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THESIS

THE DESIGN AND TESTING OF AN ANALOG OPTICAL COMMUNICATION LINK CAPABLE OF THE SIMULTANEOUS TRANSMISSION OF FOUR FREQUENCY DIVISION MULTIPLEXED AUDIO SIGNALS

by

Michael Steven Silvers

June 1987

Thesis Advisor: J.P. Powers

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The Design and Testing of An Analog Optical Communication Link Capable of the Simultaneous Transmission of Four Frequency Division Multiplexed Audio Signals

by

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Submitted in partial fullfillment of the requirements for the degree of

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ABSTRACT

A Communication Link featuring the analog transmission of four simultaneous Frequency Division Multiplexed audio signals, via optical means, was designed, constructed, and experimentally tested. Low cost and common components were utilized throughout the system. Active filter techniques were employed and extended to uncommonly high frequencies. Fidelity of the recovered waveforms proved to be exceptionally high with crosstalk between channels of less than -50 dB.

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I. INTRODUCTION

The subject of this thesis is the design, construction, and experimental testing of a communication system capable of the simultaneous transmission and receipt, via fiber optical cable, of four high fidelity waveforms in the range of 0-20 kHz. The novel aspect of this endeavor is the fact that purely analog techniques are employed.

While each of the components of this system are discussed in detail within the body of this report, a general overview of the decisions affecting the final design (Figure 1), a description of the concepts involved, and a familiarization with the total system layout are presented here.

The reader will appreciate the unusual nature of analog transmission over fiber as almost all such links utililize on-off binary pulse transmission. The analog approach taken with this system avoids the complex circuitry associated with digitization and also achieves transmission of four simultaneous channels without the need for Time Division Multiplexing (TDM).

The design of any communications system begins with the selection of the type of modulation to be used. In this case, FM was the logical choice for two reasons. First of all, FM classically provides superior noise performance over AM and, secondly, the availability of FM transmitters and receivers (Voltage Controlled Oscillators and Phase Lock Loops respectively) made these elements far simpler to implement than their coherent AM counterparts.

With the decision to use FM, the goal of transmitting four simultaneous channels of information becomes a classic Frequency Division Multiplexing (FDM) problem.

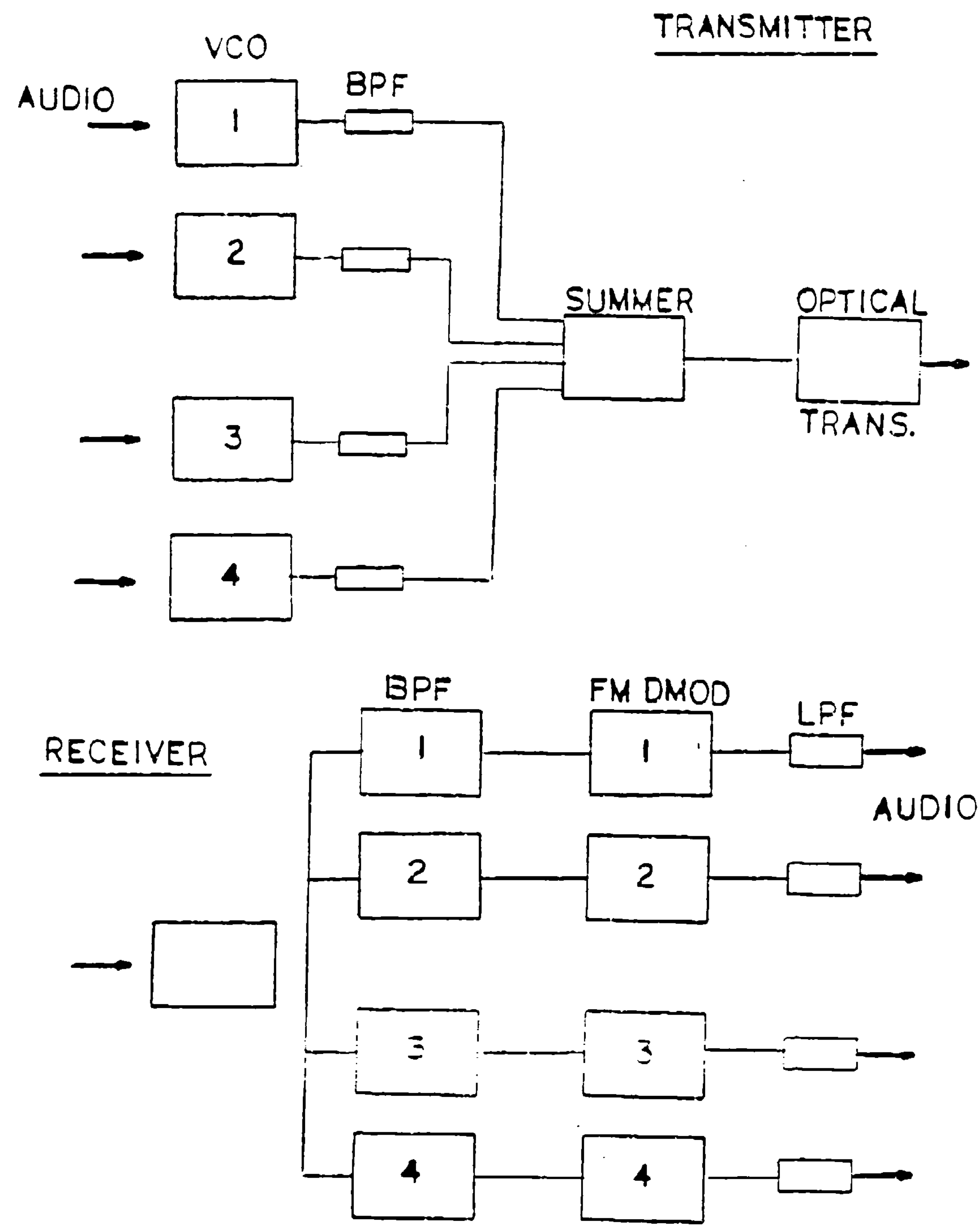


Figure 1. System Block Diagram

The actual multiplexing of the four signals is achieved with a standard voltage summer, as seen in Figure 1. This summed voltage which contains the frequency components of all four channels is then applied to an optical tranamitter for transmission.

The optical transmitter used acts as a light source with an intensity which varies in direct proportio to an input bias volage (i.e., the summed voltage plus a dc offset). For an FM carrier, the peak amplitude of the sinusoid waveform is constant, only the frequency changes. In the optical transmitter, only the intensity changes; the wavelength is fixed at 665 nm. The bridge between the two is achieved by having the optical intensity change at a rate determined by the instantaneous frequency of the FM carrier. For satisfactory transmission, therefore, all that was required was that the LED in the optical transmitter "follow" the complex, rapidly changing input voltage.

The optical receiver performs the inverse function of the optical transmitter and reproduces a FDM signal faithfully. The remainder of the system is unremarkable with the expected bandpass filters to isolate each respective channel and the associated FM receivers as also shown in Figure 1.

The basic system parameters are as follows:

1) A maximum input information signal level of 0.1 volt peak to peak.
2) Center frequencies(f_c) of:
   a) Channel 1: f_c=92 kHz
   b) Channel 2: f_c=325 kHz
   c) Channel 3: f_c=477 kHz
   d) Channel 4: f_c=700 kHz

   Note: The positioning of these channels was driven by performance characteristics of the band-pass filters and attention to harmonic interference. The final placement, however, was to a great extent, trial and error.

3) A minimum acceptable attenuation of the receiver bandpass filters of -40 dBv. This number was experimentally determined by varying a test tone set to a frequency near to an operating channel. With the carrier fixed in amplitude at 2 volts, peak-to-peak, 20 millivolts was the maximum permissible "bleed through", hence -40 dBv.

With this broad overview and Figure 1 firmly fixed in mind, the reader is now invited to examine each of the major components of this system in detail.

II. TRANSMITTER

The transmitter group, Figure 2, consists of:

1) four FM modulators
2) four bandpass filters
3) a summing amplifier and
4) an optical transmitter

To reiterate, the overall aim of these subsystems is the Frequency Division Multiplexing of four analog information channels capable of high fidelity waveform transmission in the 0 to 20 kHz range. With this purpose in mind, each subsystem is discussed as to its construction and design, its peculiarities, and its contribution towards the desired goal.

A. THE FM MODULATOR

The task of any modulator is to accept a baseband information signal as the input and to output a higher frequency carrier signal with some characteristic impressed upon it (modulation) which permits a suitable distant receiver to recover (demodulation) the original baseband information signal. In FM modulation, the type chosen for this system, the characteristic is a variation of the carrier frequency proportional to the information signal. The mathematical description of this operation is

$$S_{FM} = A\cos(\omega_0 t + \theta_0 + k_{FM}\int f(t)dt) \quad (1)$$

where A is the amplitude of the carrier, f(t) is the information signal, θ_0 is an arbitrary phase angle, k_{FM} is an arbitrary positive constant, the expression $\omega_0 t + \theta_0 + k_{FM}\int f(t)d(t)$ is the instantaneous phase angle, and S_{FM} is the FM signal itself.

There are many methods of accomplishing this feat, some very complex. The method chosen, however, is known as

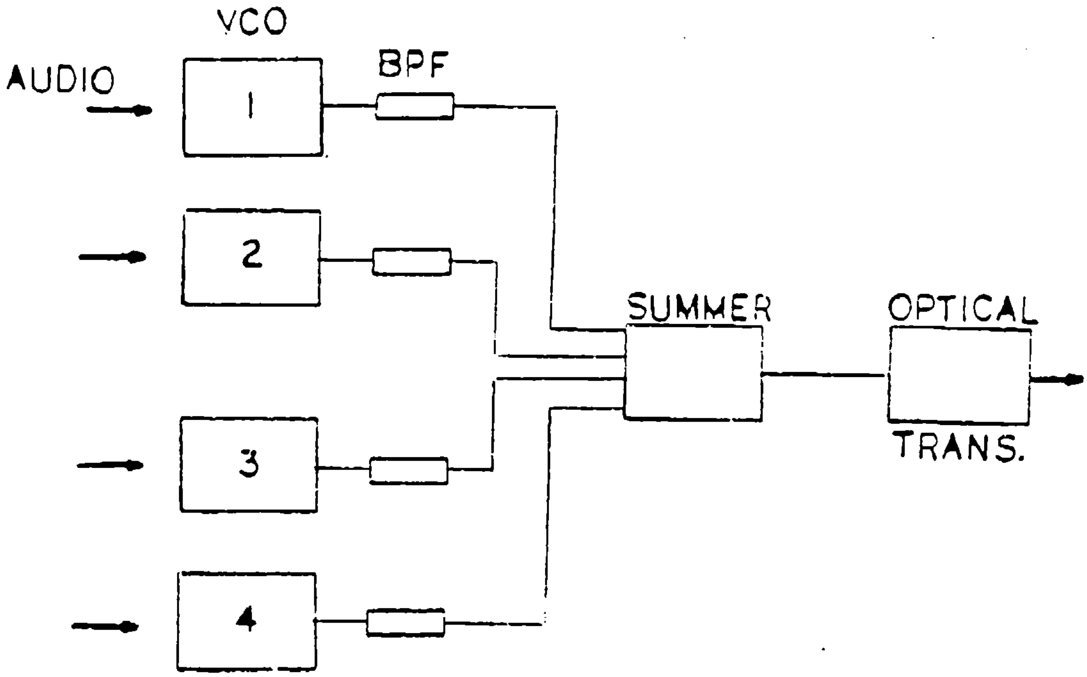


Figure 2. Transmitter Group

direct FM modulation and is made exceedingly simple through the use of a voltage controlled oscillator (VCO) as the actual modulator.

The VCO selected was the XR-2206 Monolithic Function Generator, described in Reference 9. This particular device was used because of its range of frequency operation, (0.01 Hz to 1 MHz), its low sinewave, hence carrier, distortion (0.5%), and its low FM distortion (<10%). Employed as an FM modulator, Figure 3 and Table 1, this device outputs a frequency-modulated sinewave carrier proportional to an input analog voltage, V_c. The basic governing equation is simple in that

$$f_0 = 1/RC \qquad (2)$$

where f_0 is the free running frequency of oscillation with

$$R = R5 + R6 \qquad (3)$$

and

$$C = C4 \qquad (4)$$

Additionally, the instantaneous frequency of oscillation as a function of V_c is

$$f_{inst} = \frac{1}{R(1+R/R1(1-V_c/3)} \qquad (5)$$

with V_c a maximum of 0.1 volt. The voltage to frequency conversion gain is

$$k = -0.32 / (R1 \times C) \qquad (6)$$

As the carrier itself is a sinewave, low distortion is desired in order to prevent unnecessary frequency components from entering adjacent FDM signals. These components cause interference with adjacent channels and complicate the receiver filtering operation. Therefore, the schematic of

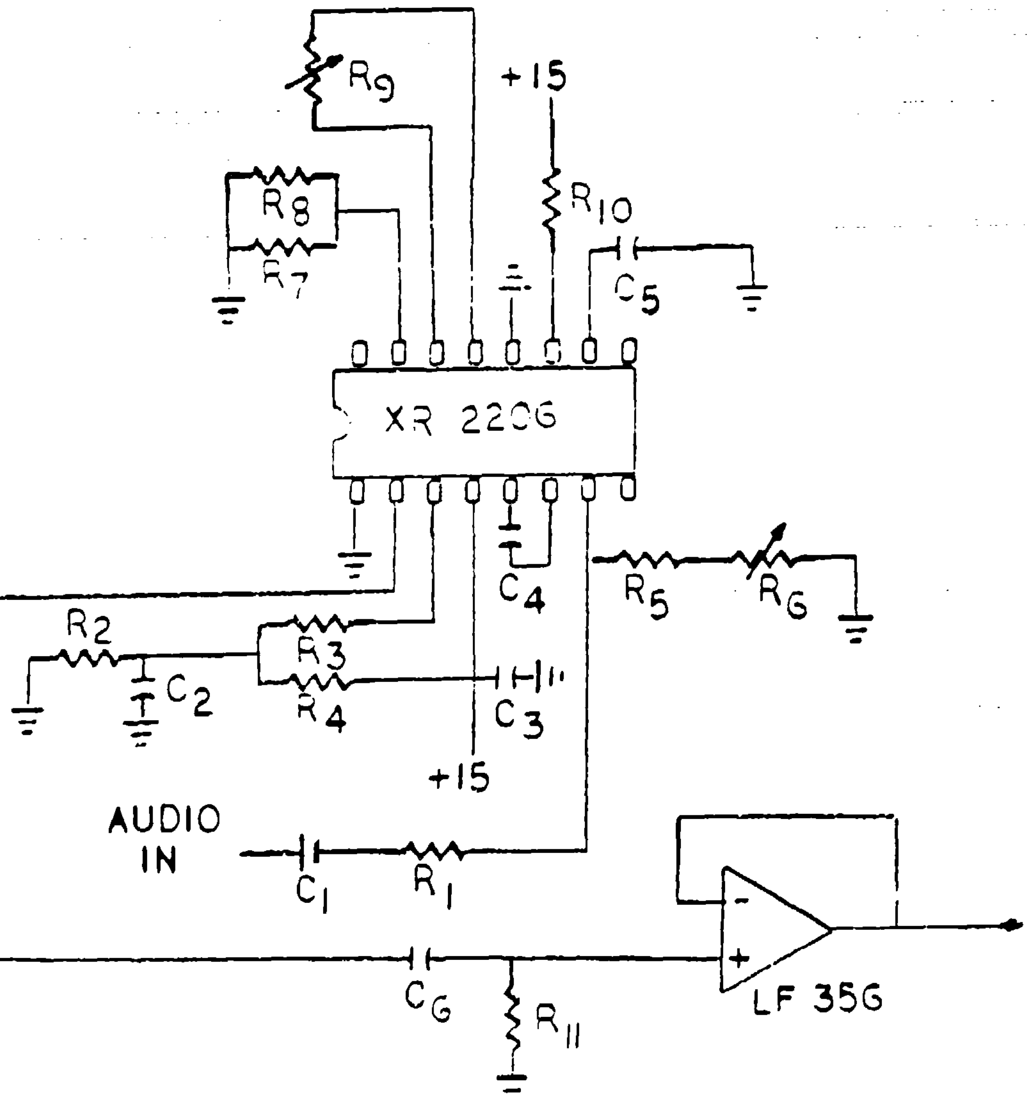


Figure 3. FM Modulator

TABLE 1

TRANSMITTER COMPONENT VALUES

Channel 1

R1 = 7.5k	C1 = 1 uf
R2 = 5.1k	C2 = 10 uf
R3 = 20.0k	C3 = 1 uf
R4 = 5.1k	C4 = 1 nf
R5 = 13.0k	C5 = 1 uf
R6 = 0-1.0k	C6 = 100 uf
R7 = 20.0k	
R8 = 30.0k	
R9 = 0-5.0k	
R10 = 10.0k	
R11 = 36.0k	

Channel 2

R1 = 3.0k	C1 = 1 uf
R2 = 5.1k	C2 = 10 uf
R3 = 20.0k	C3 = 1 uf
R4 = 5.1k	C4 = 1 nf
R5 = 2.4k	C5 = 1 uf
R6 = 0-1.0k	C6 = 100 uf
R7 = 20.0k	
R8 = 30.0k	
R9 = 0-1.0k	
R10 = 10.0k	
R11 = 36.0k	

Channel 3

R1 = 2.0k	C1 = 1 uf
R2 = 5.1k	C2 = 10 uf
R3 = 20.0k	C3 = 1 uf
R4 = 5.1k	C4 = 1 nf
R5 = 1.1k	C5 = 1 uf
R6 = 0-1.0k	C6 = 100 uf
R7 = 20.0k	
R8 = 30.0k	
R9 = 0-1.0k	
R10 = 10.0k	
R11 = 36.0k	

Channel 4

R1 = 2.4k	C1 = 1 uf
R2 = 5.1k	C2 = 10 uf
R3 = 20.0k	C3 = 1 uf
R4 = 5.1k	C4 = 750 pf
R5 = 1.5k	C5 = 1 uf
R6 = 0-500	C6 = 100 uf
R7 = 20.0k	
R8 = 30.0k	
R9 = 0-1.0k	
R10 = 10.0k	
R11 = 36.0k	

Note: Tolerance of all Resistors: +/- 5%
Tolerance of all Capacitors: +/- 10%

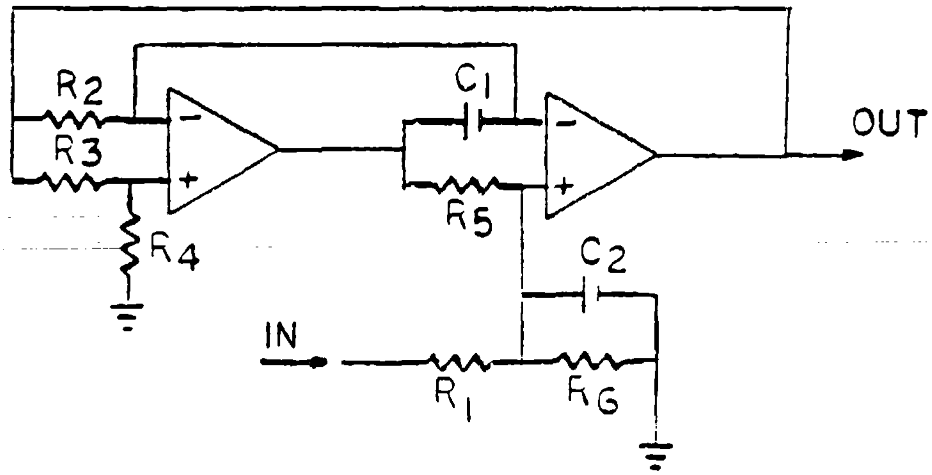
Figure 3 shows a version of this VCO which is especially adjustable for low sinewave distortion. Particularly, resistors R7, R8, and R9 serve this purpose. Resistors R7 and R8 were originally a variable potentiometer arrangement to permit fine symmetry tuning while R9 is a basic shape tuner which was left variable for experimentation. Additionally, the dC removal/voltage isolation network of R11 and C6, and a voltage follower are shown. In all respects, the XR-2206 performed according to expectations with the sole problem being that of a tendency not to return to the exact center frequency at each start-up. This quirk mandated a slight tuning capability, R6.

