right] \right) \left(\text{Same} \right) \left(\text{Same} \right) \tag{7}$$

Finally, the frequencies produced can be obtained from the multiple product as:

$$S_i^3 = \sum_{\ell=1}^{100} \sum_{m=1}^{100} \sum_{n=1}^{100} \frac{A_\ell A_m A_n}{4} \cos \left\{ \left(\begin{bmatrix} |L+M+N| \\ |L+M-N| \\ |L-M+N| \\ |L-M-N| \end{bmatrix} \omega_o t + \phi_{\ell mn} \right) \right\} \tag{8}$$

where;

$$L = 1 + \frac{\ell}{100} \qquad M = 1 + \frac{m}{100} \qquad N = 1 + \frac{n}{100}$$

The computer then carries out the triple summation, adding up the amplitudes at each frequency. The amplitudes are added in the RMS sense since the phase of the cubic components is random. The spectrum of the IM noise obtained in this way is plotted in Fig. 7. The absolute level of the noise in this plot is arbitrary.

To determine the effect of this spectrum on the S/N calculated earlier, we must evaluate the constant k in Eq. 5. This involves considering in detail the definition of tone intercept level and applying it to the individual tones we have used as the input spectrum. For our case we define the intermodulation noise as the amplitude of the cubic component at the same frequency as the signal. Hence we obtain a S/N at each of the 100 frequencies in the octave band. This can then be applied to any number of channels which one chooses to split the band into.

An additional point needs to be considered in determining the S/N, and that is the fact that the thermal noise is generated at the input to the repeater while the IM noise is generated at the output. Since there is an equalizer in the middle of the repeater, the output thermal noise spectrum goes as $f^{1/2}$. Finally, to calculate the S/N across the band, we have added the thermal noise spectrum to the IM noise spectrum. In addition, the signal spectrum is either flat (post-equalization) or tapered as $f^{1/2}$ (pre-equalization). Figure 8 displays the pre-equalized signal spectrum for n channels. The signal and noise spectrums are then used to obtain the curves shown in Figs. 9 and 10.

These curves represent the departure of the S/N from that calculated in the previous section (Fig. 5). Note that optimizing the S/N is no longer as straightforward, since each channel maximizes at a different output signal level. It should be pointed out that the S/N plotted in Fig. 5 is only the ratio of signal power to <u>thermal</u> noise power. Adding in the IM noise calculated in Section III would reduce the S/N by 1.8 db. However, the detailed consideration of the IM noise spectrum shows that the output signal level can be raised about 2 db, and hence Fig. 5 gives approximately the correct average ratio of signal power to <u>total</u> noise power. In order to determine the individual channel S/N, the values in Fig. 9 or Fig. 10 are simply added to those in Fig. 5. In addition, the required increase in signal output power should be added to Fig. 6 to obtain the total signal output power. In the case of post-equalization, the signal power in each channel can then be determined by adding -10 log n to the total signal output power. For pre-equalization, the total signal output power is added to the values in Fig. 8.

In conclusion, we feel that the S/N calculation done in Sections III and IV is rather conservative. For one, all the signals are assumed to be present for the entire 21 km of the link. In addition, the IM noise is assumed to add coherently. If it adds incoherently, like the thermal noise, the S/N is improved by 4 to 5 db. In addition, a DSB system will add 3 db to the calculated S/N. Finally, the signal is not a smooth spectrum as shown in Fig. 8, but instead consists of 100 Mc clumps separated by 50 Mc gaps, which means that only 0.6 of the thermal noise is actually relevant. All of this could add up to as much as a 10 db improvement in the S/N. It is hoped that some of these uncertainties can be resolved by measurements on a prototype link.

V. <u>Phase Non-Linearity in the IF System</u>

There are two main sources of amplitude variation and phase non-linearity in the IF system. These are due to frequency variations of the repeater amplifiers and the dispersion in the coaxial cables. According to the minimum phase criterion: if all the networks in the system are minimum phase, then the amplitude and phase responses are directly related, and therefore, it is sufficient to equalize the amplitude response and the phase will be linearized at the same time. The non-linear phase of a coaxial cable can be expressed as:

$$\begin{aligned} (e^{\alpha})\ (e^{j\phi}) &= e^{-(1+j)af^{1/2}} \\ \alpha &= -\ af^{1/2} \\ \phi &= -\ af^{1/2} \end{aligned} \tag{9}$$

This variation in phase has been verified by some measurements made at Green Bank. It leads to a variation of phase which can be expressed as 6.6°/db of attenuation. The cable is a minimum phase network, and it will be shown in Section VII that the eqalizers which are designed to produce a flat amplitude response also yield a linear phase response. In addition, it can be shown from the following that the effect of the non-linear phase of the cable on the individual channels is negligible.

We expand the phase in a Taylor series about a carrier frequency ω_o:

$$\phi(\omega_o+\Delta\omega) = \phi(\omega_o) + \frac{d\phi}{d\omega}\bigg|_{\omega_o} \Delta\omega + \frac{1}{2}\frac{d^2\phi}{d\omega^2}\bigg|_{\omega_o} (\Delta\omega)^2 + \frac{1}{6}\frac{d^3\phi}{d\omega^3}\bigg|_{\omega_o} (\Delta\omega)^3 \ldots\ldots \tag{10}$$

For SSB transmission the $(\Delta\omega)^2$ term produces departures from linear phase which are cancelled out in DSB. Hence, the non-linear phase terms for 21 km of 1-5/8" coax with a loss at 2 Gc of 33 db/km are:

$$\begin{aligned} &\text{SSB} \quad \phi = \frac{1}{2}\,\phi''\ (\Delta\omega)^2 = 1.03° && \Delta\omega = 50\ \text{Mc} \\ &\text{DSB} \quad \phi = \frac{1}{6}\,\phi'''\,(\Delta\omega)^3 = 0.026° && \omega_o = 1\ \text{Gc} \end{aligned} \tag{11}$$

These are well below the design goal of 10° for the IF system. In addition, the equalizers will reduce these phase variations to negligible effects. However, the phase variations in the amplifiers may present a problem which would require more than just a trimming equalizer in the IF receiver. Also, recent measurements have indicated an anomalous dispersion above 600 Mc in 1-5/8" coaxial cable. The degree of this difficulty will have to be determined in amplifier and prototype system tests.

VI. Measurement of the Non-Linear Phase

In the previous section it was shown that the effect of the non-linear phase of the cable on the individual IF channels is very small. However, the effect of other sources of non-linear phase, such as the repeater amplifiers, may be significant. In order to measure the non-linear phase on the IF link, it is necessary to have a system which does not require the two ends of the link to be brought together. This can be done with the system shown in Fig. 11. Since two signals are sent down the line, the phase shift can be measured between them without the need for an external reference. Essentially the system measures the change in phase shift of 50 Mc information as it is swept from 1 to 2 GC. This is equivalent to measuring the change in group delay or slope of the phase curve ($\tau = d\phi/d\omega$).

The measurement consists of sending a 50 Mc DSB modulated signal on a 1 Gc carrier and a similar signal on a carrier which is swept from approximately 1.2 to 2 Gc. At the other end both carriers are detected and the phase shift of their modulation compared. The sensitivity of the system is determined by how accurately the phase can be detected. The output of the detectors will produce 18° of phase shift for a 1 ns change in group delay. For the 1-5/8" coax considered in the last section, the variation in group delay for 21 km from 1 to 2 Gc is 1.3 ns. This will produce 24° of phase shift, whereas the actual non-linear phase shift was shown to be 1.03° for SSB and .026° for DSB. Hence, it should be easy to measure any dispersion which would be significant in terms of the individual IF channels. Note that a measurement of the change in group delay is the same as measuring the SSB non-linear phase shift:

$$\Delta\phi_{meas} = \left(\left.\frac{d\phi}{d\omega}\right|_{\omega_2} - \left.\frac{d\phi}{d\omega}\right|_{\omega_1} \right) \Delta\omega \tag{12}$$

Using the definition of the second derivative and Eq. 11

$$\Delta\phi_{meas} = \frac{d^2\phi}{d\omega^2} \quad (\omega_2-\omega_1) \quad \Delta\omega = 2\frac{(\omega_2 - \omega_1)}{\Delta\omega} \quad \phi_{SSB} \tag{13}$$

Thus we have a direct way of making a swept frequency measurement of the non-linear phase shift without needing a separate reference cable.

VII. Design of the Cable Equalizers

The main source of attenuation in coaxial cables is the skin depth resistance in the conductors which produces an attenuation in db proportional to $f^{1/2}$. In addition, a small amount of dielectric loss will be present, giving a loss proportional to f. The skin depth loss, which is shown in Fig. 12, is very nearly a linear slope of approximately 10 db/octave in the interval f_o to $2f_o$. The curvature or departure from this slope is a maximum of 0.4 db less attenuation at a frequency of 1.43 f_o. The slope and curvature for various amounts of cable loss are shown in Table I. It can be seen that the curvature is about 4% of the slope. Hence, the major equalization problem will be to correct for the 10 db/octave slope across the band. In the equalizer design, we have chosen to equalize the slope first and then try to trim up the curvature so that the spectrum will be flat to a few tenths of a db for each repeater section. This will give a total attenuation variation for the 21 km link of a few db which can be removed by the trim equalizer in the IF receiver.

The slope equalizer is based on the simple, constant input impedance network which is shown in Fig. 13. The reactances act to divide the input power between the two resistors. Hence, one can arrange the reactances so that most of the high frequency energy goes to one resistor while most of the low frequency energy goes to the other resistor. For the slope equalizer, transmission lines are used for the reactive elements instead of inductors and capacitors. Transmission lines are chosen because they are easier to construct than lumped elements at frequencies above a few hundred megacycles. They also have a more rapid reactance variation which will allow steeper slopes to be obtained from the equalizer.

The transmission line form of the equalizer is shown in Fig. 14. The line lengths are chosen to be 1/4 wavelength at about 2.6 f_o. At this frequency the shorted stub is an open circuit while the open stub is shorting out the 50Ω resistor, so that the equalizer theoretically has no loss. As the frequency is reduced, more and more of the input power is dissipated in the 50Ω resistor. However, at all frequencies the input impedance remains 50Ω. The frequency response of the equalizer for various values of characteristic impedance of the shorted stub is shown in Fig. 15 . In Table I we have indicated the characteristic impedance which gives the proper equalization for the cable lengths listed.

TABLE I

SLOPE AND CURVATURE FOR VARIOUS CABLE ATTENUATION

Loss at 2 f_o	Loss at f_o	Slope db/octave	Curvature db	Length 1-5/8" f_o=1Gc	Equalizer Z_o
5 db	3.5 db	1.5	.06	.14km	110Ω
10 db	7.1 db	2.9	.13	.29km	72Ω
15 db	10.6 db	4.4	.19	.43km	50Ω
20 db	14.1 db	5.9	.25	.57km	37Ω
25 db	17.7 db	7.3	.32	.71km	28Ω
30 db	21.2 db	8.8	.38	.86km	18Ω
35 db	24.8 db	10.2	.44	1.00km	10Ω
40 db	28.3 db	11.7	.51	1.14km	
45 db	31.8 db	13.2	.57	1.29km	
50 db	35.3 db	14.7	.63	1.43km	

When a prototype equalizer was tested, the open stub was found to be very sensitive to stray capacitance,since it is a balanced line. As a result, we have modified the design by replacing this line with a capacitor. The input impedance is no longer constant; however, the maximum VSWR is only 1.10. The high frequency attenuation is also increased slightly. A plot of the frequency response of a 24Ω equalizer with an open stub and with a capacitor is shown in Fig. 16. In addition, we have also plotted the measurements which were made on an experimental model of the capacitor equalizer. It appears to be within 0.25 db of the theoretical response except at the high end. The slope and high frequency loss can be changed by varying the value of capacitance.

The total response of the equalizer plus cable can be seen in Fig. 17 to be quite flat. In addition, the phase shift is very linear across the band. This is a result of the minimum phase properties of the equalizer and cable. Note that the curvature is rather small, indicating that the slope equalizer has compensated for some of the curvature. The amount of residual curvature for this type of equalizer is usually less than 0.5db, and for this particular case is 0.1 db. However, some additional curvature may be added by the dielectric losses in the cable.

In order to compensate for the slight roll-off at the band edges, we follow the slope equalizer wihh a transmission line curvature equalizer (Fig. 18). This equalizer consists of an open circuited stub ($\lambda/2$ at f_o), which is lightly coupled to the main line. The transmission stub is an open circuit at f_o and $2f_o$, and thus the equalizer has no attenuation at these frequencies. At 1.5 f_o the stub is a short circuit and provides maximum attenuation which can be varied by adjusting the amount of coupling. The frequency response of the curvature equalizer is a cosine function with maximums at f_o and $2f_o$ (Fig. 18). To compensate for 0.5 db of curvature requires a coupling of -30db (-15db in each direction through the coupler). Measurements presently being made on an experimental model of the curvature equalizer indicate the configuration of the coupler has some influence on the frequency response. Preliminary results are plotted in Fig. 18.

The function of the slope and curvature equalizers is to reduce the gain variation in each repeater section of the IF link to a few tenths of a db. The equalizers will mainly be used to compensate for the cable dispersion and attenuation variations; however, some equalization of the repeater amplifiers may also be possible. Assuming the buildup of gain variation is somewhat correlated as the signal proceeds down the IF link, one might expect about ten times the individual section variations to apply for the entire link. Hence, a variation of a few db must be expected at the IF receiver.

To equalize these variations, a trim equalizer is being designed (Fig. 19). This is a time delay equalizer which adds to the main signal, a signal with various delays inserted. The frequency response of this equalizer is similar to the curvature equalizer. Each delay section produces a cosine variation in the frequency spectrum. The rate of variation of this cosine function is controlled by the amount of delay and the amplitude by the amount of coupling. Hence, by having a sufficient number of delays equally spaced, a Fourier cosine series can be built up and any frequency response can be synthesized (Fig. 19).

At present it is difficult to say how many terms or delays will be needed. Ten is certainly an upper limit, and probably four or five would be sufficient. Experiments on the trim equalizer have begun, and it appears to be quite feasible. Details of the coupling to the main line still have to be worked out. An alignment procedure will have to be studied since one is synthesizing the inverse of the IF link response with Fourier components. This is not as bad as it sounds, since a sweep generator will be located at the far end of the link, and the trim equalizer, which will have about five adjustments, can very easily be tuned to give minimum signal variation across the band. In addition, final trimming can be done by adjusting the level of the output of each video detector in the IF receiver.

One final consideration is the fact that the telescopes are sometimes tapped into the cable at points in between the repeaters (Fig. 1). In this case the gain and equalization must be adjusted to compensate only for the length of coax between the telescope and the next repeater. One rather simple way of doing this is to use a section of high loss coax. This can be inserted before coupling into the link such that the total attenuation slope betwen the telescope and the next repeater is the same as that in one section of the main line. The repeater on the telescope is then identical to those used in the IF link, except that the output power must be 10 db higher to compensate for the coupler loss. High loss coax is available with a loss of as much as 60 db/100 ft. at 2 Gc. We are presently using this coax to simulate the loss between repeaters.

A number of the problems which we have discussed in this report can probably be resolved in the next few months by measurements on several experimental equalizers. However, some of the more general considerations, such as the IM and thermal noise buildup in the IF link, the gain variations across the band, the phase linearity in cascaded repeater amplifiers, the curvature due to dielectric losses, and the spectrum of the IM noise can be determined only by the full scale test of the IF system. Thus, we have progressed about as far as we can in analyzing the equalizer and signal-to-noise problems in the IF transmission system. It is now necessary to proceed with the experimental method.