B. TRANSMITTER BANDPASS FILTERS

The bandpass filter of Figure 4 and Table 2 is known as a Generalized Immittance Convertor (GIC). A thorough discussion of its attributes is included in Chapter III as it is used on a grander scale in the receiver. For the moment, therefore, only its purpose in the transmitter will be discussed.

As mentioned previously in the section on the FM modulator, great care was taken to ensure that the carrier was as pure a sinewave as possible. Even with the fine tuning circuit, however, some slight distortion was still evident. The task of these filters is to remove this remaining distortion. Figures 5,6,7, and 8 are the spectra of the unmodulated carriers after filtering. A word about the spectral analysis figures is in order at this point. These figures are direct plots from the Hewlett-Packard 3566B Spectrum Analyser. The important quantities (e.g., see Figure 5) are:

1) Center: This is the frequency at the center vertical line.



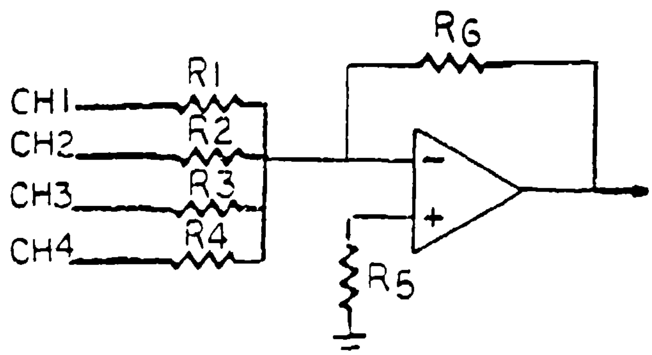


Figure 4. Transmitter Bandpass Filter and Summing Amplifier

TABLE 2

TRANSMITTER BANDPASS FILTER AND SUMMING AMPLIFIER COMPONENT VALUES

BANDPASS FILTER

Channel 1

R1 = 5.1k   C1 = C2 = 750 pf
R2 = R3 = R4 = R5 = 2.4k
R6 = 50.0k

Channel 2

R1 = 10.0k   C1 = C2 = 220 pf
R2 = R3 = R4 = R5 = 1.8k
R6 = 10.0k

Channel 3

R1 = 11.1k   C1 = C2 = 150 pf
R2 = R3 = R4 = R5 = 1.0k
R6 = 20.0k

Channel 4

R1 = 8.2k   C1 = C2 = [illegible] pf
R2 = R3 = R4 = R5 = [illegible]
R6 = 20.0k

VOLTAGE SUMMING AMPLIFIER

R1 = 20.0k, R2 = 20.0k, R3 = 39.0k, R4 = 39.0k, R5 = 5.1k
R6 = 1.0k

2) Span: This is the frequency coverage of the entire graph.

3) DL: When present, this is the horizontal line denoting a 0.0 dBm reference level and is the line above and to the right of the inscription.

4) Ref: This is the noise floor height in dBm.

5) 10 dB/: This indicates that each horizontal grid equals 10 dB.

6) MKR: When present, this indicates the frequency and distance in dB below the DL of the small diamond marker This is included when highlighting of a single component is desired.

Other quantities which appear but were not used are

7) RES BW: Resolution Bandwidth permits the expansion of the trace, (i.e., small values yield a large expansion and large values yield a small expansion).

8) VBW: Video Bandwidth permits smoothing of the trace, (i.e., small values yield a smooth trace and large values yield a ragged trace).

9) SWP: Sweep time is the time taken to present an updated spectrum calculation on the screen

10) ATTEN: Attenuation is the amount that the input signal is attenuated.

Returning now to Figures 5,6,7, and 8 (the spectra of the unmodulated carriers after filtering) it is apparent that any local spurious frequencies are indeed removed. Note: By increasing the span, the numbered harmonics are still detectable, however, they occur at the predicted positions, are attenuated 50 dBm, and are of no consequence.

C. SUMMING AMPLIFIER

The actual multiplexing of the four FM signals is achieved via a standard voltage summer, Figure 4 and Table

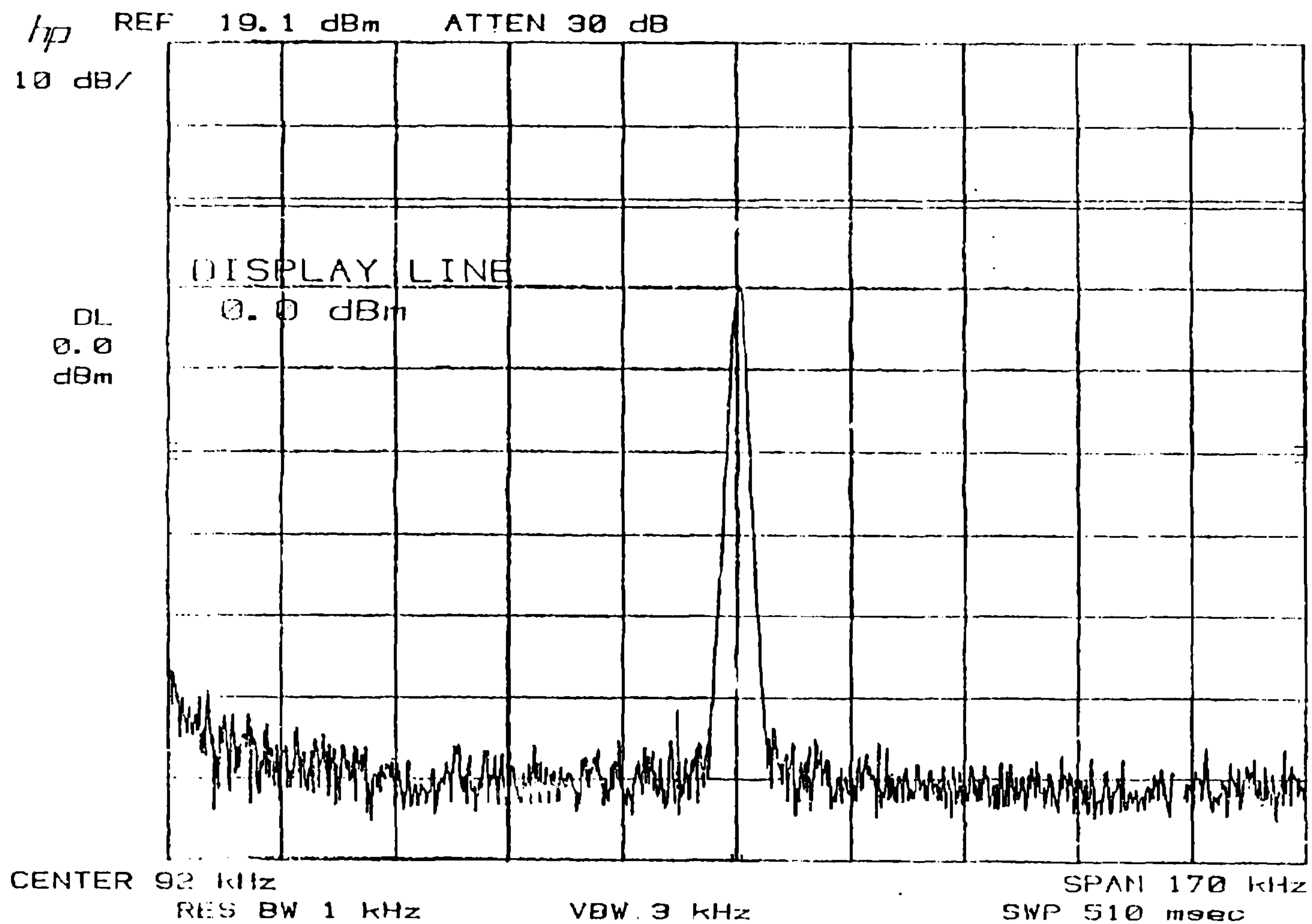


Figure 5. Carrier Spectral Analysis, Channel 1

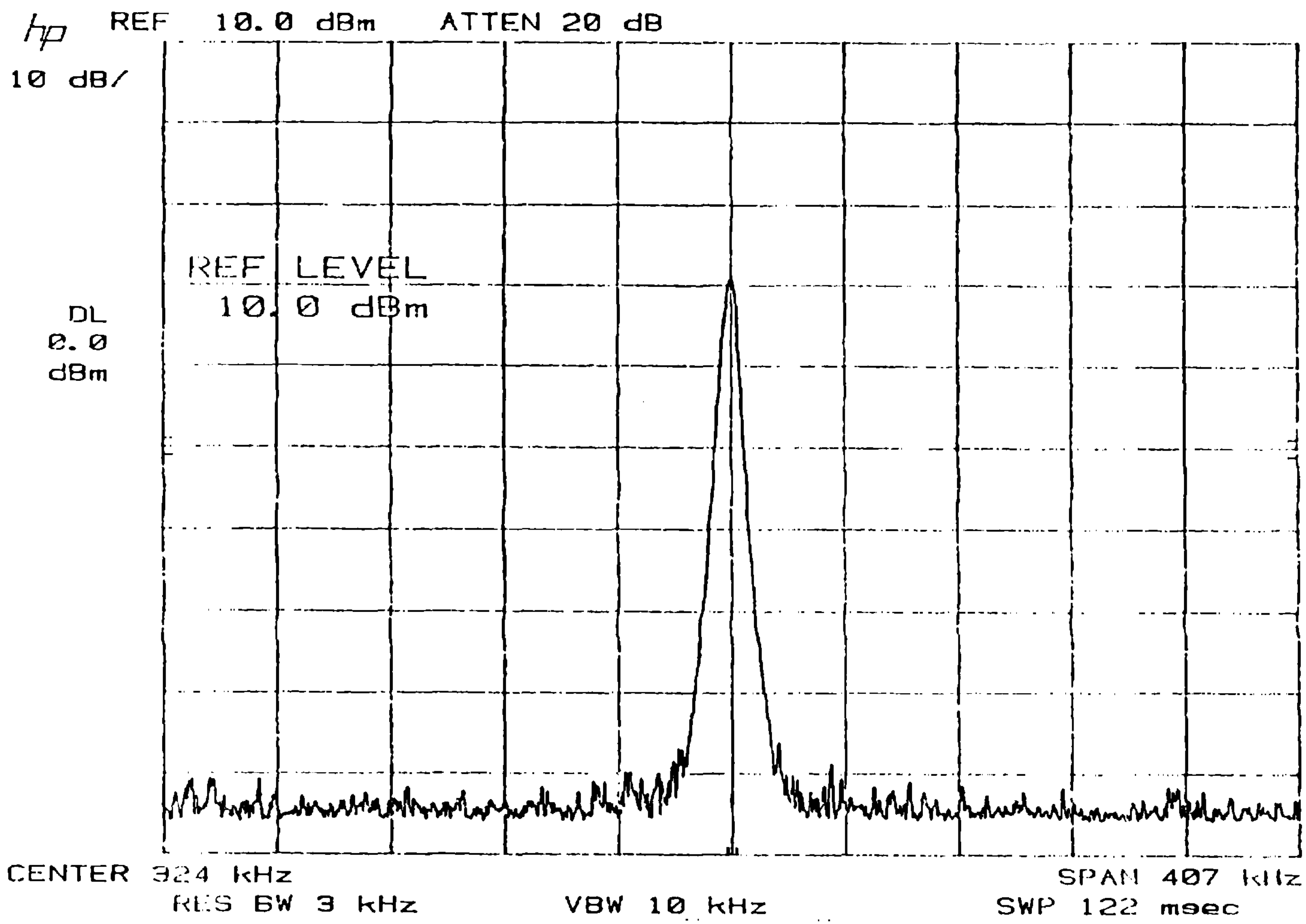


Figure 6. Carrier Spectral Analysis, Channel 2

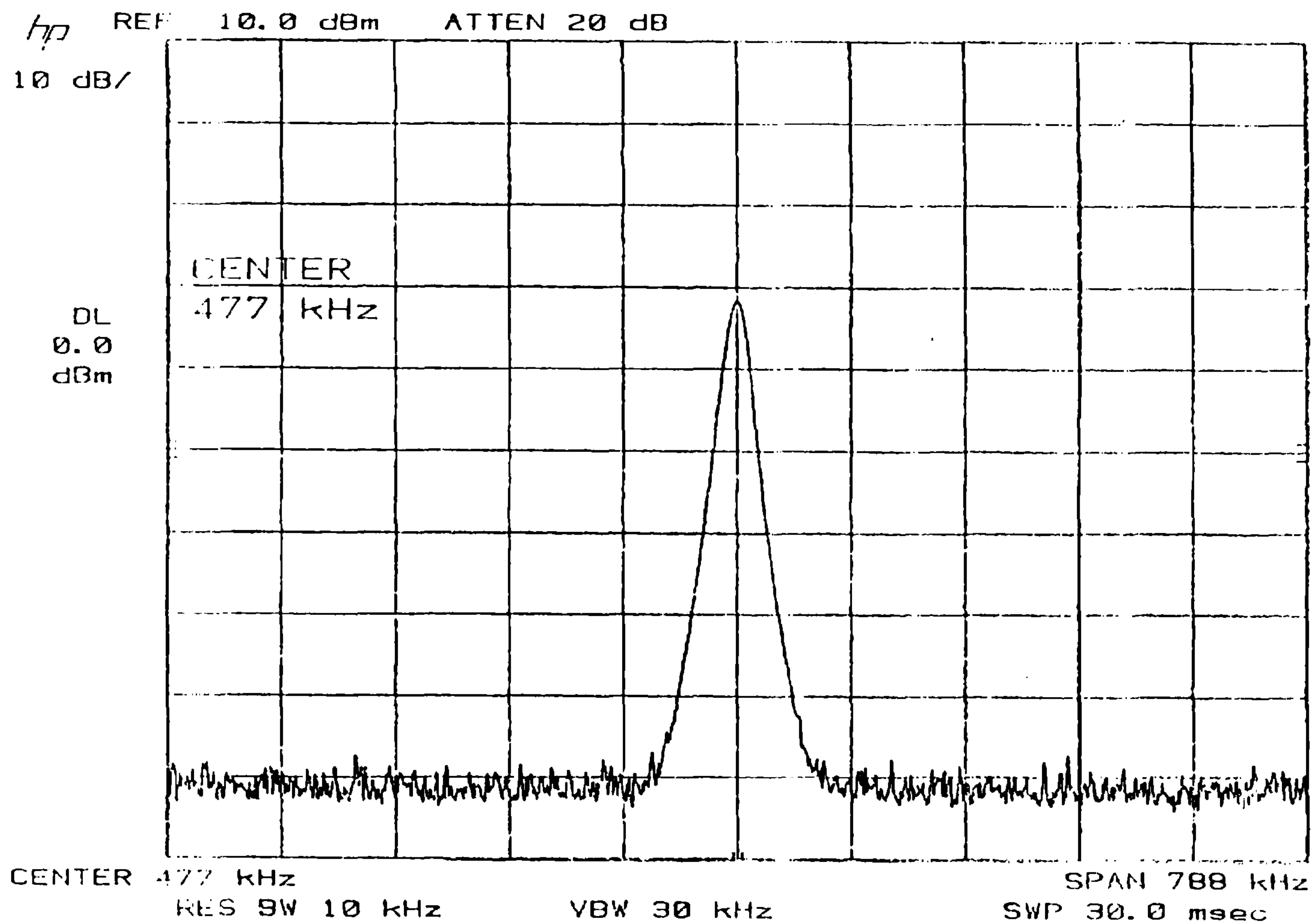


Figure 7. Carrier Spectral Analysis, Channel 3

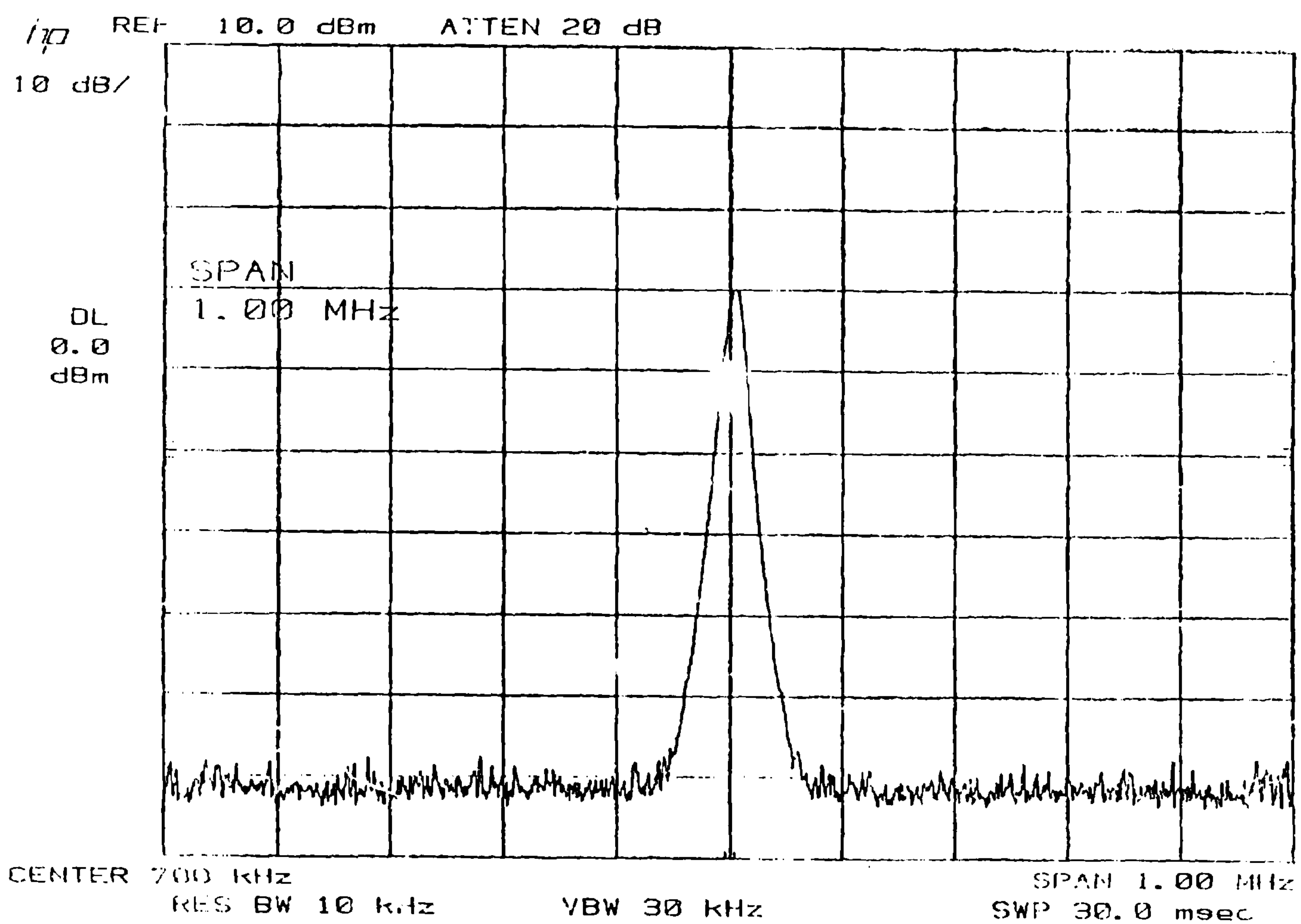


Figure 8. Carrier Spectral Analysis, Channel 4

2. Due to filter gains, the signals arriving at the summer are of different amplitudes and the various values of R1 to R4 are adjustments for this condition. The reader will also note that this summing "amplifier" is actually a summing attenuator. This is necessary because the optical receiver has an analog transmission range of only 0.5 volts. Therefore, the maximum amplitude of each of the four summed inputs is reduced to 0.1 volt to allow for periods of maximum coincidence. Failure to account for this limitation results in clipping of the transmitted waveform with the attendant formation of unpredictable harmonics which grossly interfere with the received signal.