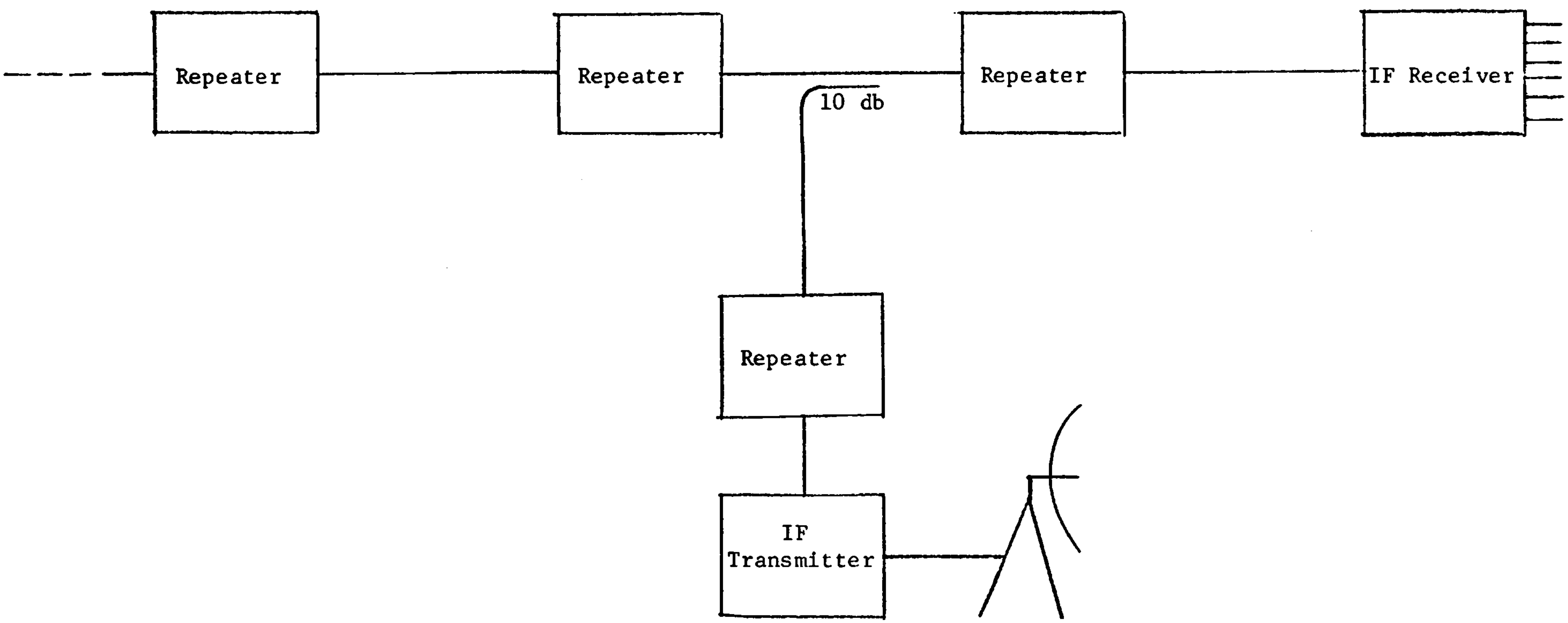


Figure 1

IF Transmission System Block Diagram

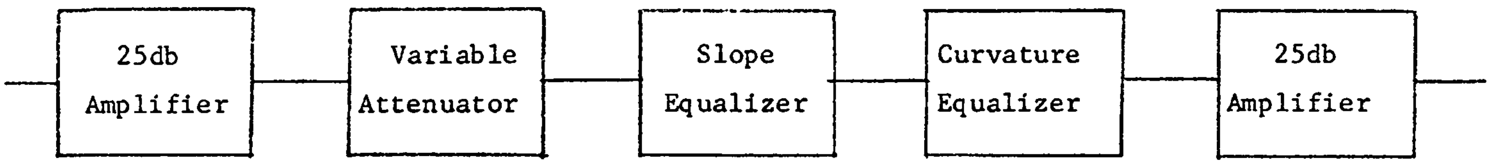


Figure 2

Repeater Block Diagram

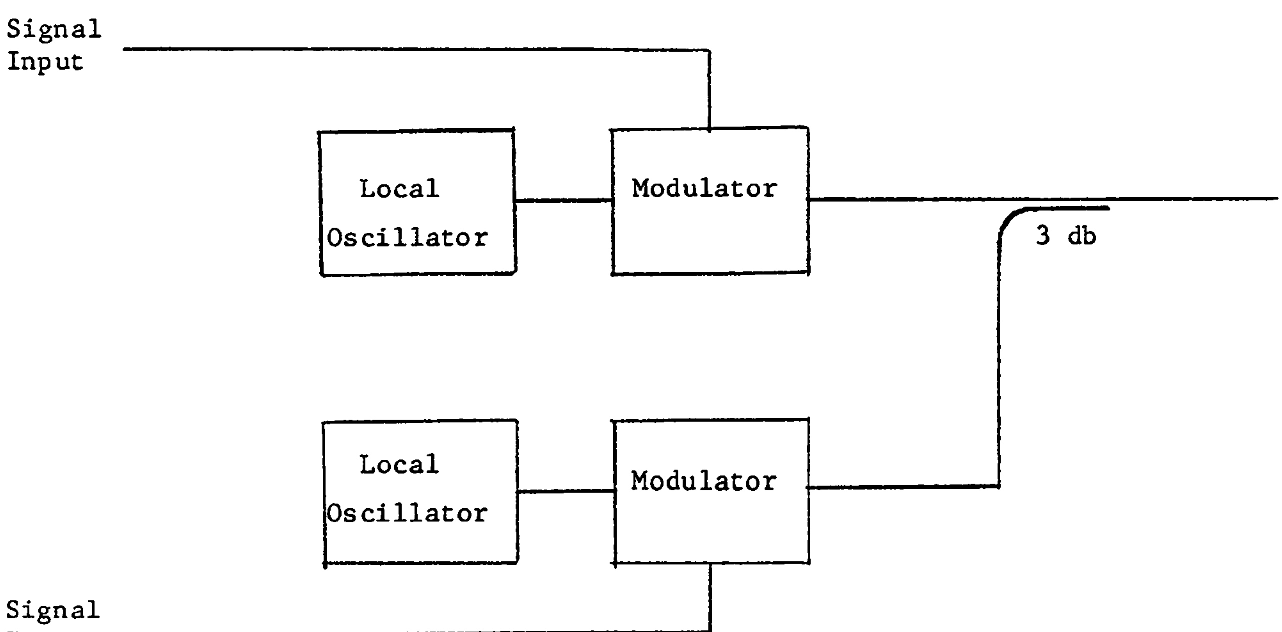


Figure 3

IF Transmitter Block Diagram

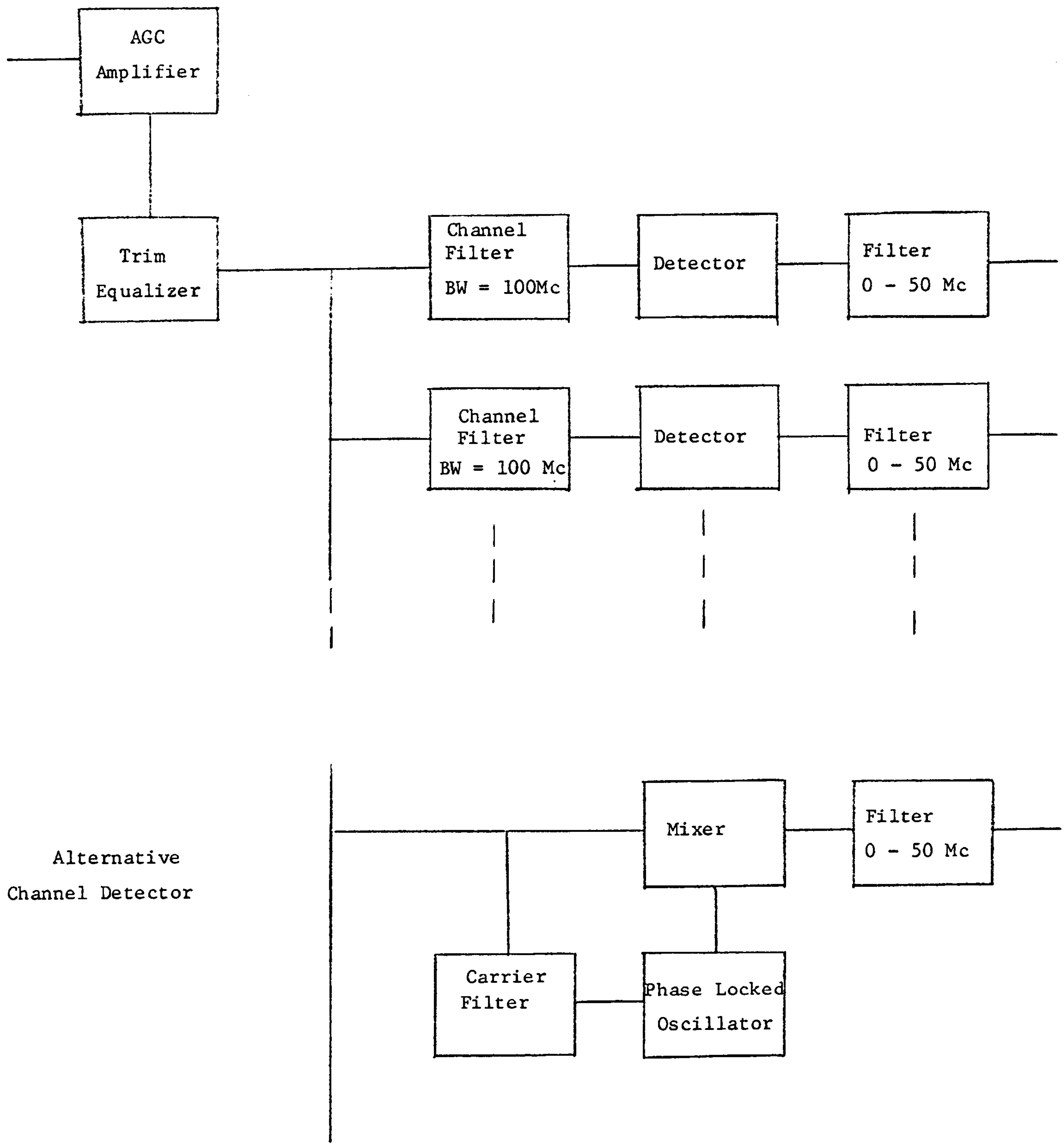


Figure 4

IF Receiver Block Diagram

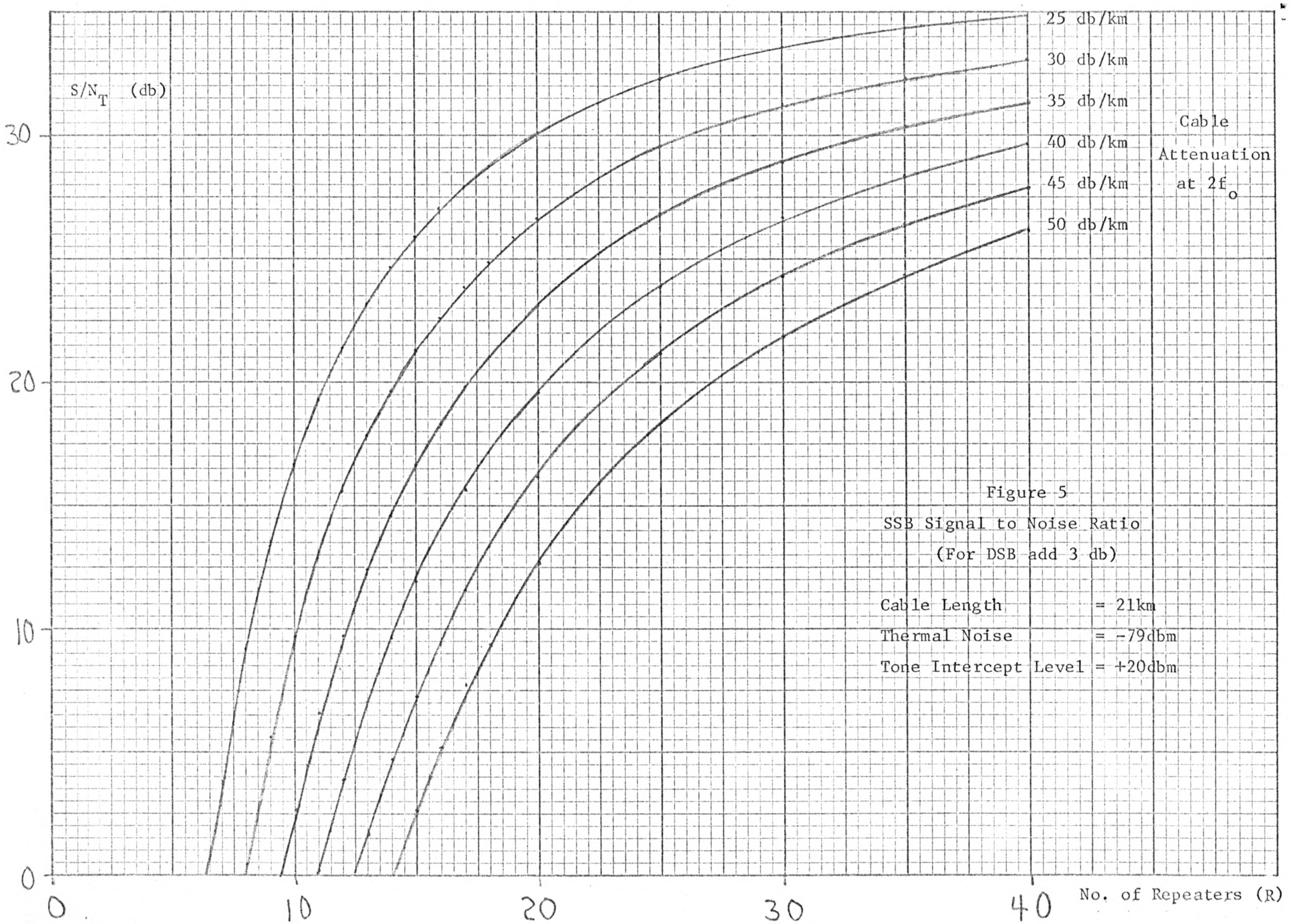


Figure 5

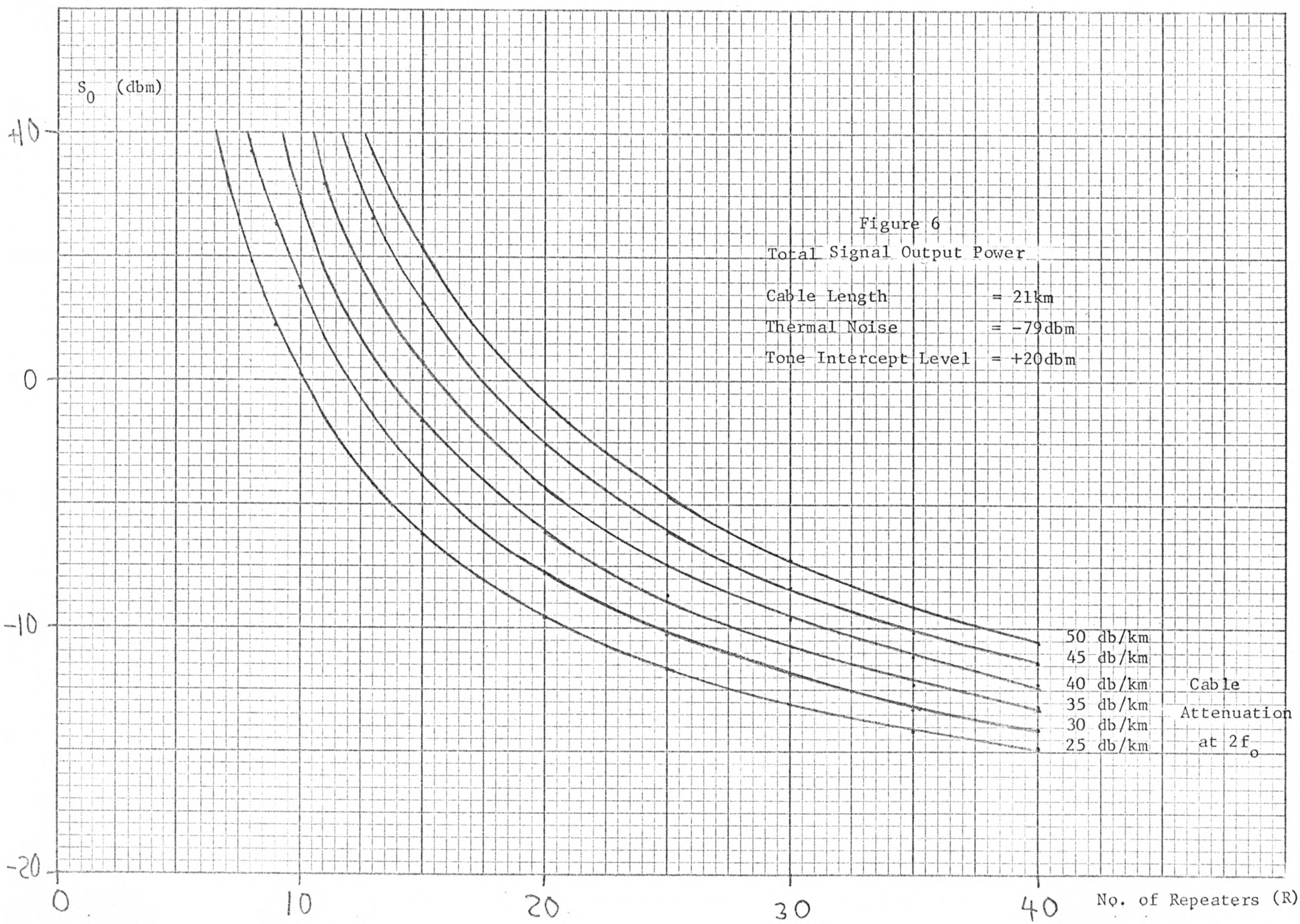
SSB Signal to Noise Ratio

(For DSB add 3 db)

Cable Length = 21km

Thermal Noise = -79dbm

Tone Intercept Level = +20dbm



S_0 (dbm)
+10
0
-10
-20
0
10
20
30
40
No. of Repeaters (R)
Figure 6
Total Signal Output Power
Cable Length = 21km
Thermal Noise = -79dbm
Tone Intercept Level = +20dbm
50 db/km
45 db/km
40 db/km
35 db/km
30 db/km
25 db/km
Cable Attenuation at $2f_o$

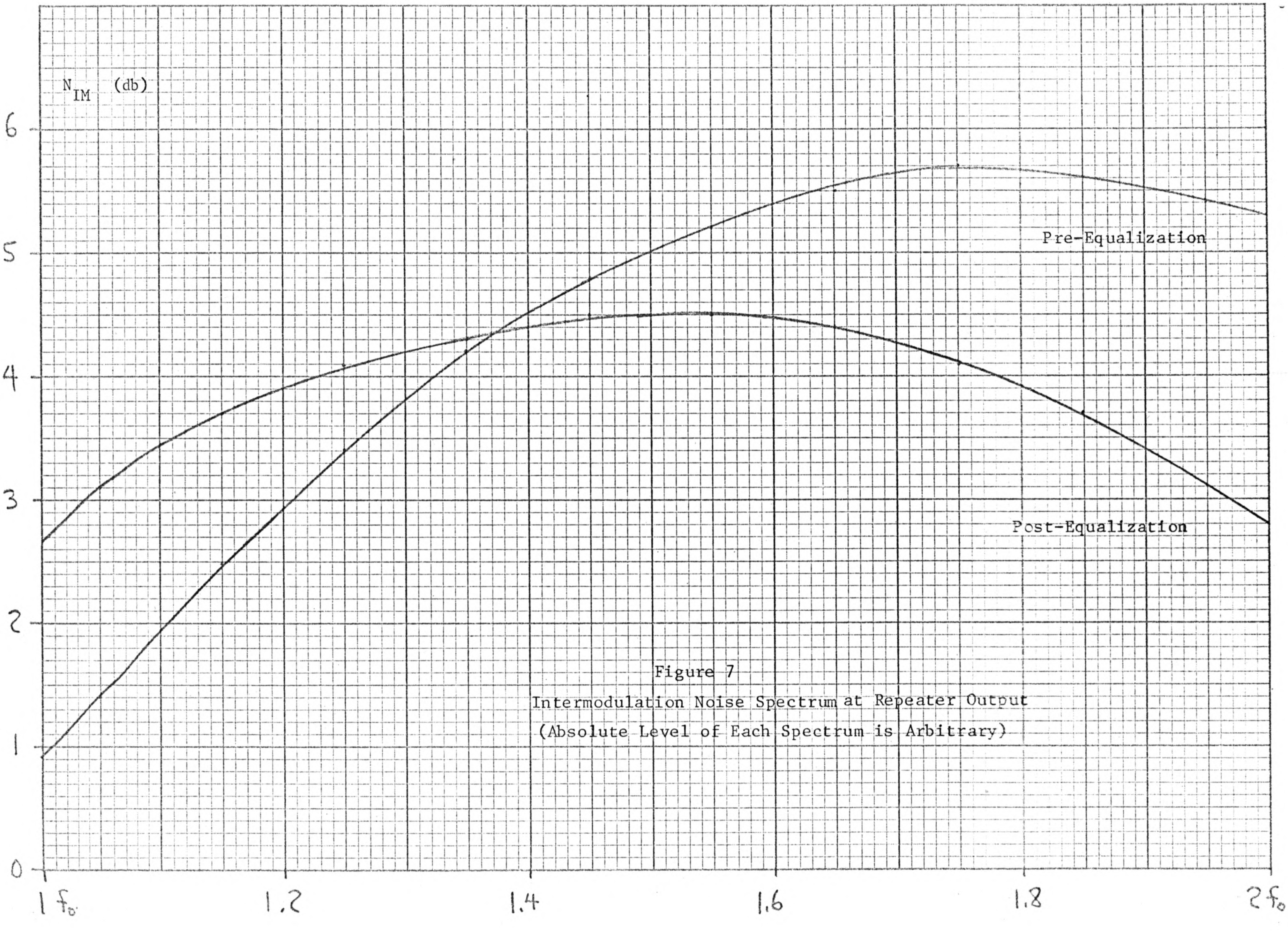


Figure 7
Intermodulation Noise Spectrum at Repeater Output
(Absolute Level of Each Spectrum is Arbitrary)