D. OPTICAL TRANSMITTER

The optical transmitter chosen was a Hewlett-Packard HFBR-1402 analog capable LED device (Reference 1). The complete schematic is given in Figure 9 and Table 3.

Biasing range for the LED, as shown, is 2.2 to 2.7 volts, permitting the aforementioned 0.5 volt peak-to-peak analog waveform transmission. The lower level is actually a free choice commensurate with the transmission of suitable power at the minimum waveform level. The values listed give good performance over reasonable distances, (i.e., 100's of meters).

The driving circuitry itself is mostly concerned with the application of the biased signal to the transmitter. A voltage divider network, R4 and R5, is employed to provide the needed -2.2 volts (including inversion of the summer) to one input of a voltage summer while the FDM signal to be transmitted serves as the other input. A voltage follower is included for voltage isolation. Capacitor C1 serves to shunt any power supply transients to ground and resistor R8 provides current protection. The wiring of the actual HFBR-1402 is direct and in accordance with Reference 1.

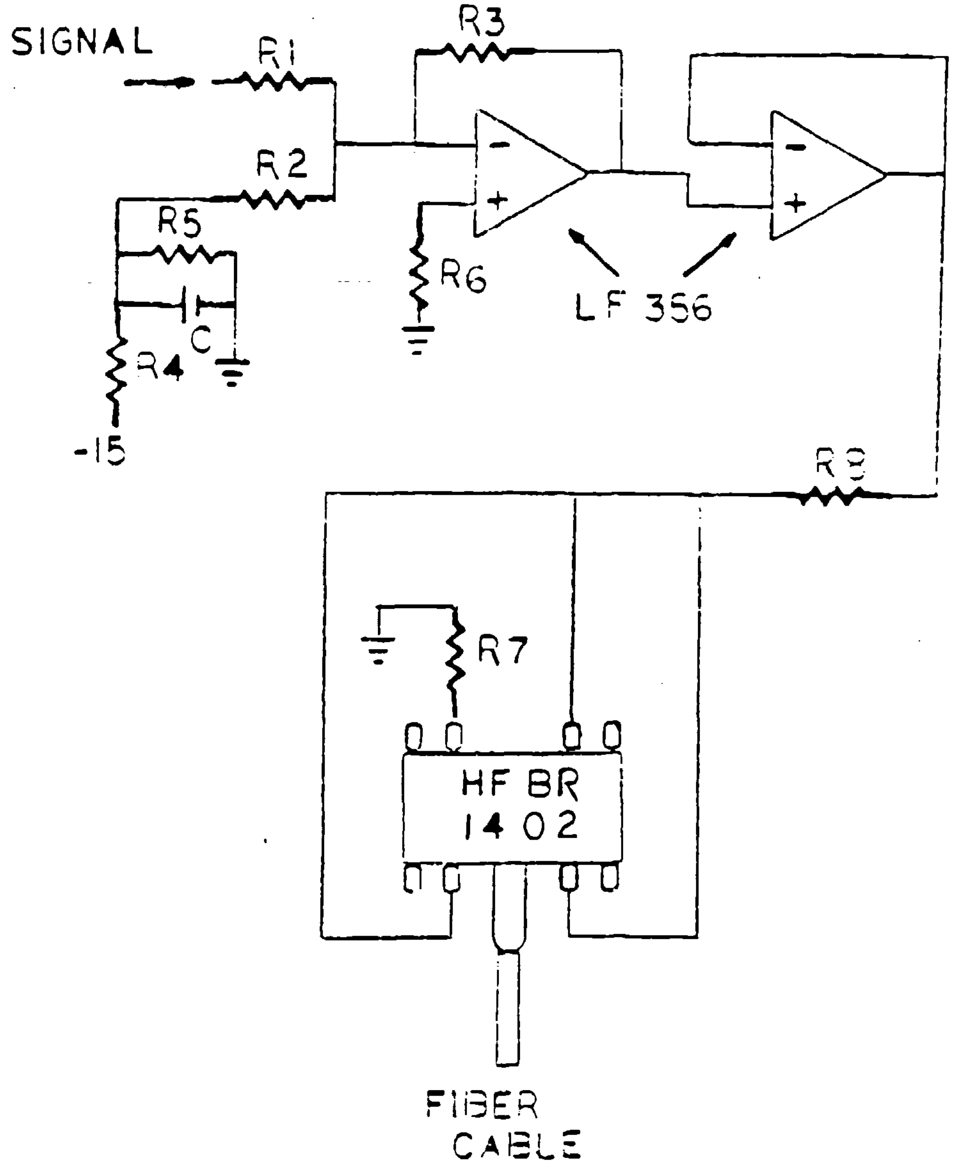


Figure 9. Optical Transmitter

TABLE 3

OPTICAL TRANSMITTER COMPONENT VALUES

R1 = R2 = R3 = R4 = 51.0k	C = 2000 pf
R5 = 11.0k	
R6 = 10.0k	
R7 = 51	
R8 = 10	

In summary, the transmitter consists of an FM modulator which accepts audio signals and applies FM modulation to a subcarrier sine wave, a bandpass filter which lowers harmonic distortion of the modulated wave, a voltage summer which accomplishes the multiplexing, and an optical transmitter which translates the electrical signal to an optical signal for insertion into the optic cable.

III. RECEIVER

The receiver group, detailed in Figure 10 consists of:

1) an optical receiver and associated power amplifiers,
2) a parallel arrangement of bandpass filters,
3) a phase-lock loop FM demodulator, and
4) low pass filters and a power amplifier.

It is appropriate to discuss each of these primary functional sub-systems of the receiver group in full detail.

A. OPTICAL RECEIVER AND ASSOCIATED POWER AMPLIFIERS

Low-cost optical receivers fall into two basic categories: those with an internal logic comparator/threshold device, suitable only for receipt of pulse transmissions, and those capable of full analog response. The nature of this system requires the latter. A review of readily available devices resulted in the selection of the HFBR-2404 by Hewlett Packard due to its high numerical aperture, simple drive circuit, and "breadboard" compatible mounting.

A schematic of the entire optical receiver subsystem is given in Figure 11 and Table 4. The reader will note that power is provided to the optical receiver from a 10 volt source via a voltage divider/follower combination. A variable voltage regulator would have served as well or better, however, the simple arrangement used is sufficient. The purpose of the supply is that the HFBR-2404 (max VCC of +7v) be provided with +5 volts from the same power supply as the power amplifiers, which require greater than +5 volts. This arrangement permits the use of any combination of supply voltages desired.

The actual receiver wiring is in accordance with data from Reference 1 for the 2402 device and operates satisfactorily for the 2404. (No explicit diagrams for the

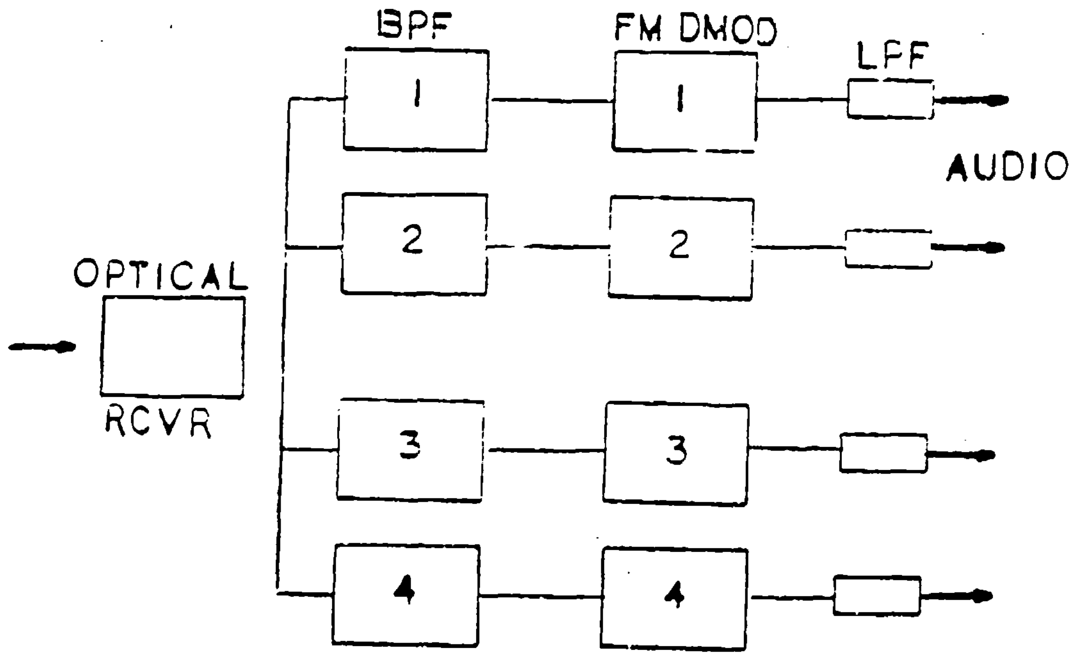


Figure 10. Receiver Group

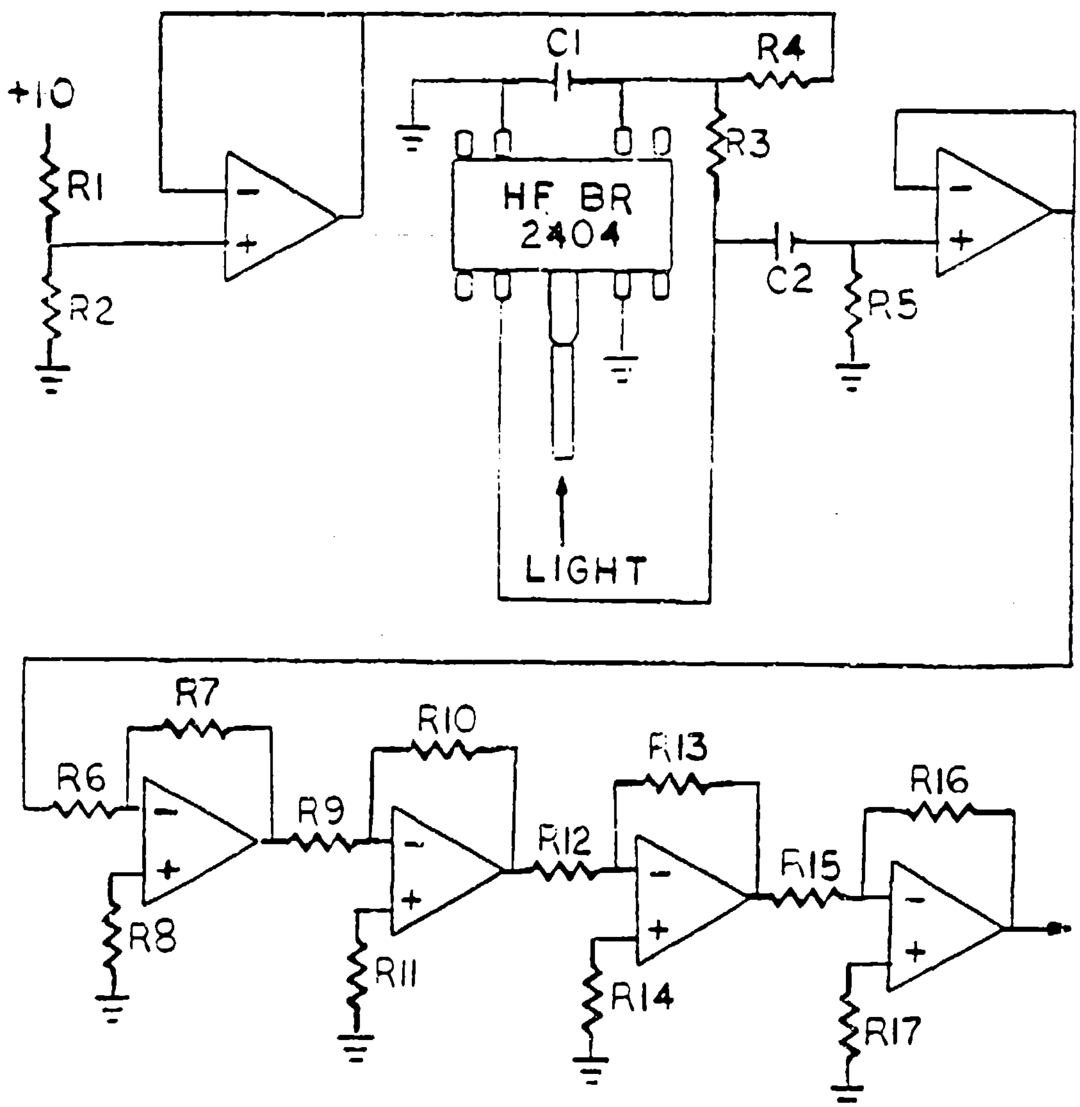


Figure 11. Optical Receiver

TABLE 4

OPTICAL RECEIVER COMPONENT VALUES

R1 = 51.0k	R10 = 2.0k	C1 = 100 uf
R2 = 51.0k	R11 = 500	C2 = 1500 pf
R3 = 510	R12 = 1.0k	
R4 = 300	R13 = 15.0k	
R5 = 100.0k	R14 = 500	
R6 = 1.0k	R15 = 1.0k	
R7 = 2.0k	R16 = 3.9k	
R8 = 500	R17 = 500	
R9 = 1.0k		

Note: All Op-Amps are LF 356

2404 could be located.) The only exception is the addition of R1 which provides current protection and also lowers the DC value of the output.

The final output signal level is on the order of millivolts and requires considerable power amplification before a usable voltage is obtained.

Design of the power amplifiers proved more difficult than first anticipated due primarily to the small signal input levels involved. The first step was the elimination of the dC component of the signal for voltage isolation. This was accomplished with the simple RC network of C2 and R5 and a standard voltage follower. Amplification was gradual with experimentation revealing the following optimum values: stage 1--2x max, stage 2--2x max, stage 3- -15x max), and stage 4--10x max (with 4x chosen). These values resulted in an overall amplification of 240 and yielded a clear output signal of 5 volts peak-to-peak.

It is essential that the input impedances of the amplifier stages be kept as low as possible to minimize additive amplifier noise. Not doing so caused tremendous noise problems in early work on this receiver.

B. RECEIVER BANDPASS FILTERS

The requirement for a parallel arrangement of bandpass filters to isolate the frequencies of interest (i.e., the modulated subcarrier) from the received composite signal is intrinsic in the nature of a Frequency Division Multiplexed communications system, AM or FM. This isolation must be as complete as possible to avoid "crosstalk": a situation in which frequencies of sufficient strength from adjacent channels overlap into the channel of interest and cause interference. Measurement of this "crosstalk" is one of the benchmarks of multiplexed systems and will be addressed fully in the System Performance Chapter.

The selection of the appropriate design and the determination of the optimum center frequencies for the four bandpass filters to minimize "crosstalk" occupied most of the effort required to complete this communication link.

Preliminary design of this communications system was quite straightforward and the requirement for four bandpass filters, of wide passband and +40 dBV attentuation between channels, appeared equally straightforward utilizing Active Filter techniques. There existed many "handbooks" and "cookbooks" as well as complete texts on the subject. The problem became one of arranging four FM modulated carriers separated in the frequency domain so as to avoid excessive "crosstalk" and yet not to extend to such high frequencies that Active Filters could not be used. (The exact extent of this frequency limitation was not made clear in the references, only alluded to.)

Adequate attentuation between channels had been experimentally determined earlier to be 40 dBV. A good rule of thumb for most active filter designs is 20 dBv per decade per second order stage (References 2 and 3). The logistics involved dictated that three stages were the largest filter that could be conveniently constructed. (Additional stages would have required an additional "breadboard".) These considerations allowed a 6th-order, 60 dBv/decade filter skirt. (This ability to cascade individual "stand alone" stages is one of the primary attributes of Active Filters.) It was hoped that these 60 dBV/decade filters could come very close to the 40 dBV between channels required if separation were kept to at least 150 kHz. This was later verified to be the case (Figures 22-25).

Given a broad passband, the filters available, and minimum 150 kHz separation, the four channels required a bandwidth of at least 450 kHz. With the channel spacing and filter parameters determined, construction of the filters began. At this point a very unpleasent surprise arose in that none of the designs in the available literature functioned at all above 250 kHz. This fact was ascertained experimentally at great effort as the literature itself (References 2-5) made little to no mention of the frequency limits of the designs. State Variable, Biquad, Positive Feedback, Sallen-Key, Multiple Feedback, and several other filter designs with no particular title were tried with no success. The completion of this part of the endeavor is due in no small part to Professor Sherif Michael who provided a design known as the Generalized Immittance Converter GIC (Reference 6). Professor Michael had much experience with this design and the version utilized is his design with one minor variation provided by the author.

The next step was the center frequency selection. It was found that up to the 5th harmonic had to be considered when selecting channel spacing. The problem of harmonic avoidance therefore, grew "exponentially" with each additional channel, so much so in fact, that the final channel selection was achieved by the less than aesthetically pleasing method of sweeping a FM modulated carrier through the frequency spectrum from 0 to 800 kHz and choosing the frequency position of minimum interference effect on the remaining 3 channels.