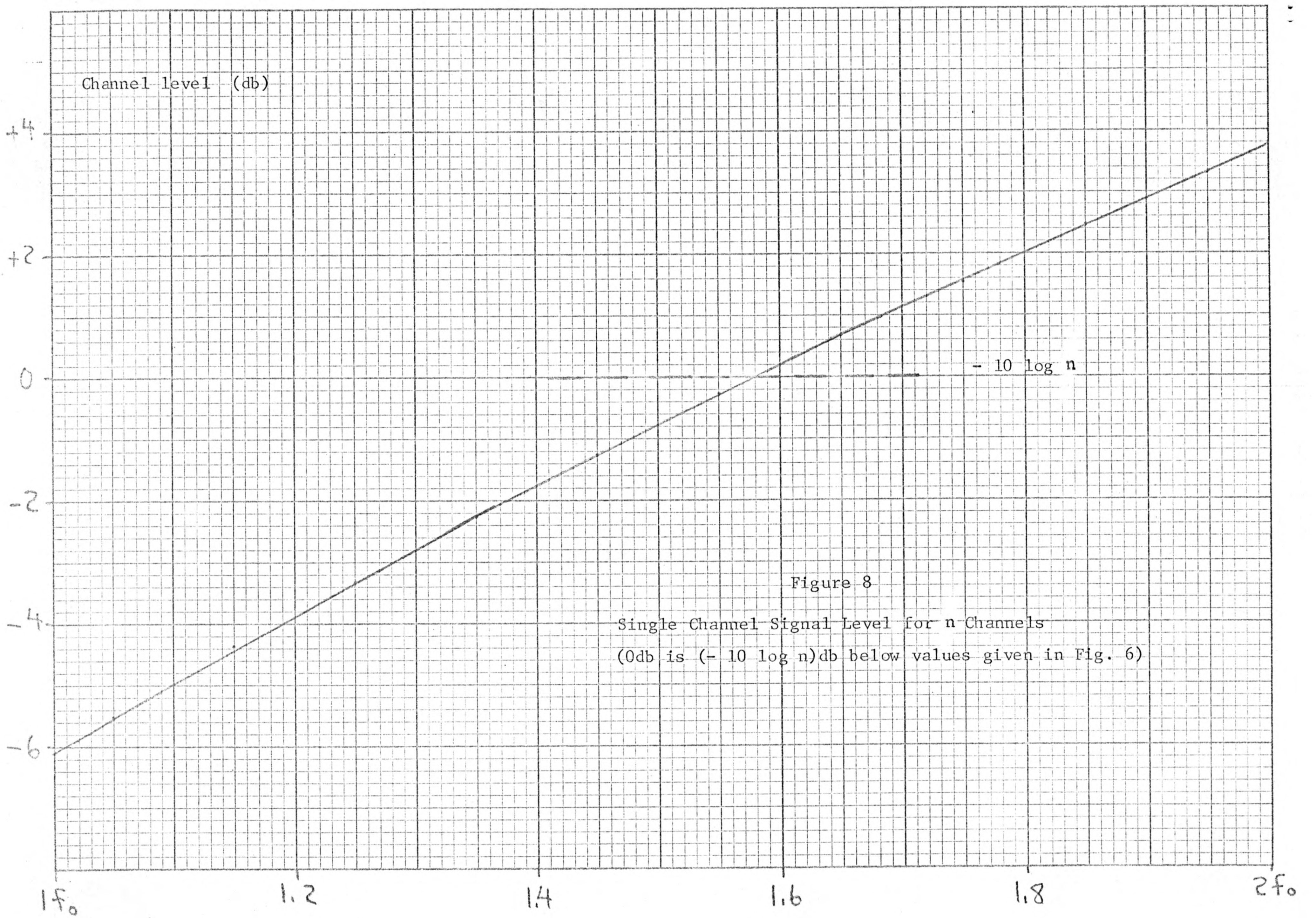


Figure 8

Single Channel Signal Level for n Channels

(0db is (- 10 log n)db below values given in Fig. 6)

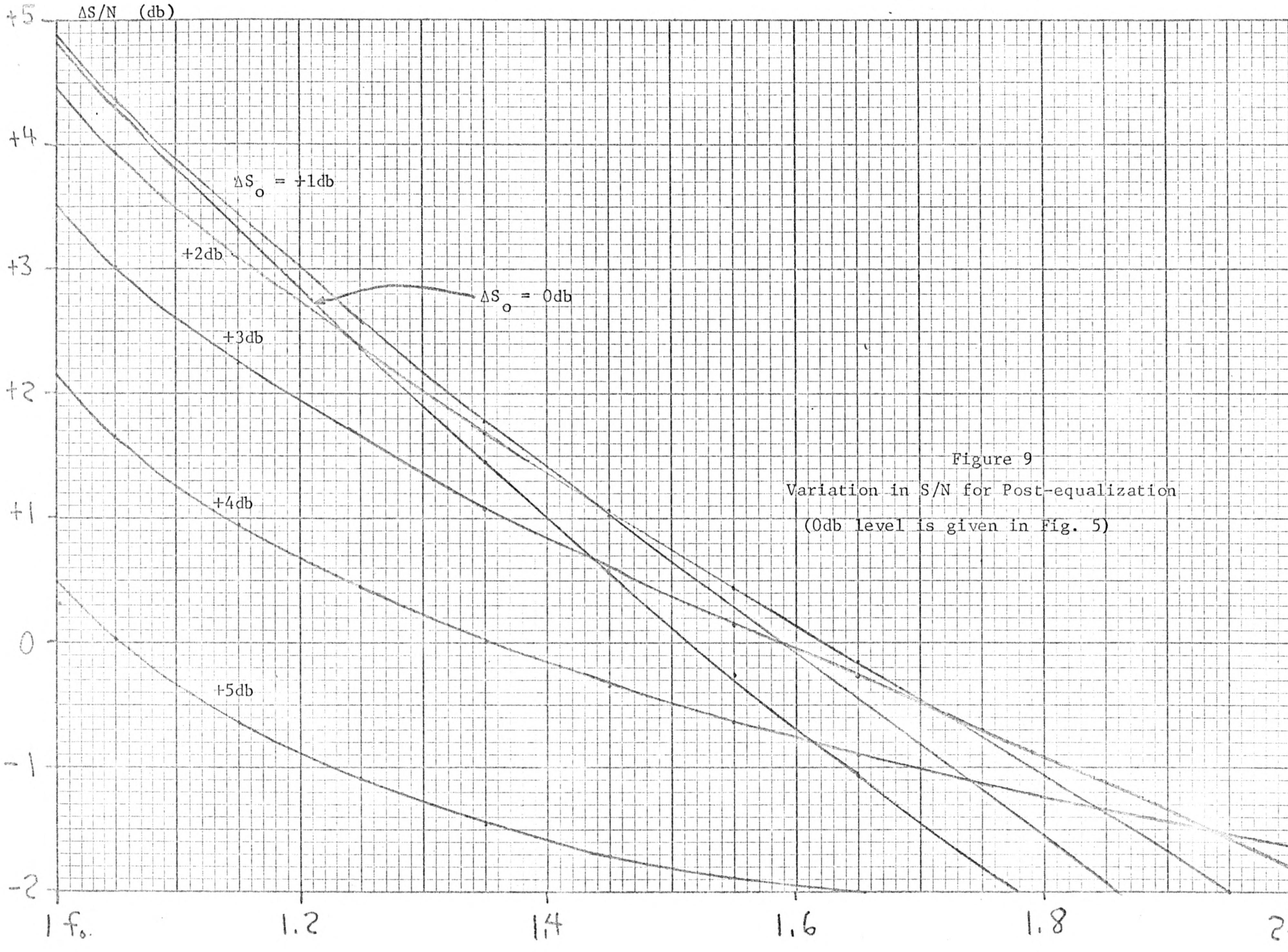


Figure 9

Variation in S/N for Post-equalization

(0db level is given in Fig. 5)

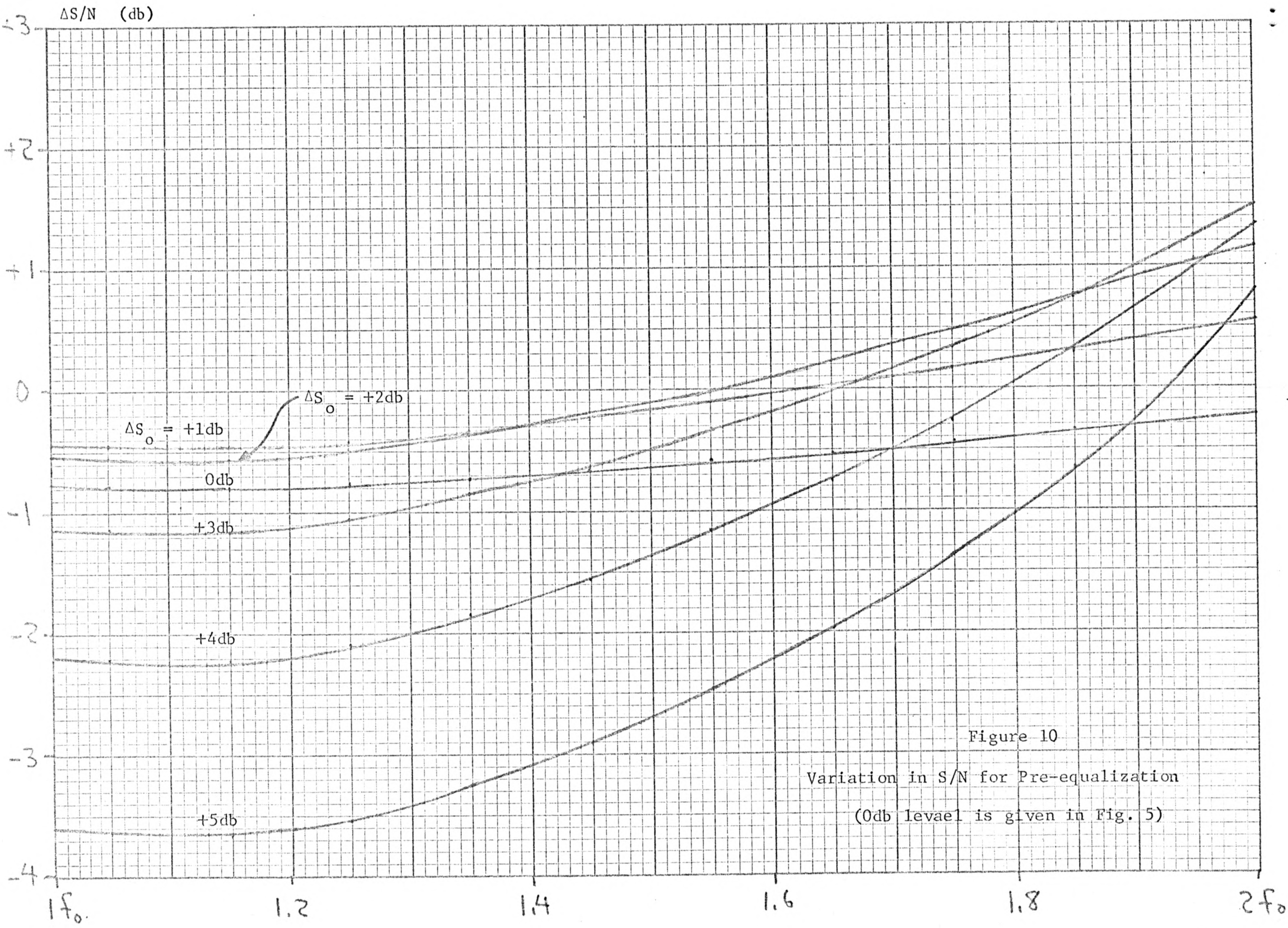


Figure 10

Variation in S/N for Pre-equalization

(0db levael is given in Fig. 5)

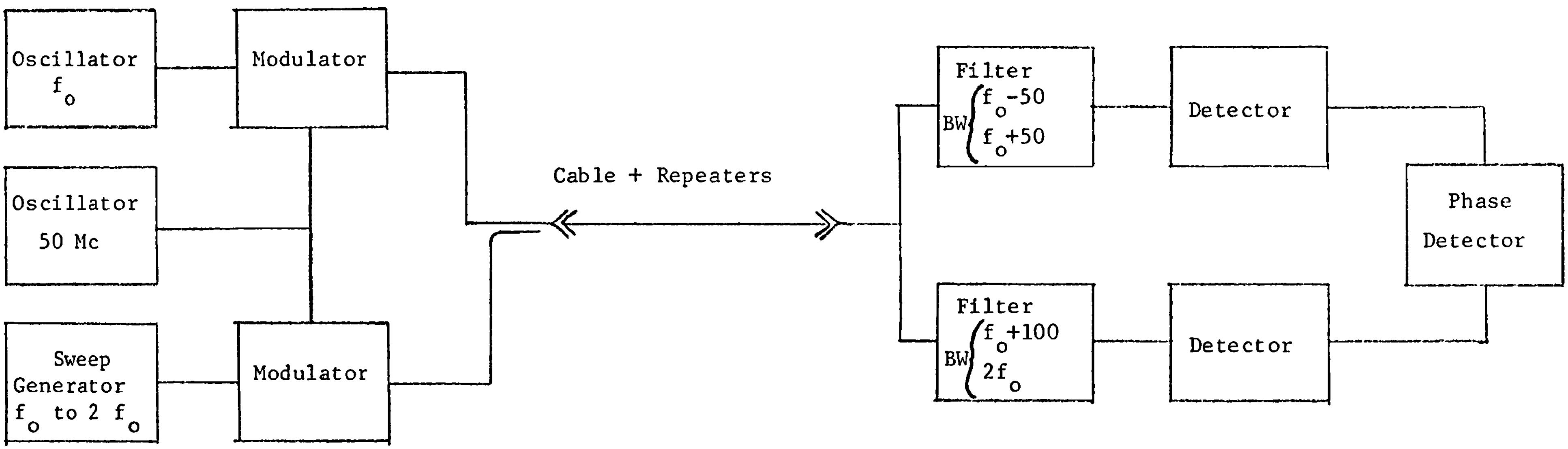
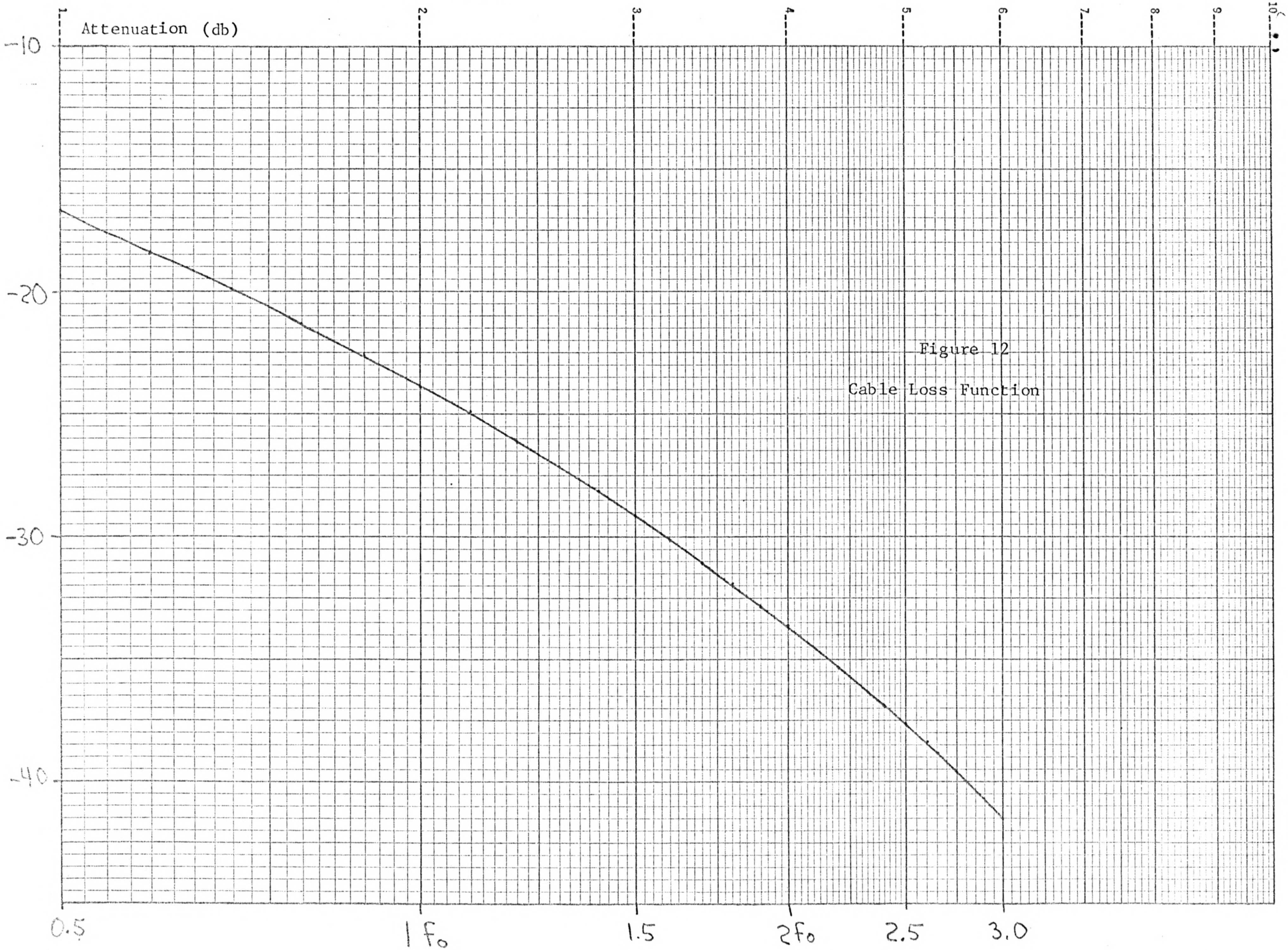


Figure 11

Non-linear Phase Measurement System


Attenuation (db)
-10
-20
-30
-40
0.5
1 fo
1.5
2fo
2.5
3.0
Figure 12
Cable Loss Function