Figure 12 and Table 5 provide complete schematics of the 6th order, 3 stage, modified GIC bandpass filter which was ultimately used. (Note the voltage isolators on both ends.) The arrangement is very direct and very simple to utilize. A generic single GIC stage is illustrated in Figure 13. The design equations of this particular band-pass filter are extraordinarily simple. The center frequency is given by

$$f_c = 1/2\pi RC \qquad (7)$$

where

$$R = R1 = R2 = R3 = R4 \qquad (8)$$

and

$$C = C1 = C2 \qquad (9)$$

The Q and gain of this filter are equally straightforward with

$$Q = R5/R \qquad (10)$$

and

$$\text{Gain} = 2. \qquad (11)$$

The reader will note the slight difference between the generic GIC of Figure 13 and the diagram of the actual filter used, Figure 12 (i.e., the addition of one extra resistor per stage, particularly resistors R6,R12, and R18).

It was experimentally discovered that the standard GIC design provided by Professor Michael performed well at frequencies approaching 800 kHz. A reduction of gain and timing component shifts of approximately 30% were the only effects of high frequency operation, save one, with both of these effects being easily compensated for. However, the one effect which did occur and was not tolerable, was the tendency for the filter to go into a "lockup" mode when powering up. To make matters worse, the phenomenon appeared randomly. After experimentation with various values of R and C failed to resolve the difficulty, it was discovered that the filter could be cleared of its lockup condition by momentarily shorting the + terminal of the second operational amplifier in the stage to ground. The author then realized that a path to ground at this point would provide stable operation at very high frequency. The exact mechanism of this effect is beyond the scope of this thesis but could be the subject of future study.

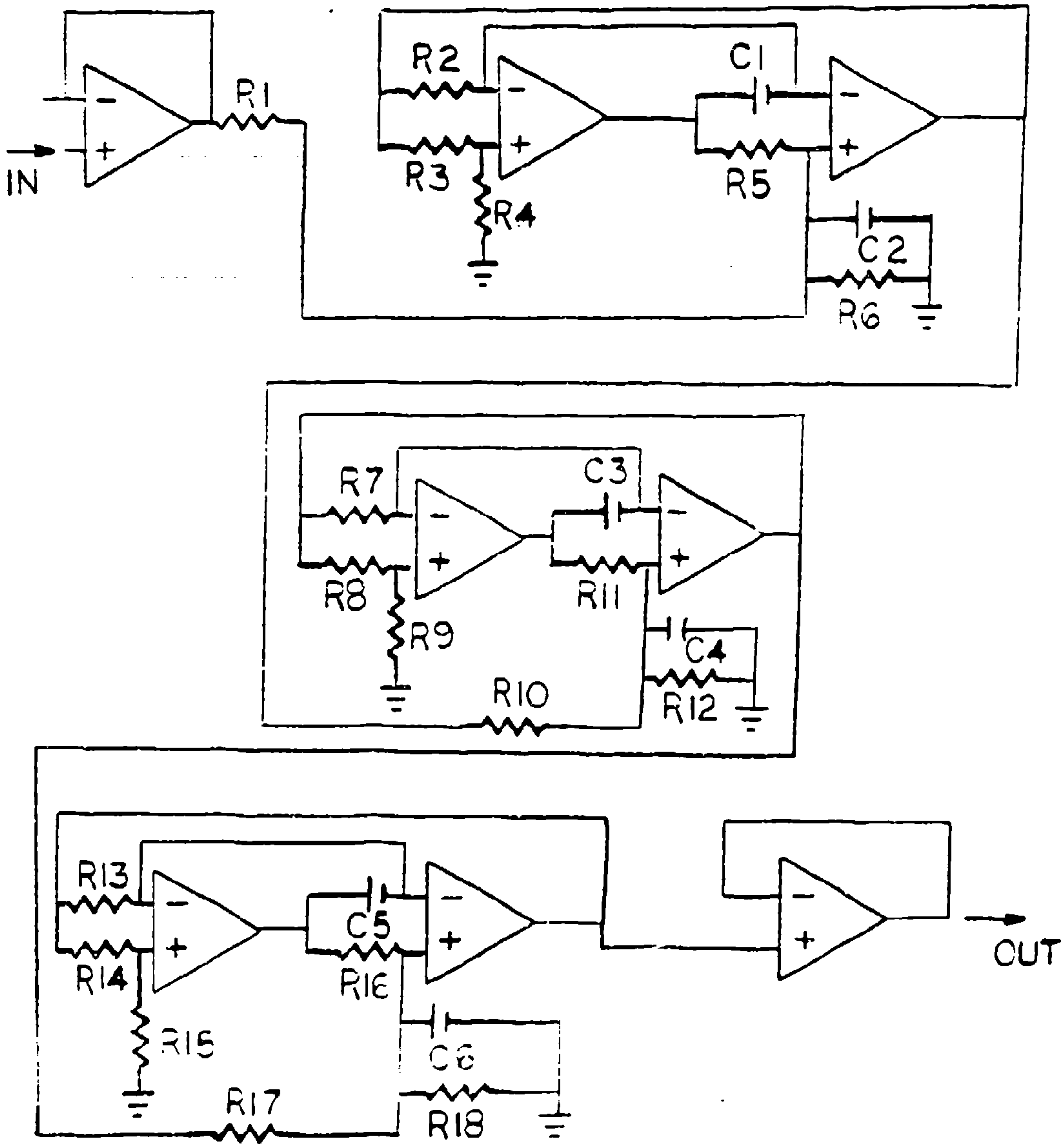


Figure 12. Receiver Bandpass Filter

TABLE 5

RECEIVER BANDPASS FILTER COMPONENT VALUES

Channel 1

Stage 1
Q Resistor R1 = 20.0k
R2 = R3 = R4 = R5 = 2.4k
Stability Resistor R6 = 50.0k
C1 = C2 = 850 pf

Stage 2
Q Resistor R10 = 20.0k
R7 = R8 = R9 = R11 = 2.4k
Stability Resistor R12 = 51.0k
C3 = C4 = 680 pf

Stage 3
Q Resistor R17 = 10.0k
R13 = R14 = R15 = R16 = 2.4k
Stability Resistor R18 = 50.0k
C5 = C6 = 750 pf

Channel 2

Stage 1
Q Resistor R1 = 15.0k
R2 = R3 = R4 = R5 = 1.8k
Stability Resistor R6 = 20.0k
C1 = C2 = 220 pf

Stage 2
Q Resistor R10 = 15.0k
R7 = R8 = R9 = R11 = 1.8k
Stability Resistor R12 = 20.0k
C3 = C4 = 220 pf

Stage 3
Q Resistor R17 = 9.1k
R13 = R14 = R15 = R16 = 1.8k
Stability Resistor R18 = 20.0k
C5 = C6 = 220 pf

Channel 3

Stage 1
Q Resistor R1 = 13.0k
R2 = R3 = R4 = R5 = 820
Stability Resistor R6 = 20.0k
C1 = C2 = 370 pf

Stage 2
Q Resistor R10 = 13.0k
R7 = R8 = R9 = R11 = 820
Stability Resistor R12 = 5.1k
C3 = C4 = 288 pf

Stage 3
Q Resistor R17 = 8.2k
R13 = R14 = R15 = R16 = 820
Stability Resistor R18 = 20.0k
C5 = C6 = 320 pf

Channel 4

Stage 1
Q Resistor R1 = 20.0k
R2 = R3 = R4 = R5 = 1.0k
Stability Resistor R6 = 51.0k
C1 = C2 = 182 pf

Stage 2
Q Resistor R10 = 20.0k
R7 = R8 = R9 = R11 = 1.0k
Stability Resistor R12 = 51.0k
C3 = C4 = 168 pf

Stage 3
Q Resistor R17 = 10.0k
R13 = R14 = R15 = R16 = 1.0k
Stability Resistor R18 = 51.0k
C5 = C6 = 168 pf

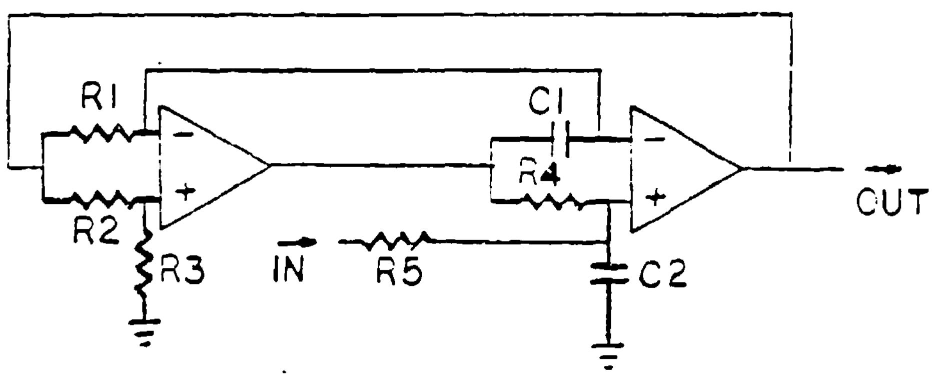


Figure 13. Generic GIC Bandpass Filter

In summary, an experimentally determined resistance of 20 kohms placed in the indicated position of Figure 12, prevented filter lockup and permitted channels 3 and 4 to be operated at 475 kHz and 700 kHz respectively. (Note: Resistances of 20 kohms or higher do not markedly affect either the center frequency or the gain characteristics of the filter although a reduction in Q appears to occur.) The filter can be forced into lockup by applying a momentary + supply voltage to the + terminal of the second Op-Amp. A resistance of 5.1 kohms was found to allow auto-recovery from this condition but at the cost of reduced gain.

The performance characteristics of the filters are felt to be very good with high phase linearity and gentle sloping peak gain curves. It was found that the Phase Lock Loop FM receiver was insensitive to AM thereby permitting the use of filters with gently sloping peaks which were far easier to construct than ones with sharp skirts.

The Q of these filters was chosen in accordance with Carson's Rule for estimating the bandwidth and the results were generally good. The stages were, of course, staggered somewhat to provide the width and slope desired. Reference 3 provides a thorough discussion of this staggering technique.

Included as Figures 14 to 21 are the gain and phase characteristic plots as taken from a HP 3575A Gain Phase meter fed by a swept FM signal and plotted on a standard X-Y plotter. A discussion of each set, gain and phase, is presented below.

Channel 1 (Figure 14 and 15) with f_c=90 kHz has a gain of almost 1 to 1 and a 3 dB passband of 54 kHz. The phase is linear throughout the passband with zero phase shift occurring at 94 kHz. The "wraparound" effect apparent in the phase diagrams was the result of the recorder resetting in response to the Gain/Phase Meter's output undergoing a sign change at 180 degrees.

Channel 2 (Figures 16 and 17) with f_c=325 kHz has a gain of 4.5 dB and a passband of 51 kHz. The phase is once again very linear in the passband with zero phase shift at 325 kHz.

Channel 3 (Figures 18 and 19) has a high center frequency of 477 kHz. As noted before, the gain of these filters tends toward attenuation at high frequencies and this is evident with this filter's -7.8 dB gain at f_c. All other characteristics are desirable with a passband of 90 kHz and linear phase. Zero phase shift occurred at 479 kHz.

Channel 4 (Figures 20 and 21) has a higher center frequency of 700kHz. This channel suffers even more attenuation (-10 dB) at f_c but again maintains good shape and phase linearity. The passband is 81 kHz with zero phase shift occurring at 700 kHz.

For a final check on the performance of the filters in parallel, the system was brought up to full operation with all four channels transmitting. The filter to be examined then had its carrier removed. Figures 22 to 25 illustrate the spectrum as seen by the receivers through their respective filters in this condition. Notice that the largest interference component in any of them is -36.5 dBm for channel 3 (Figure 24) with the others well into the -40 dBm range. ("Marker zoom" indicates the level of these components.) Although not a direct measure of crosstalk immunity, it does dramatically highlight the filters' isolation capabilities, especially concerning the 40 dBv attenuation between channels requirement mentioned earlier.

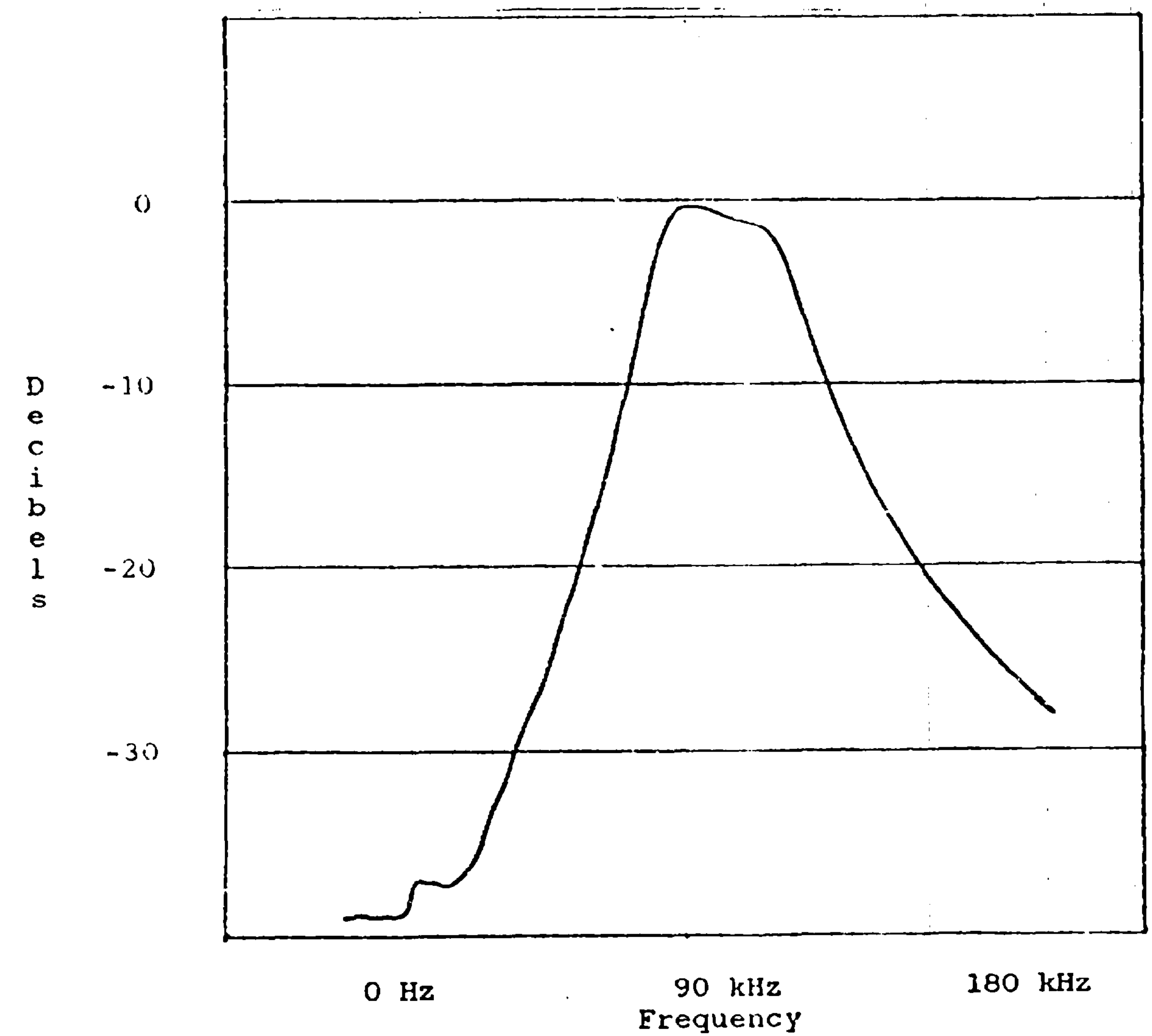


Figure 14. Gain Characteristic, Channel 1

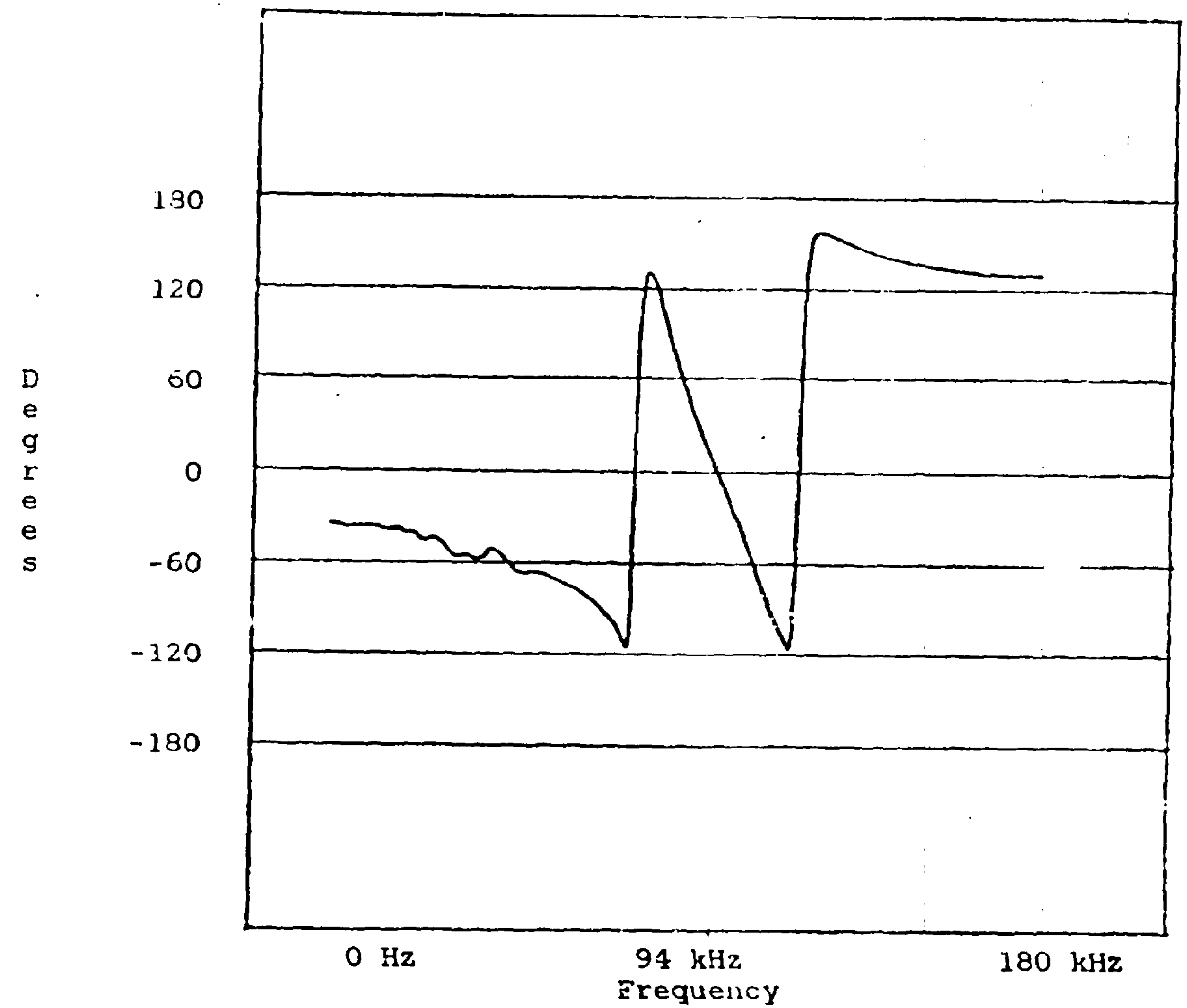


Figure 15. Phase characteristic, channel 1

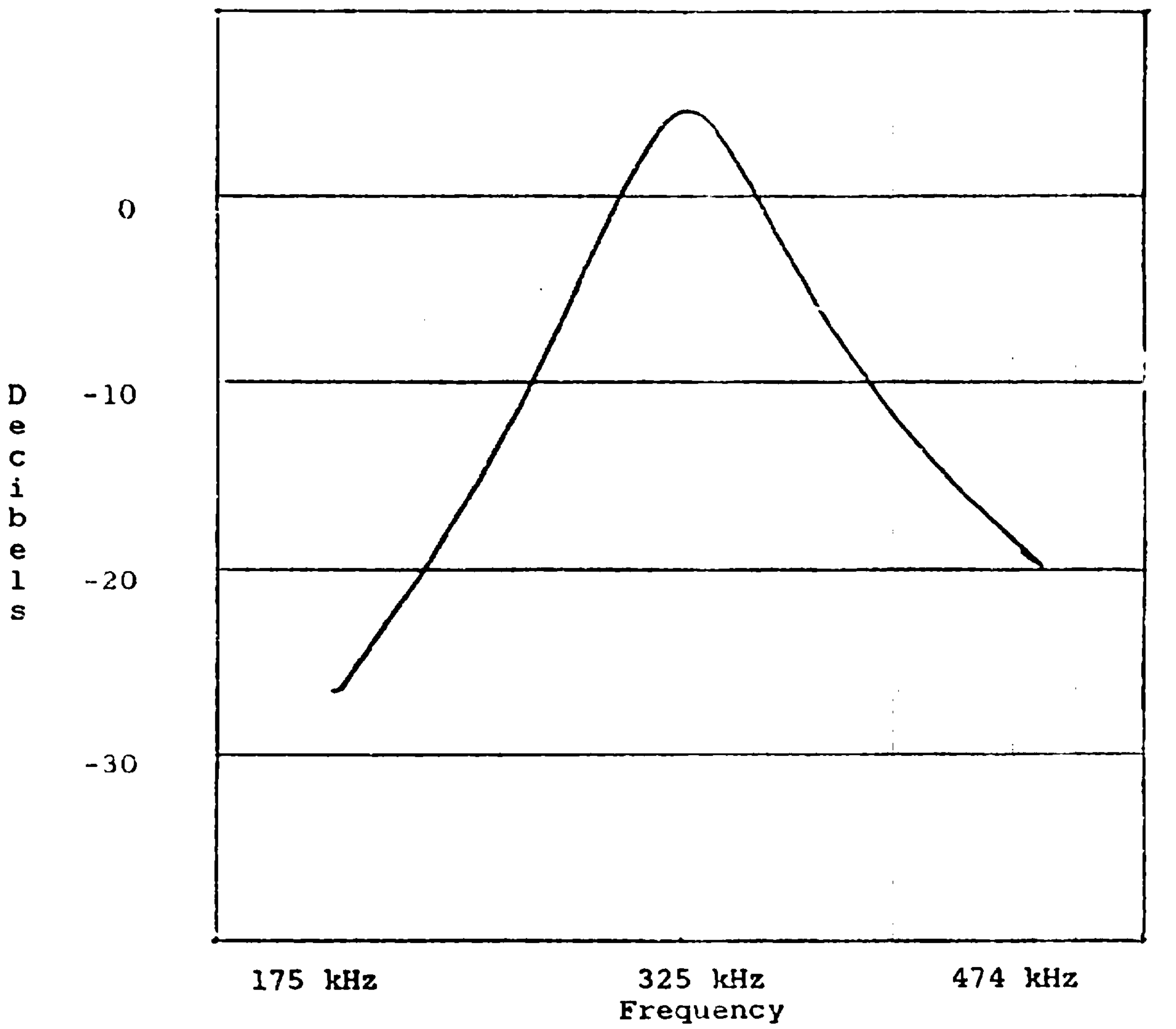


Figure 16. Gain Characteristic, Channel 2

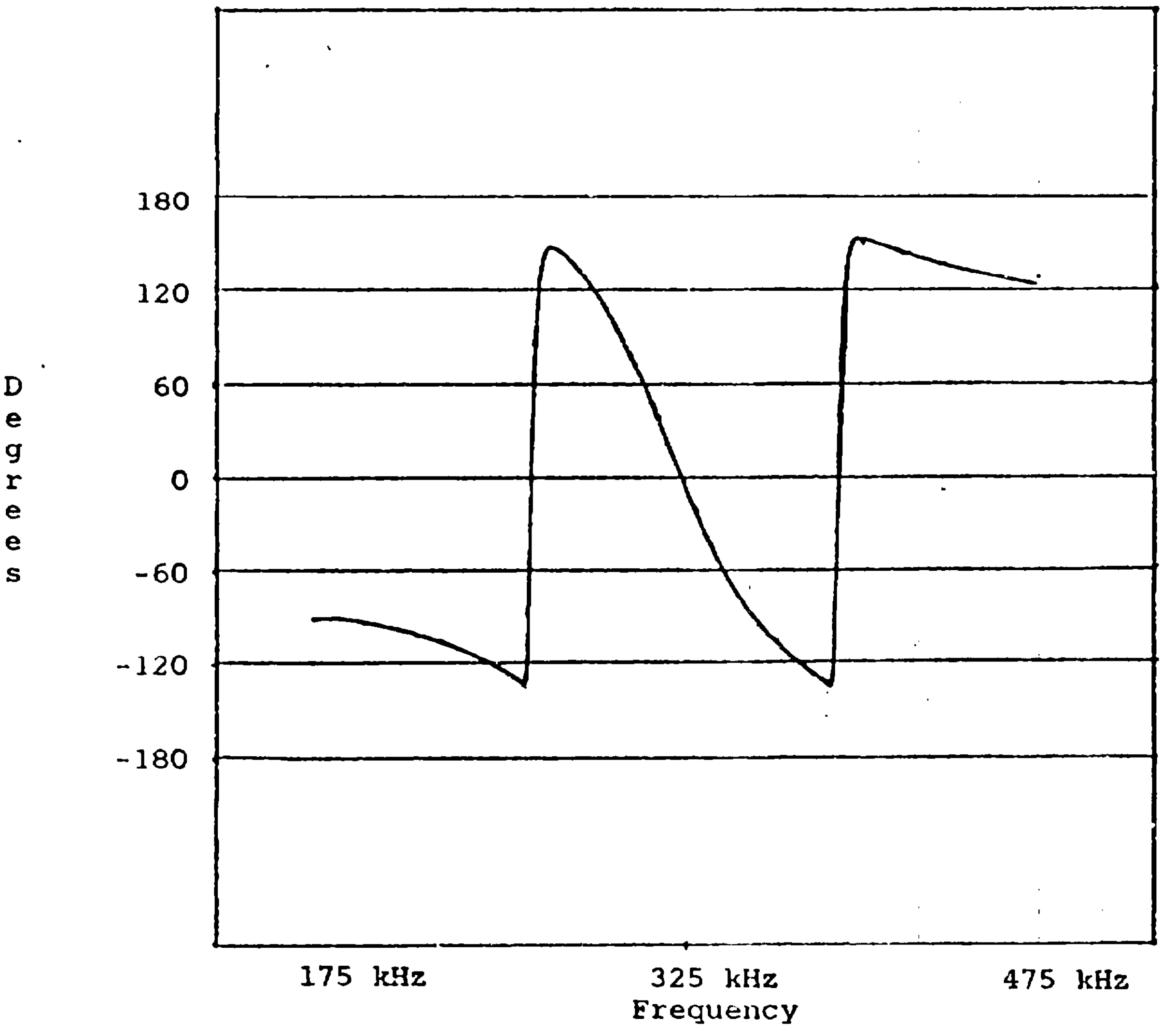


Figure 17. Phase characteristic, channel 2

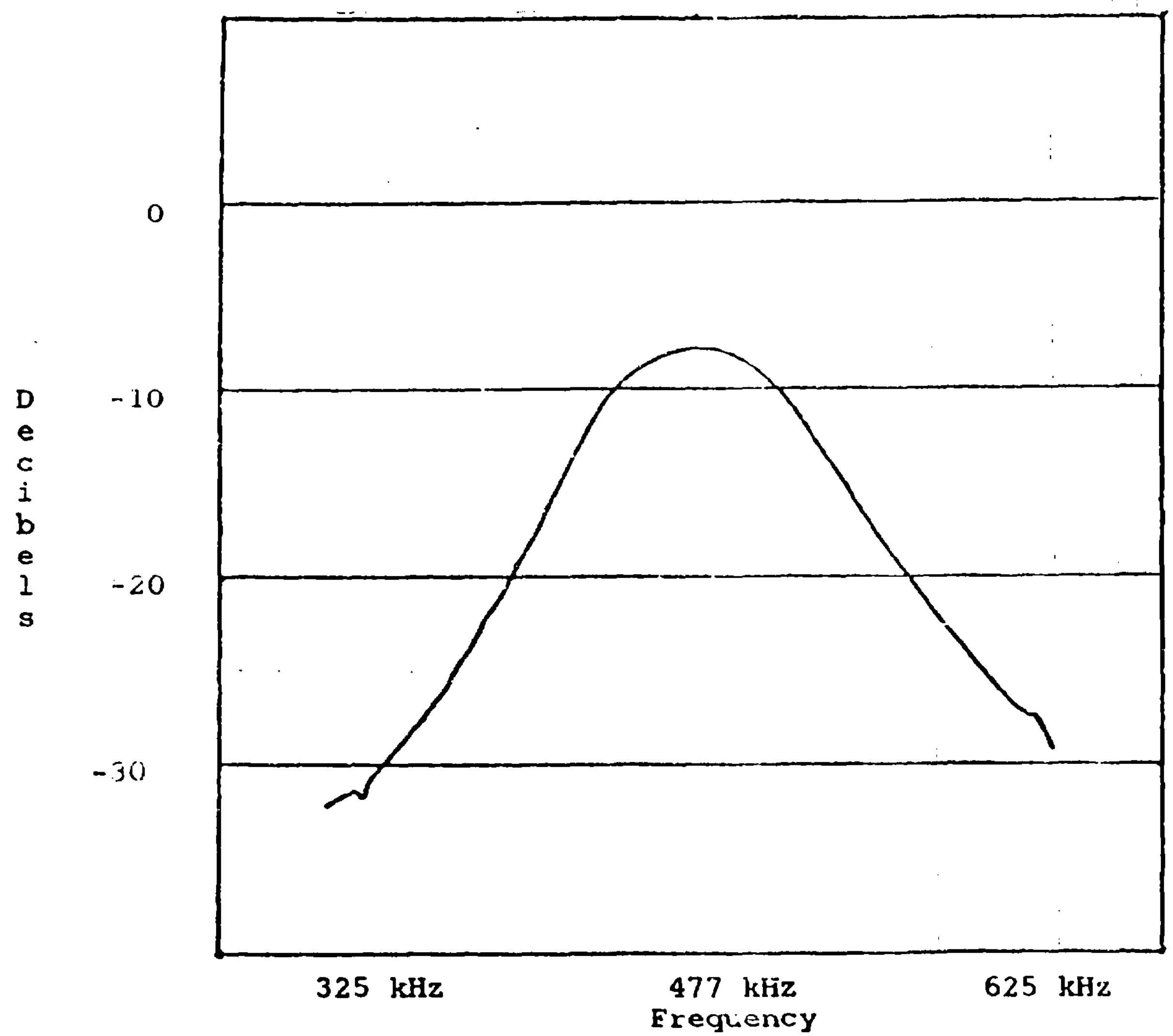


Figure 18. Gain Characteristic, Channel 3

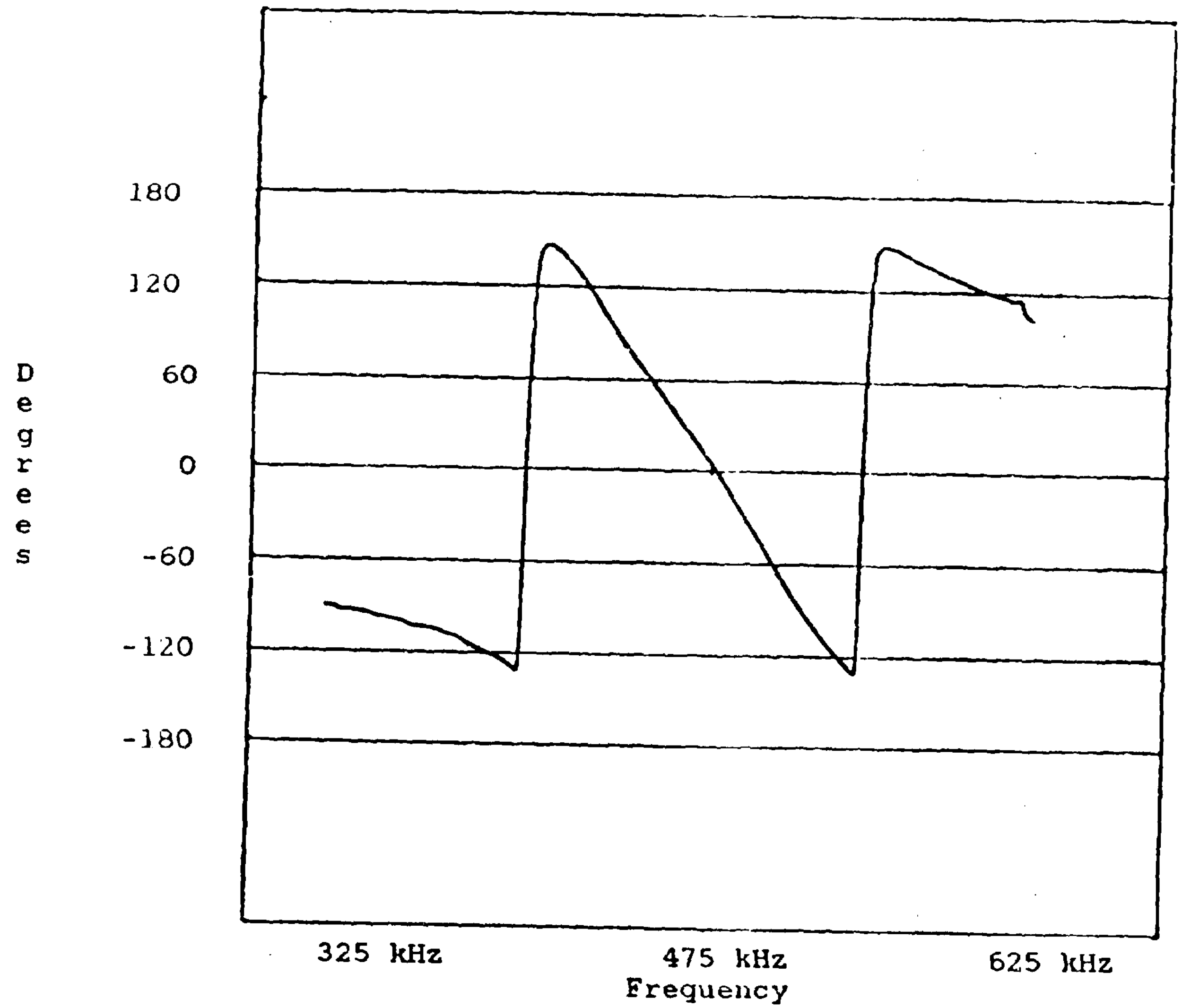


Figure 19. Phase characteristic, channel 3

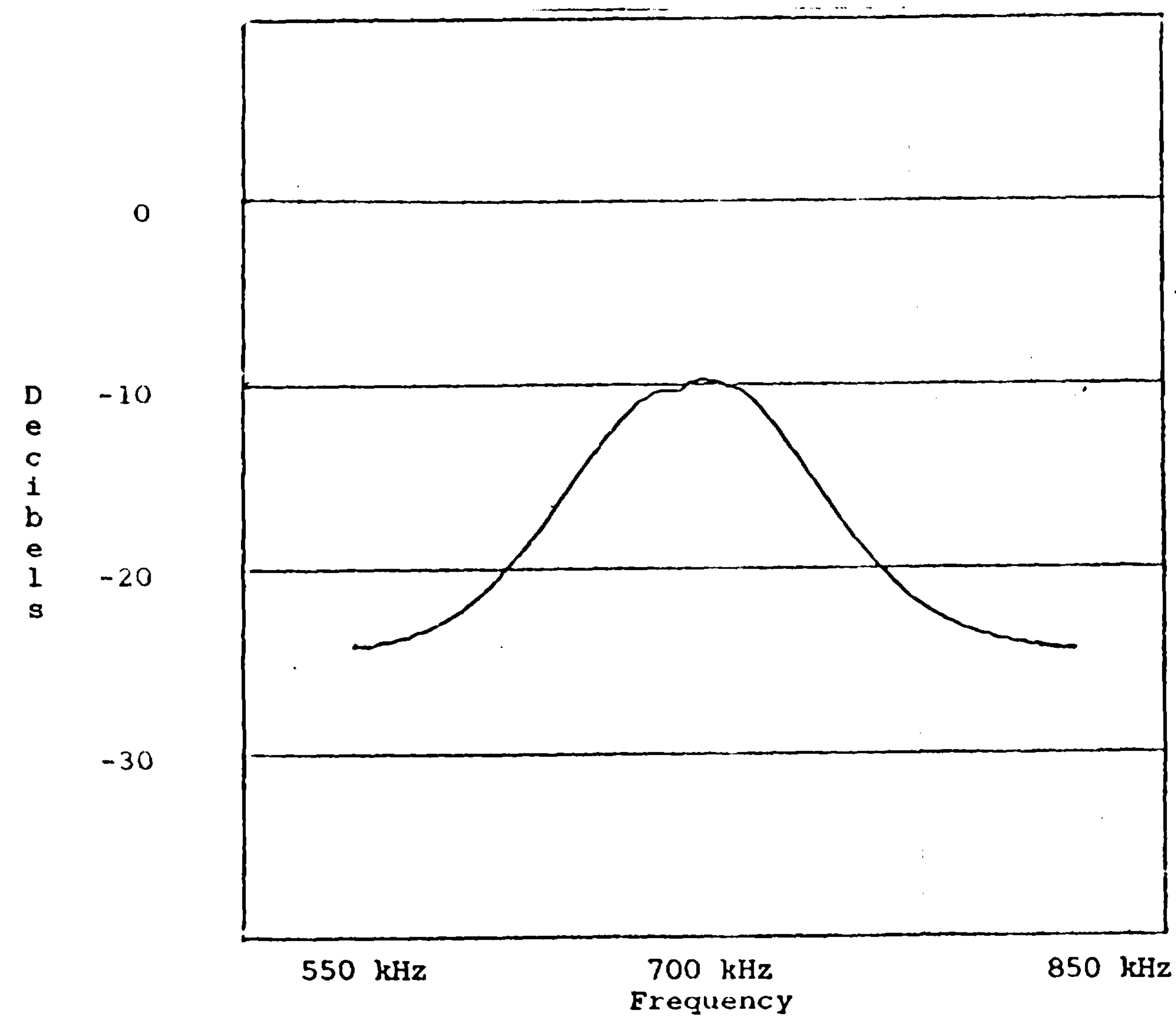


Figure 20. Gain Characteristic, Channel 4

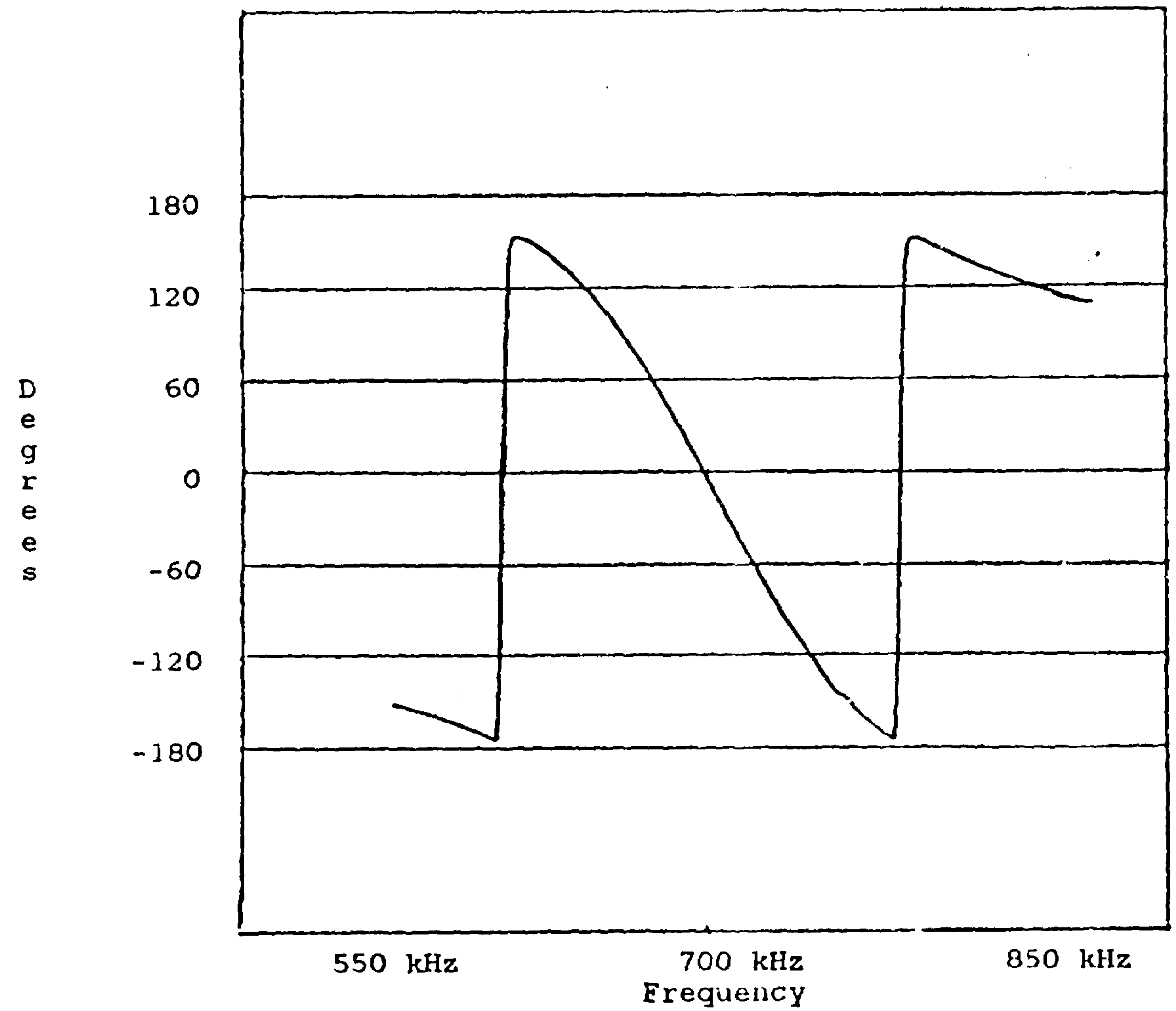


Figure 21. Phase characteristic, channel 4

C. PHASE LOCK LOOP FM DEMODULATOR

The task of FM demodulation was greatly simplified by the use of the Signetics NE/SE 565 Phase-Lock Loop chip. Reference 7 provided the necessary schematic, Figure 26 and Table 6, which was used with the exception of the addition of R1 to reduce signal strength. This was necessary due to the exceptional sensitivity of this device.

The free running frequency of the demodulator is

$$f_o = 12/(4 \times R4 \times C2) \qquad (12)$$

while the Lock Range is

$$f_{lc} = 8 \times (f_o/V_{cc}) \qquad (13)$$

with

$$V_{cc} = +/- \text{ supply voltage} \qquad (14)$$

The Capture Range is

$$f_{cp} = .159 \times \sqrt{2\pi f_{lc}/3600 \times C3} \qquad (15)$$

D. RECEIVER LOW PASS FILTER

The output of the FM demodulator is adulterated with high frequency components so precision lowpass filtering is required. The 3-stage, 6th-order lowpass, filter with f_c=20 kHz (Figure 27 and Table 7) was employed with excellent results. The design chosen was a Sallen and Key lowpass filter from Reference 2.

This design performed flawlessly in accordance with the source equations. The component value selection process, while not difficult, is more complicated than that of the GIC bandpass filter and is not included for that reason.

A word of caution is necessary concerning the input voltage follower isolation section of this filter. For unknown reasons, an additional resistance of 1 megohm, R1, was required to achieve full isolation. Poor performance resulted before this isolator was included. (Note: Although not part of the actual filter, the power amplifier section is included for completeness.)

In summary, the receiver consists of an optical receiver to translate optical signals to electrical signals, a bank of bandpass filters to demultiplex the resulting composite signal, a FM demodulator to recover the information signal, and a lowpass filter to remove high frequency noise from the recovered signal.