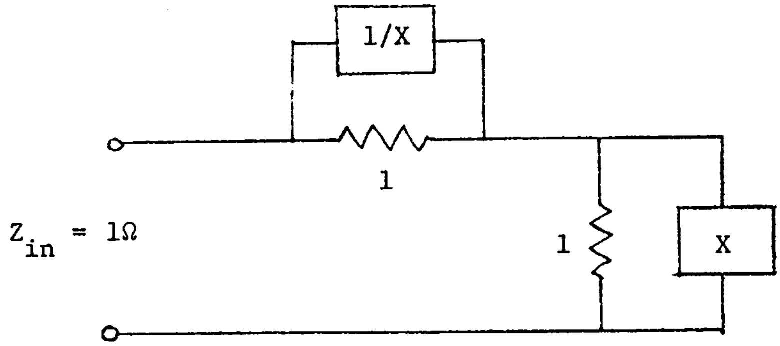


Figure 13

Constant Input Impedance Network

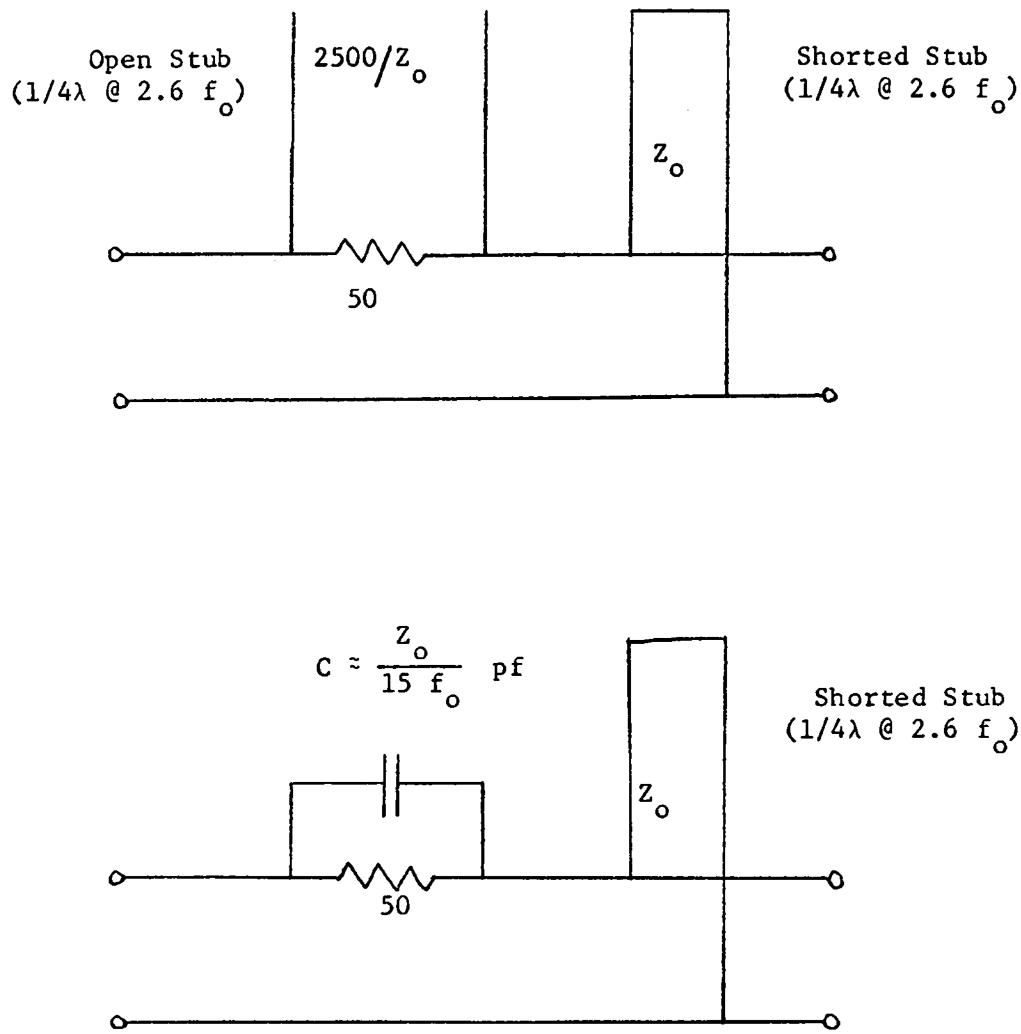


Figure 14

Slope Equalizer

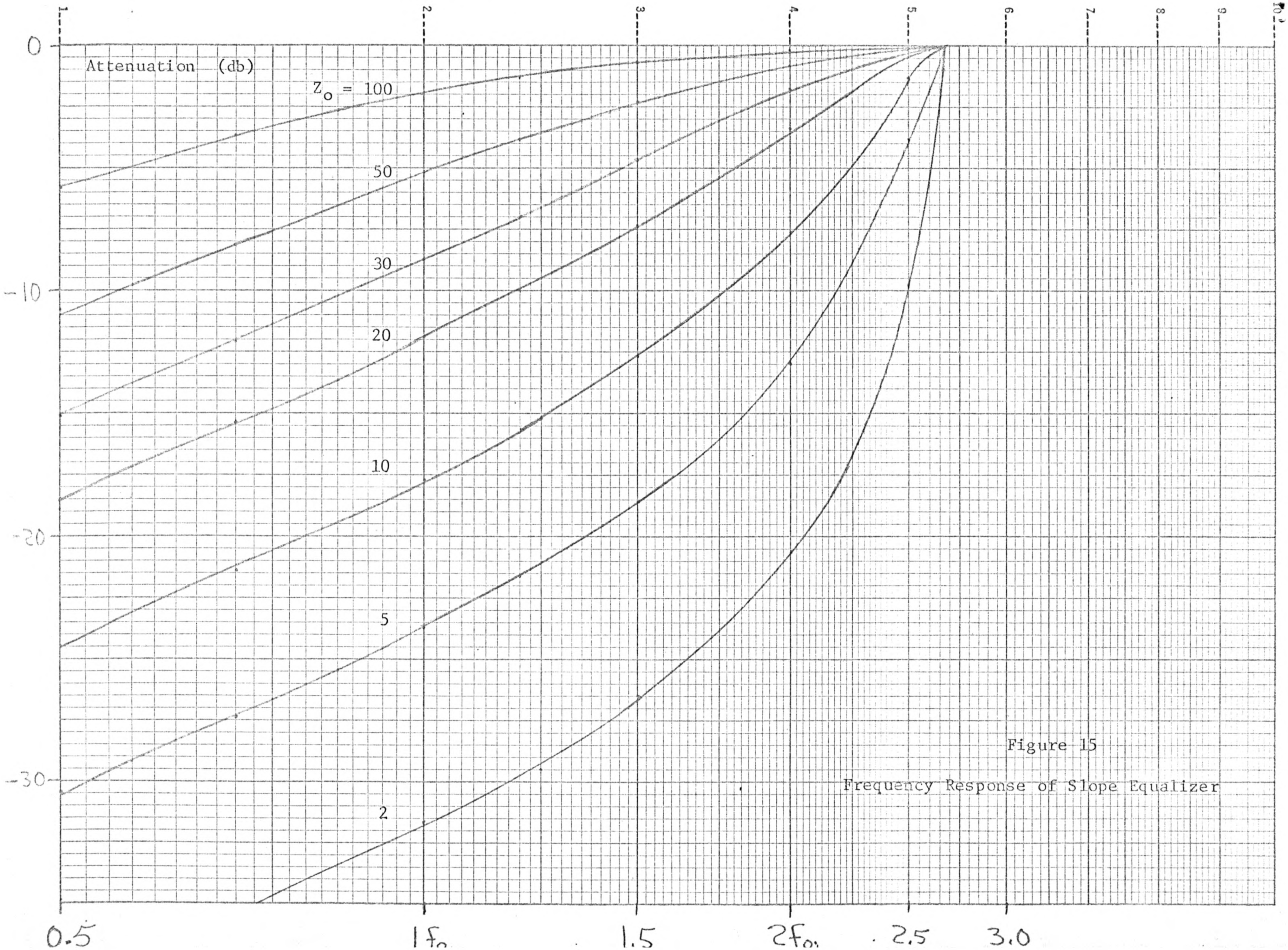


Figure 15

Frequency Response of Slope Equalizer

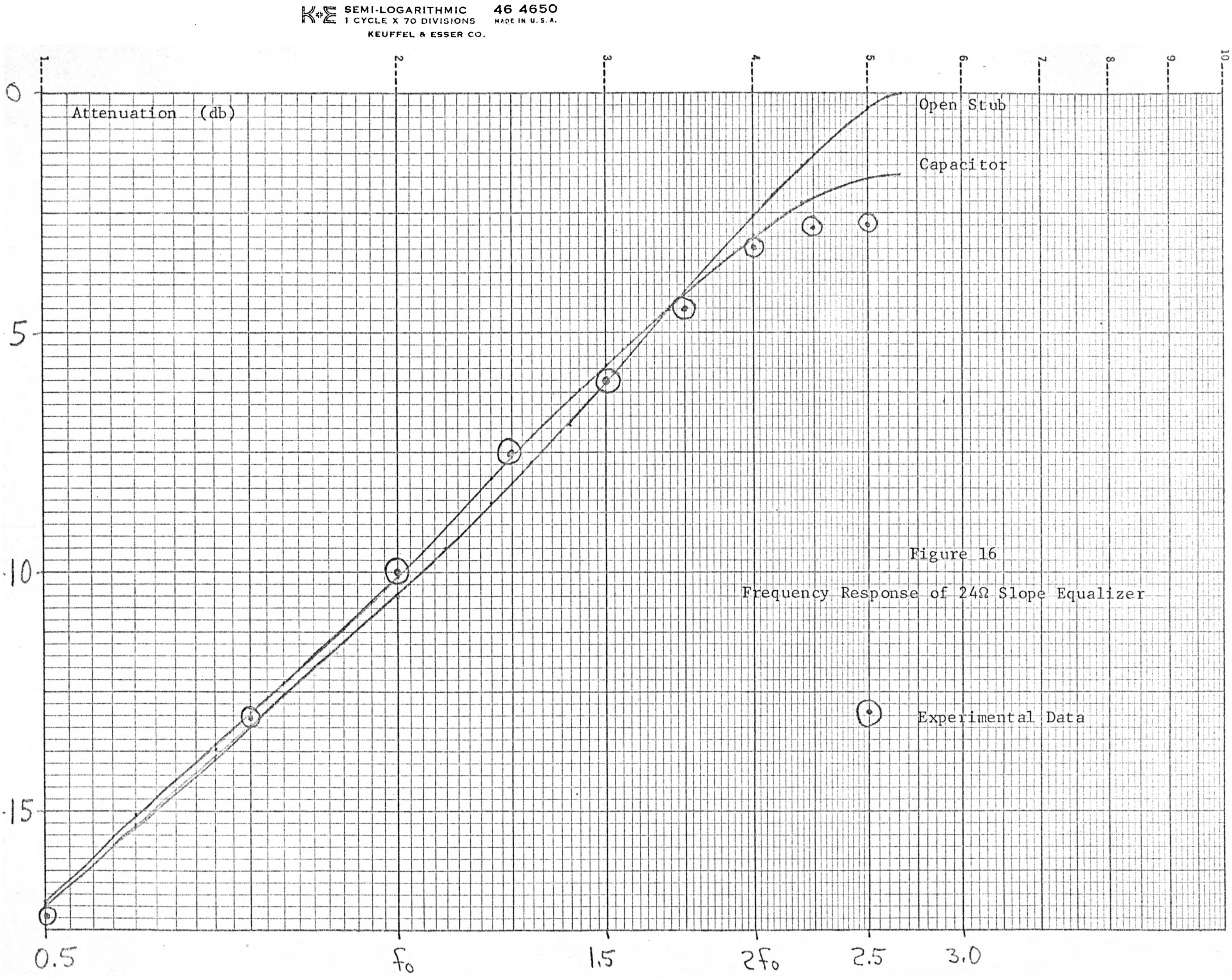