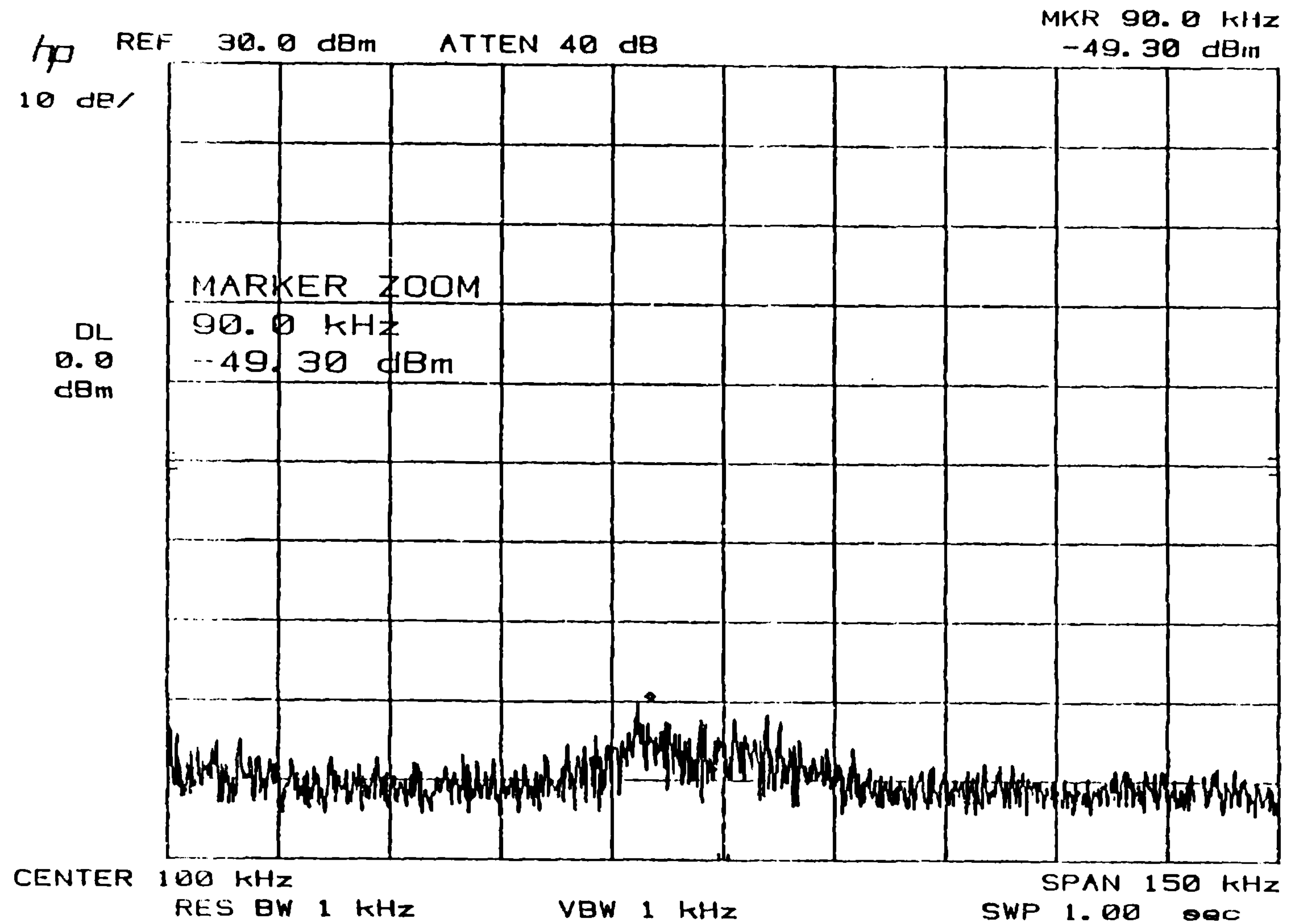


Figure 22. Spectral "window look", Channel 1

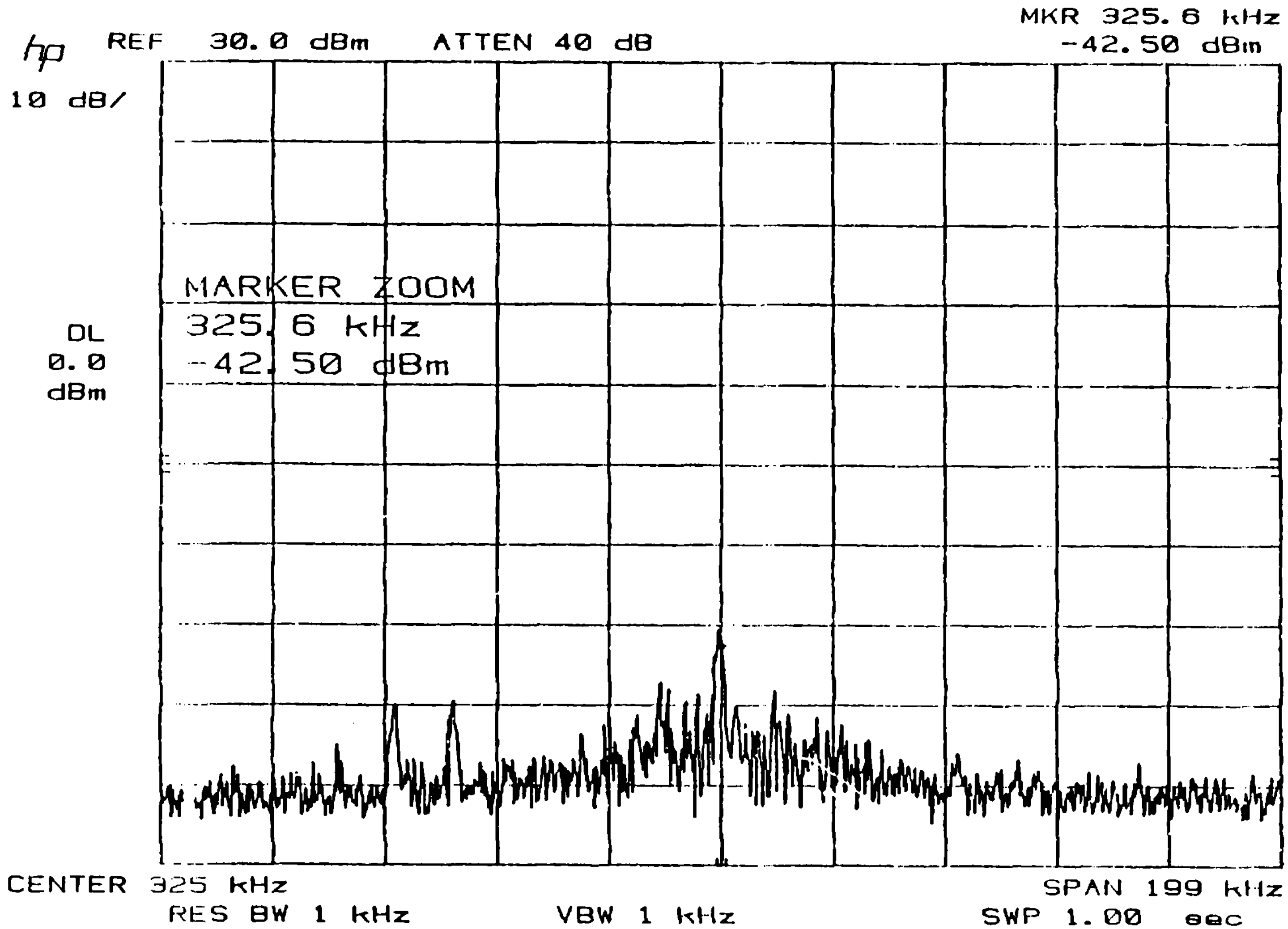


Figure 23. Spectral "window look", Channel 2

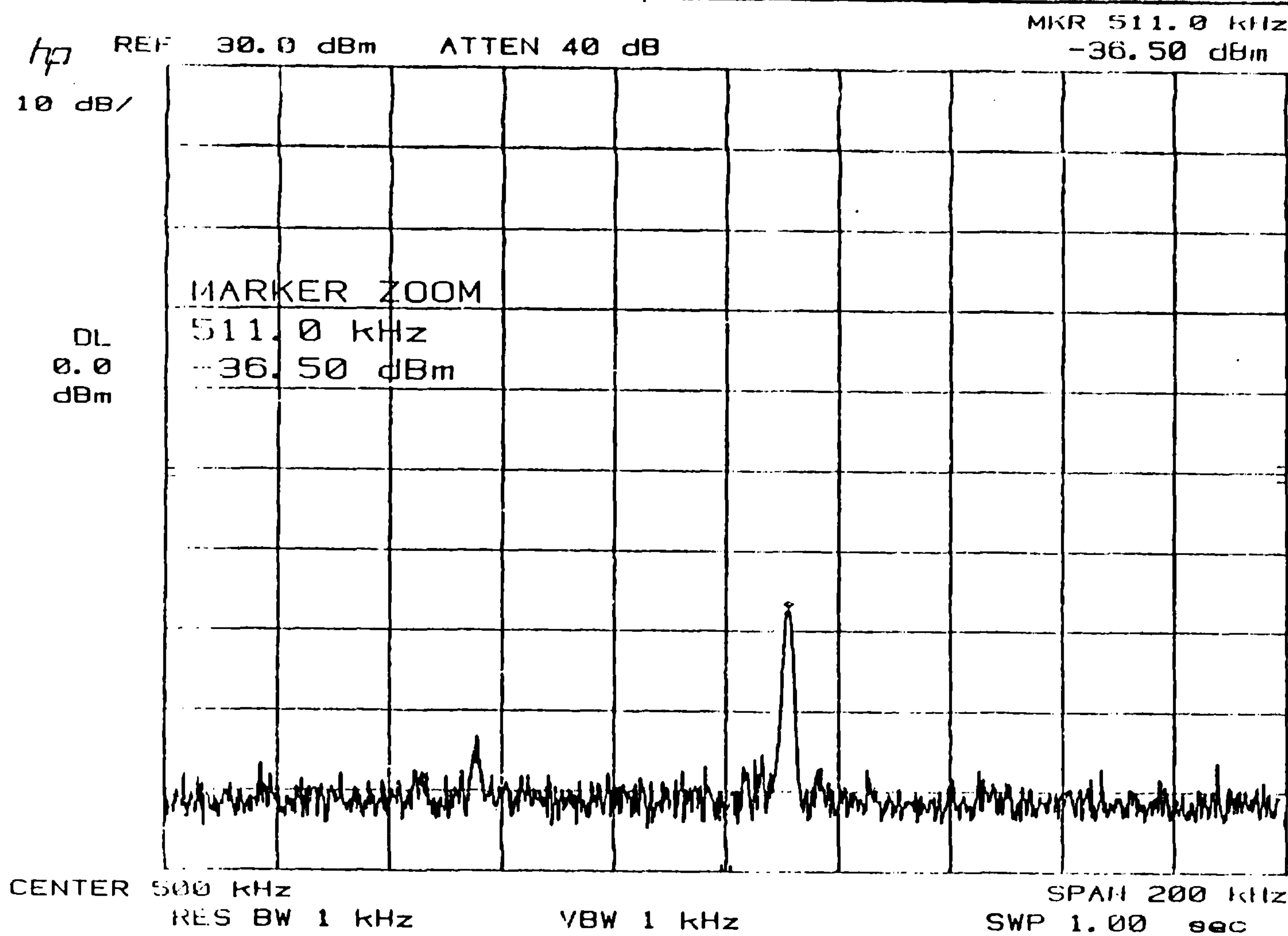


Figure 24. Spectral "window look", Channel 3

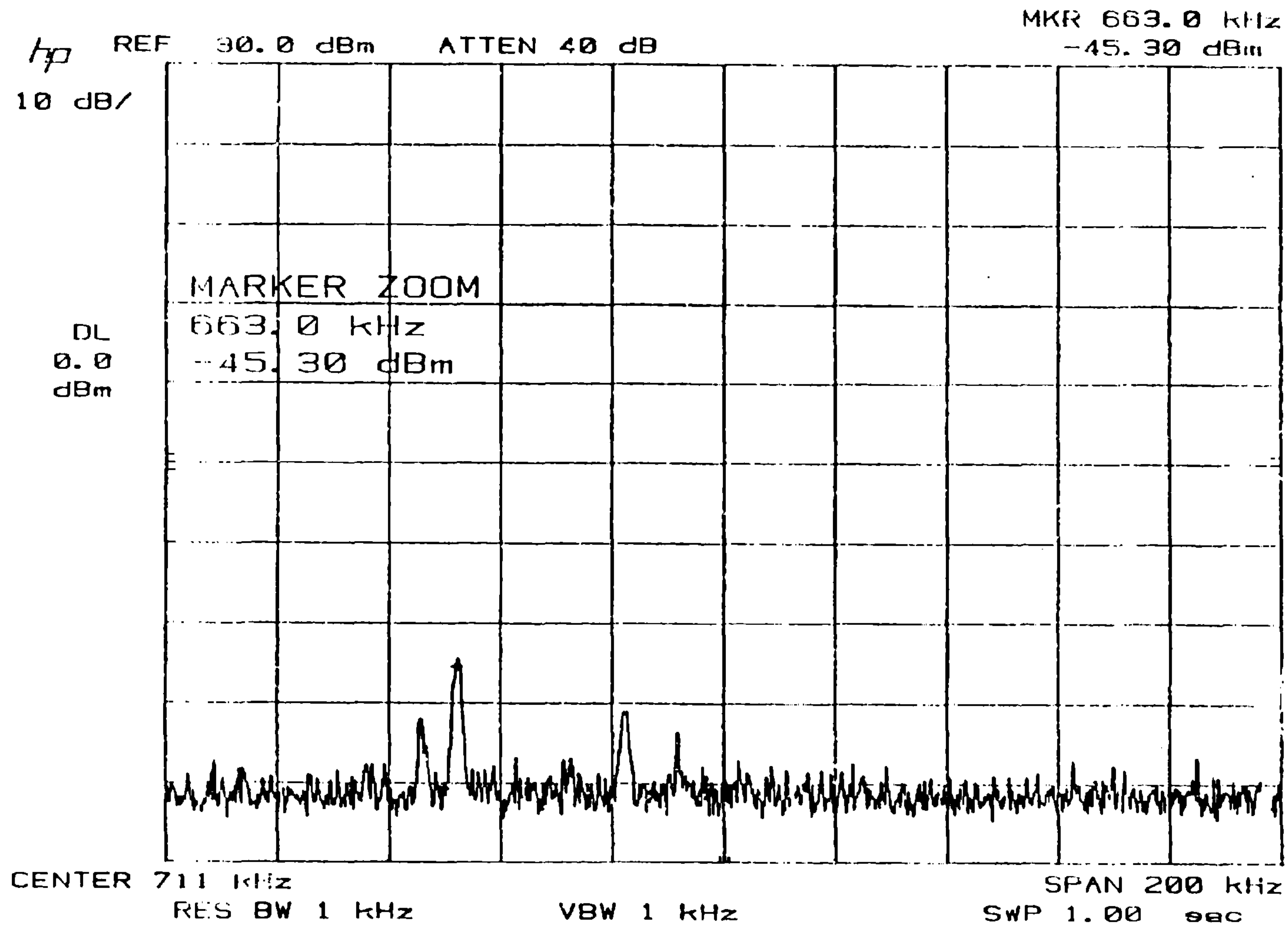


Figure 25. Spectral "window look", Channel 4

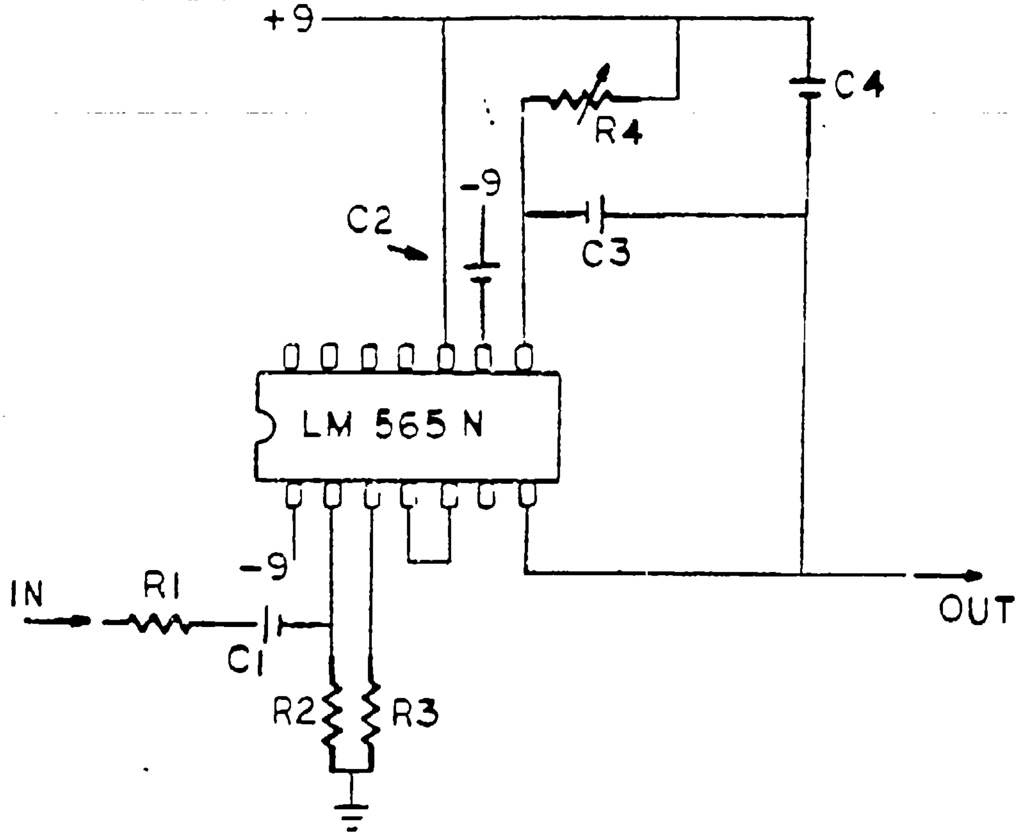


Figure 26. Phase Lock Loop FM Demodulator

TABLE 6

FM RECEIVER COMPONENT VALUES

Channel 1

R1 = 20.0k	C1 = 10 nf
R2 = 1.0k	C2 = 1 nf
R3 = 1.0k	C3 = 1 nf
R4 = 0-5k	C4 = 100 pf

Channel 2

R1 = 20.0k	C1 = 10 nf
R2 = 1.0k	C2 = 680 pf
R3 = 1.0k	C3 = 1 nf
R4 = 0-5k	C4 = 100 pf

Channel 3

R1 = 20.0k	C1 = 10 nf
R2 = 1.0k	C2 = 560 pf
R3 = 1.0k	C3 = 1 nf
R4 = 0-5k	C4 = 100 pf

Channel 4

R1 = 20.0k	C1 = 10 nf
R2 = 1.0k	C2 = 370 pf
R3 = 1.0k	C3 = 1 nf
R4 = 0-5k	C4 = 100 pf

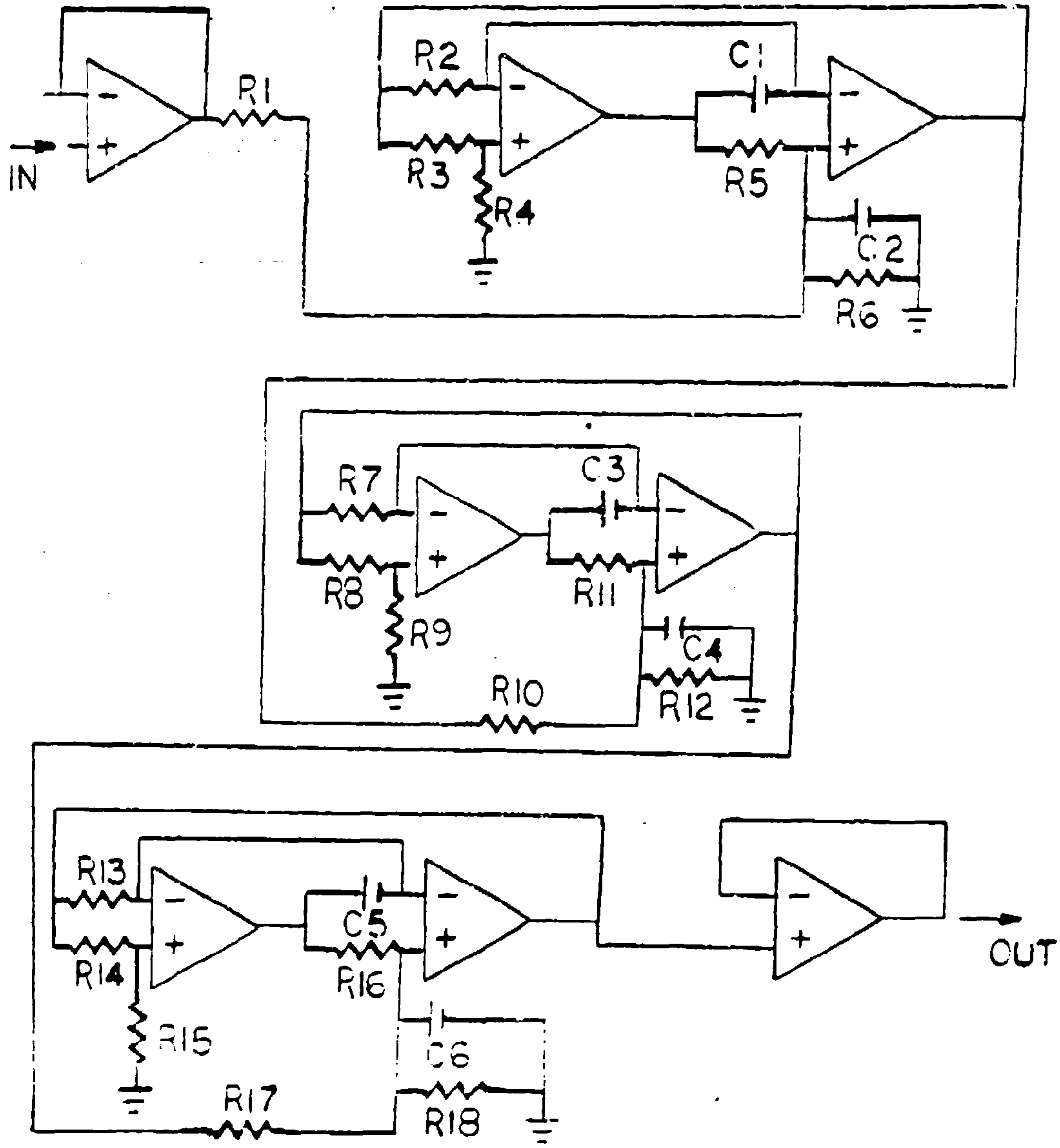


Figure 27. Receiver Lowpass Filter and Power Amplifier

TABLE 7

RECEIVER LOWPASS FILTER AND POWER AMPLIFIER
COMPONENT VALUES

Note: All channels use the same values

R1 = 1 Meg	R7 = 1.0k	C1 = 10 nf
R2 = 680	R8 = 160.0k	C2 = 5 nf
R3 = 680	R9 = 51.0k	C3 = 0.022 uf
R4 = 510	R10 = 300.0k	C4 = 0.0025 uf
R5 = 1.0k	R11 = 20.0k	C5 = 22 nf
R6 = 1.0k		C6 = 10 nf
		C7=1 uf

Note: All Op-Amps are 741 CN

IV. SYSTEM PERFORMANCE

A. CROSSTALK

The "crosstalk" measurement was achieved by the spectral analysis of a 5 kHz test signal transmitted on the channel of interest while adjacent channels were carrying a 3 kHz test signal. Figures 28-31 show the results. Notice that, to the limit of the spectrum analyzer's capability, (-50 dB), there were no observed frequency components of the 3 kHz signal present. Figure 32 is the spectrum of the 3 kHz signal for comparison. This absence of interfering components is the definition of a "crosstalk free" channel.

B. HARMONIC DISTORTION

Spectral analysis of a transmitted sinewave information signal of 9 kHz (Figures 33-36) shows the maximum harmonic distortion of any channel to be -45 dB. This is considered adequate against a commercial standard of -50 dB. The marker arrows on Figure 33-36 indicate the position of these harmonics.

C. GAIN

A flat gain versus frequency response curve for end-to-end transmission is ideal. That goal was essentially reached with channels 2 and 4 (Figures 38 and 40) while channels 1 and 3 (Figures 37 and 39) performed less spectacularly exhibiting reduced gain and fidelity above 16 kHz. The reason for this is simple; the solution is not.

The bandwidth of the transmitted signal increases as the frequency of the information signal increases. In accordance with Carson's Rule an estimation of FM bandwidth is given by