Figure 16

Frequency Response of 24Ω Slope Equalizer

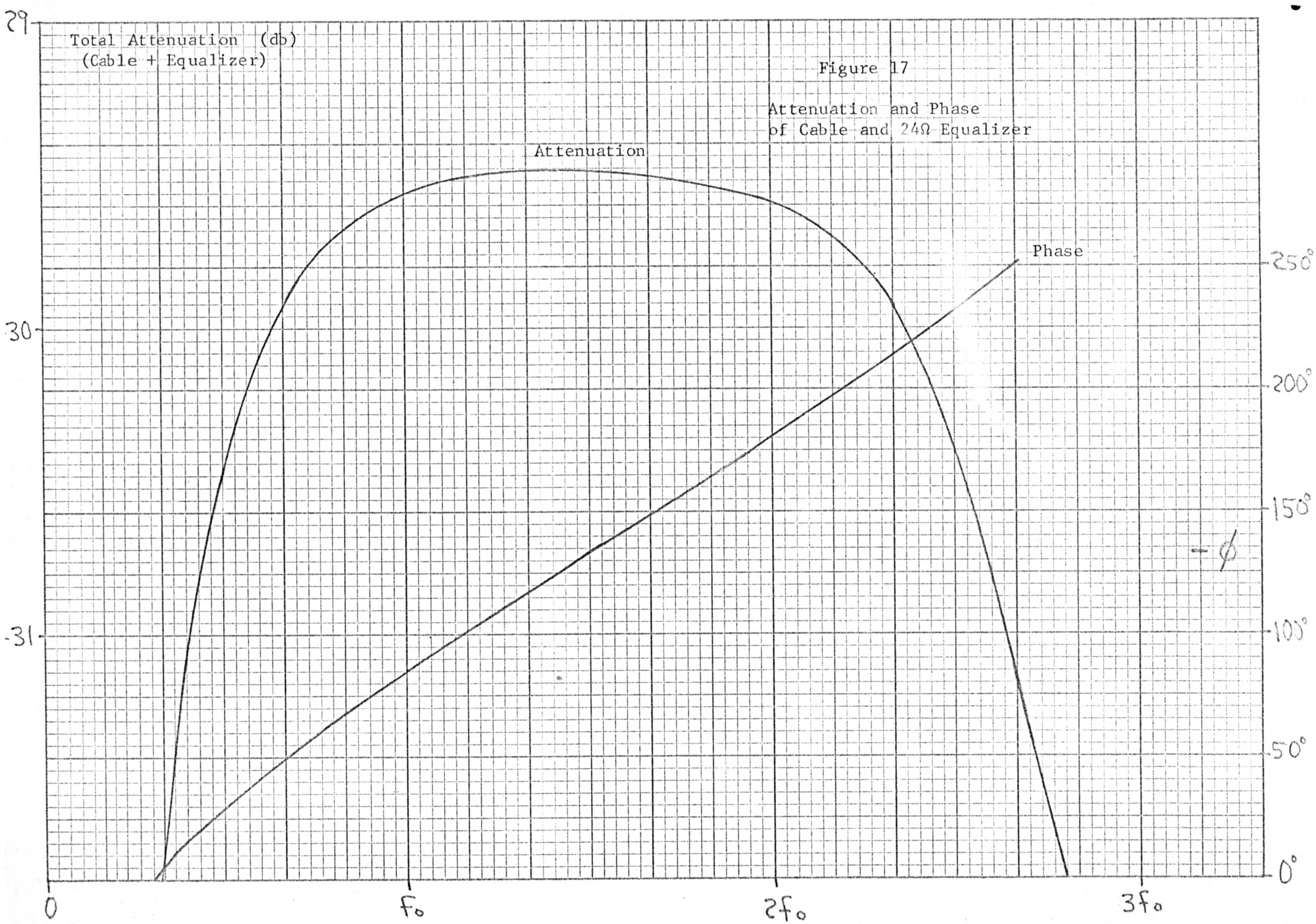


Figure 17

Attenuation and Phase of Cable and 24Ω Equalizer

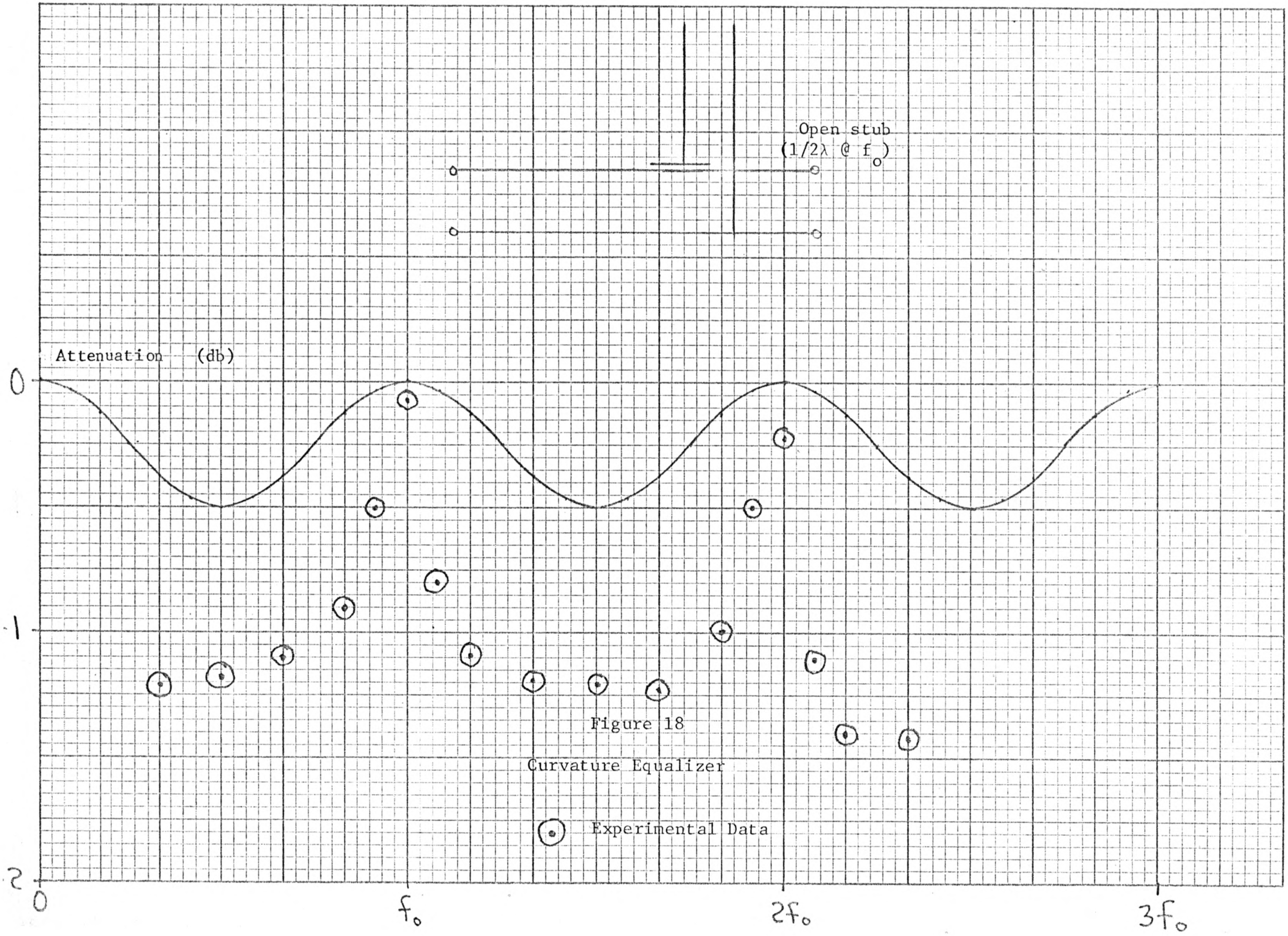


Figure 18

Curvature Equalizer

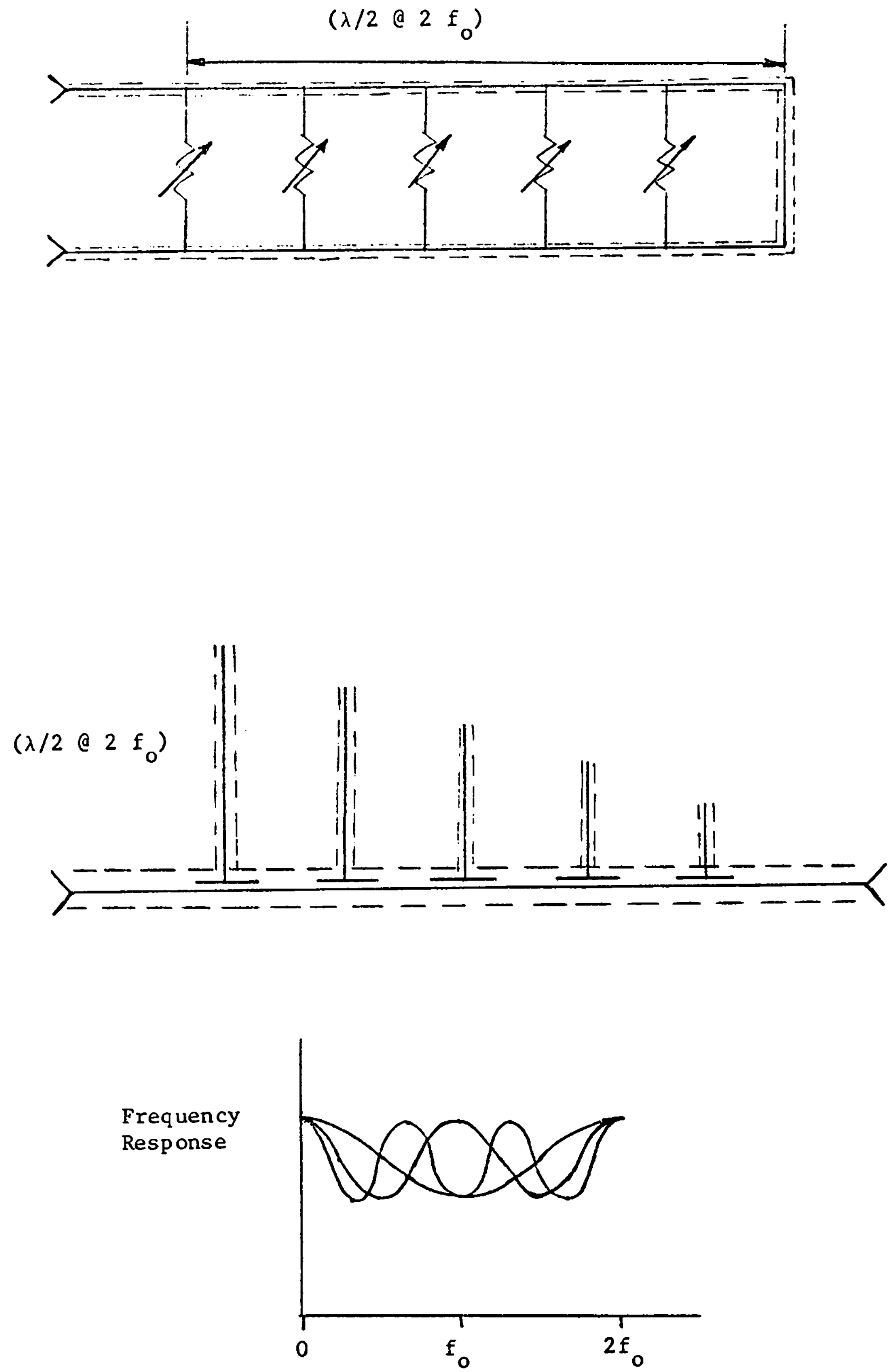


Figure 19

Trim Equalizer