$$FM_{BW} = 2 \times FM_R \times BW_I \qquad (16)$$

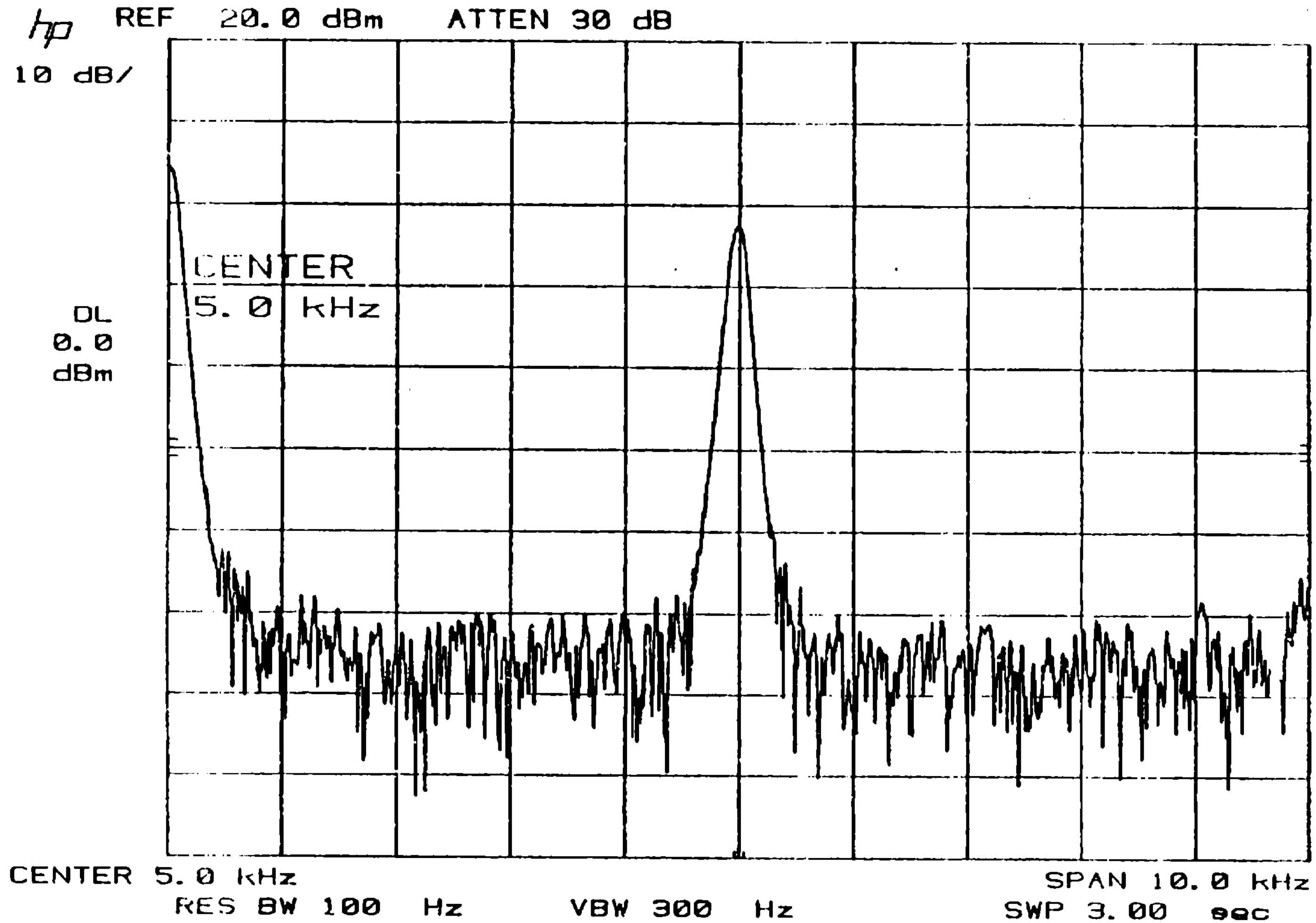


Figure 28. "Crosstalk" Spectral Analysis, Channel 1

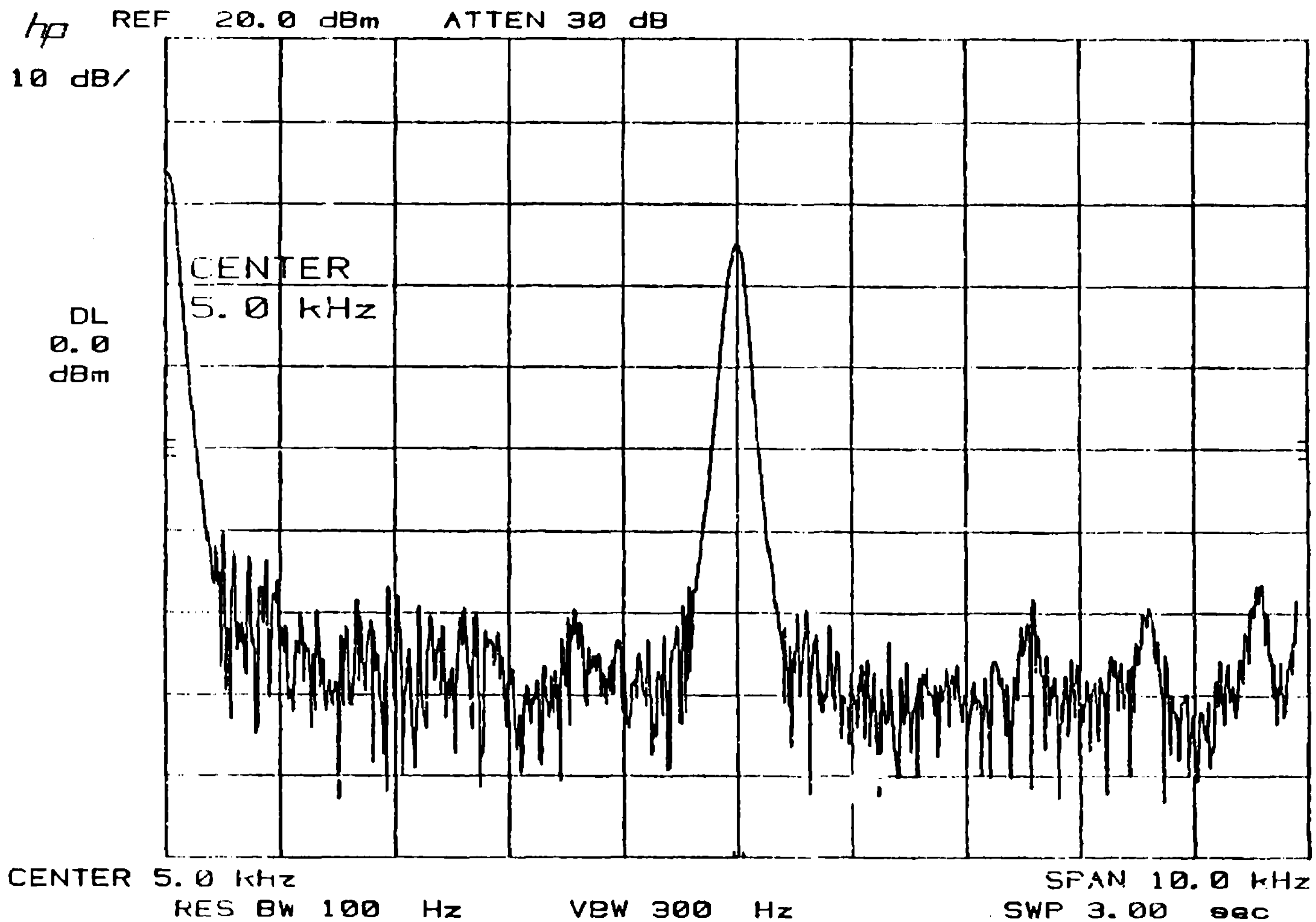


Figure 29. "Crosstalk" Spectral Analysis, Channel 2

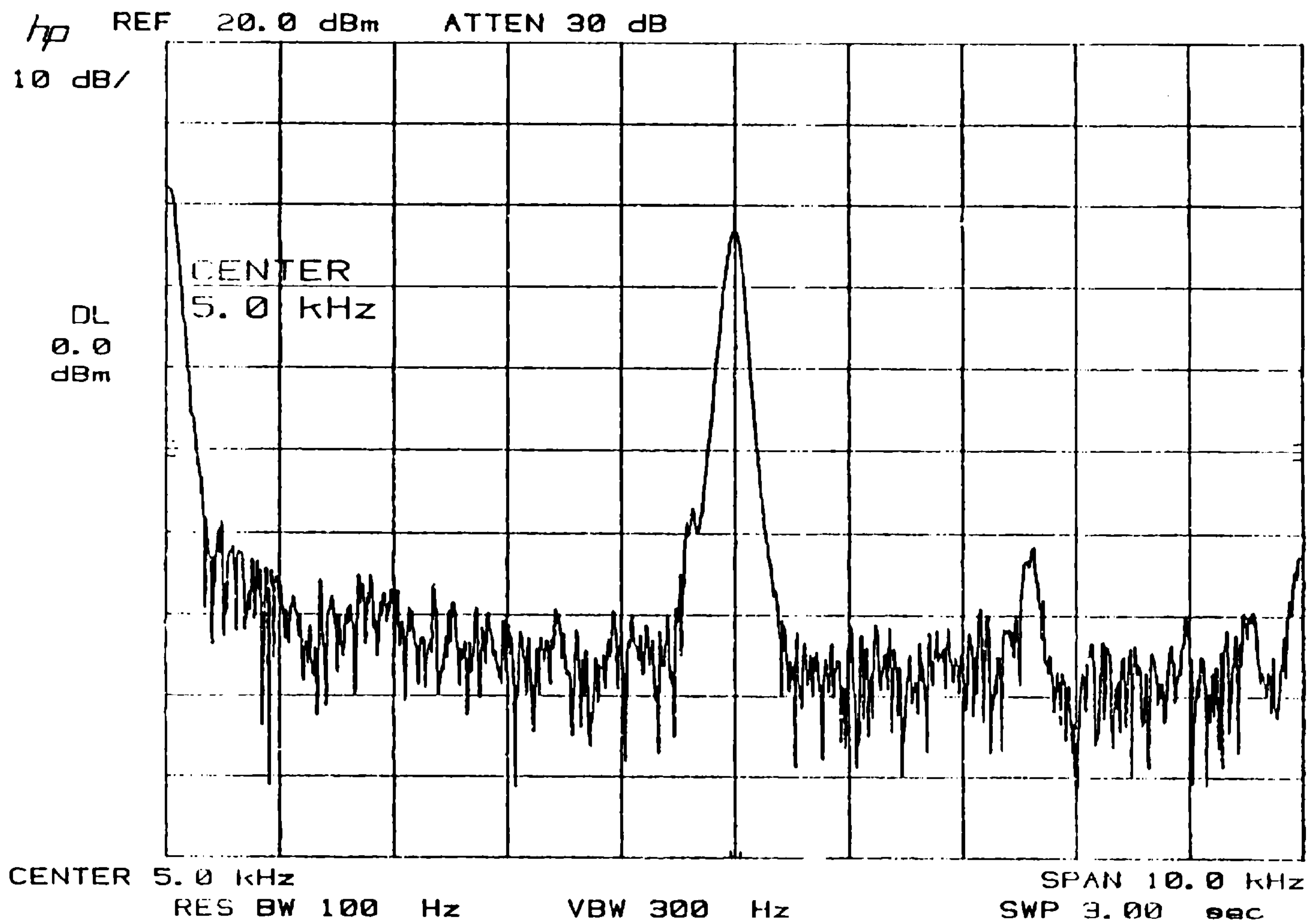


Figure 30. "Crosstalk" Spectral Analysis, Channel 3

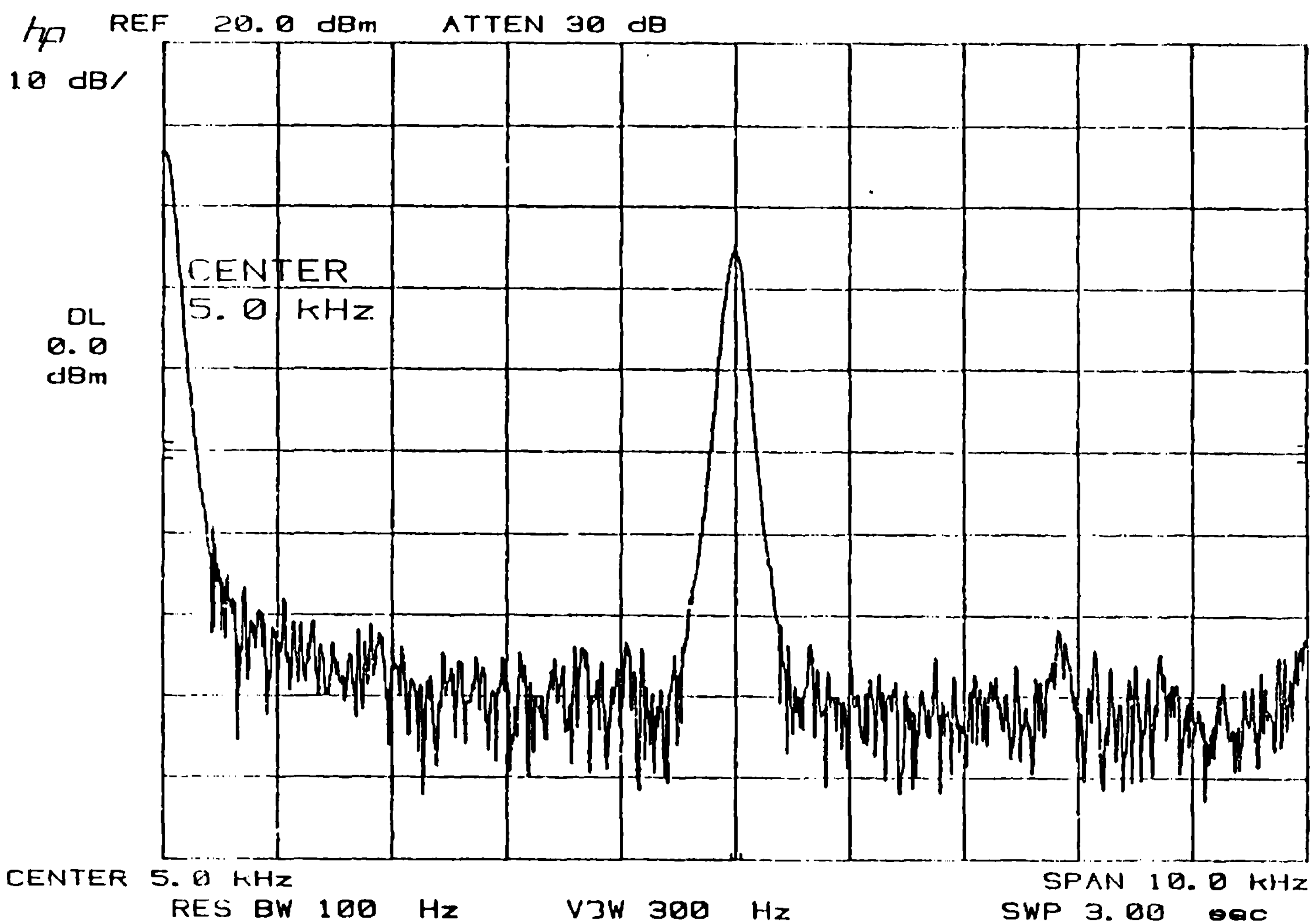


Figure 31. "Crosstalk" Spectral Analysis, Channel 4

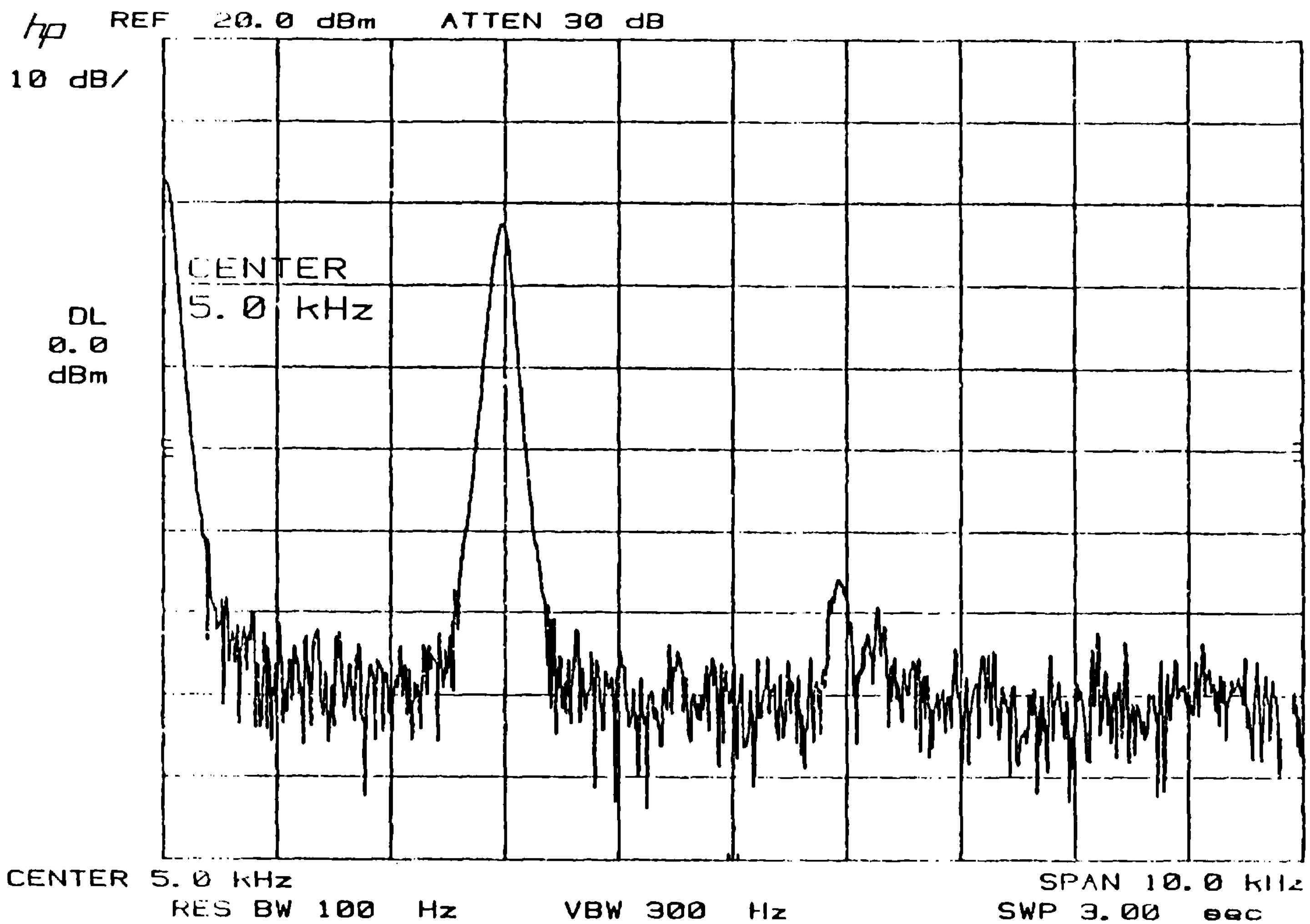


Figure 32. Adjacent Signal Spectral Analysis

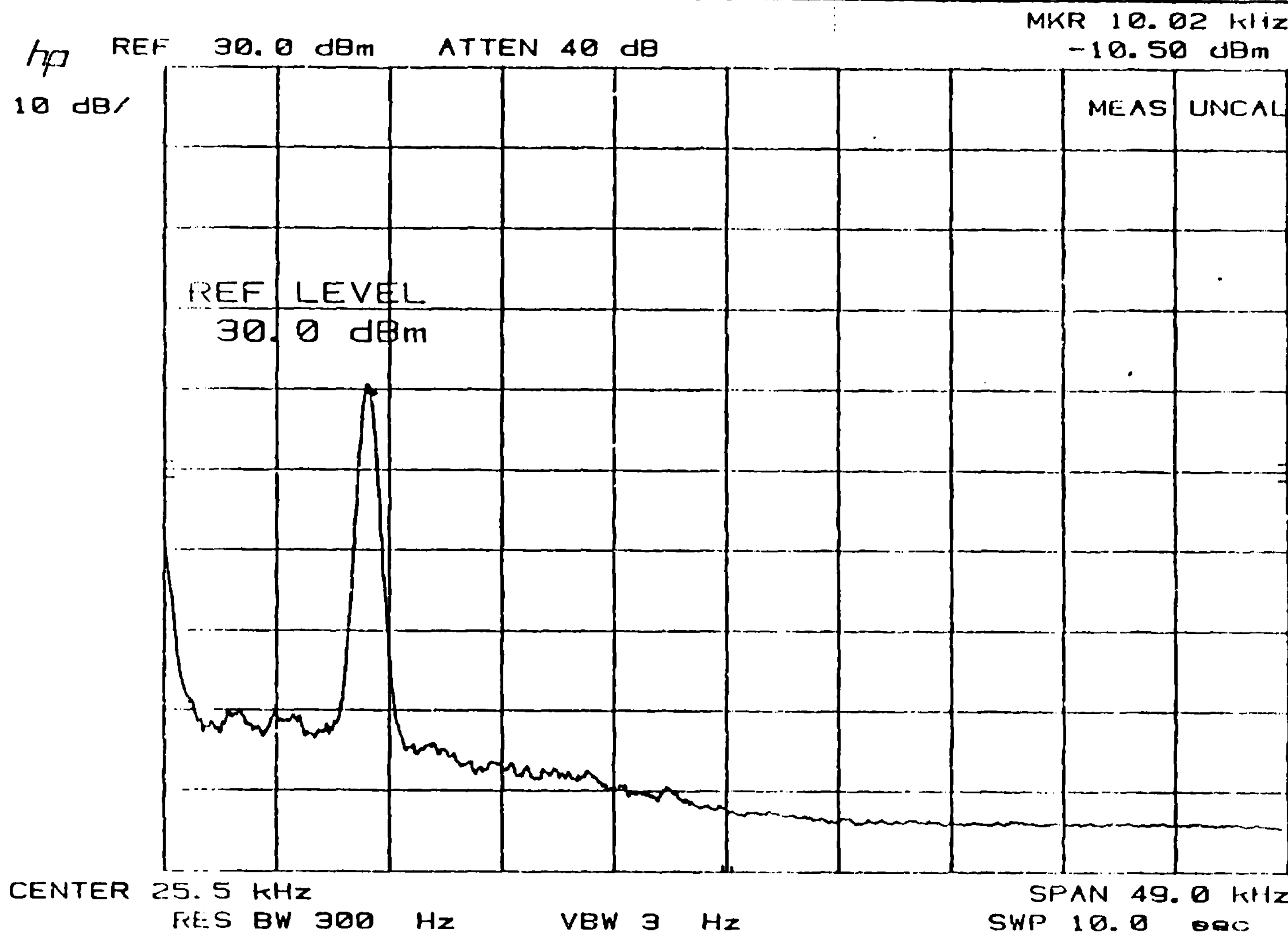


Figure 33. Harmonic Spectral Analysis, Channel 1

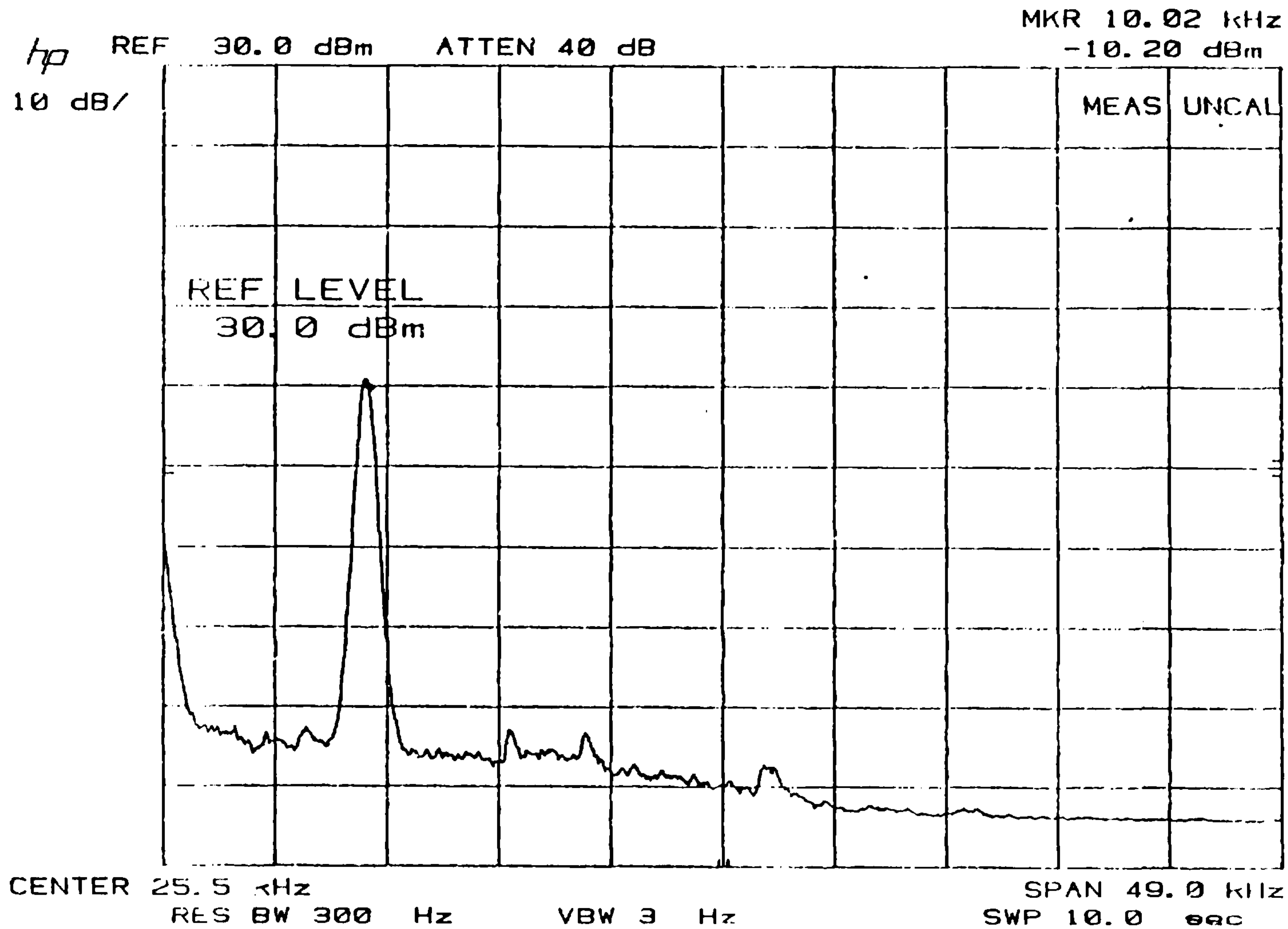


Figure 34. Harmonic Spectral Analysis, Channel 2

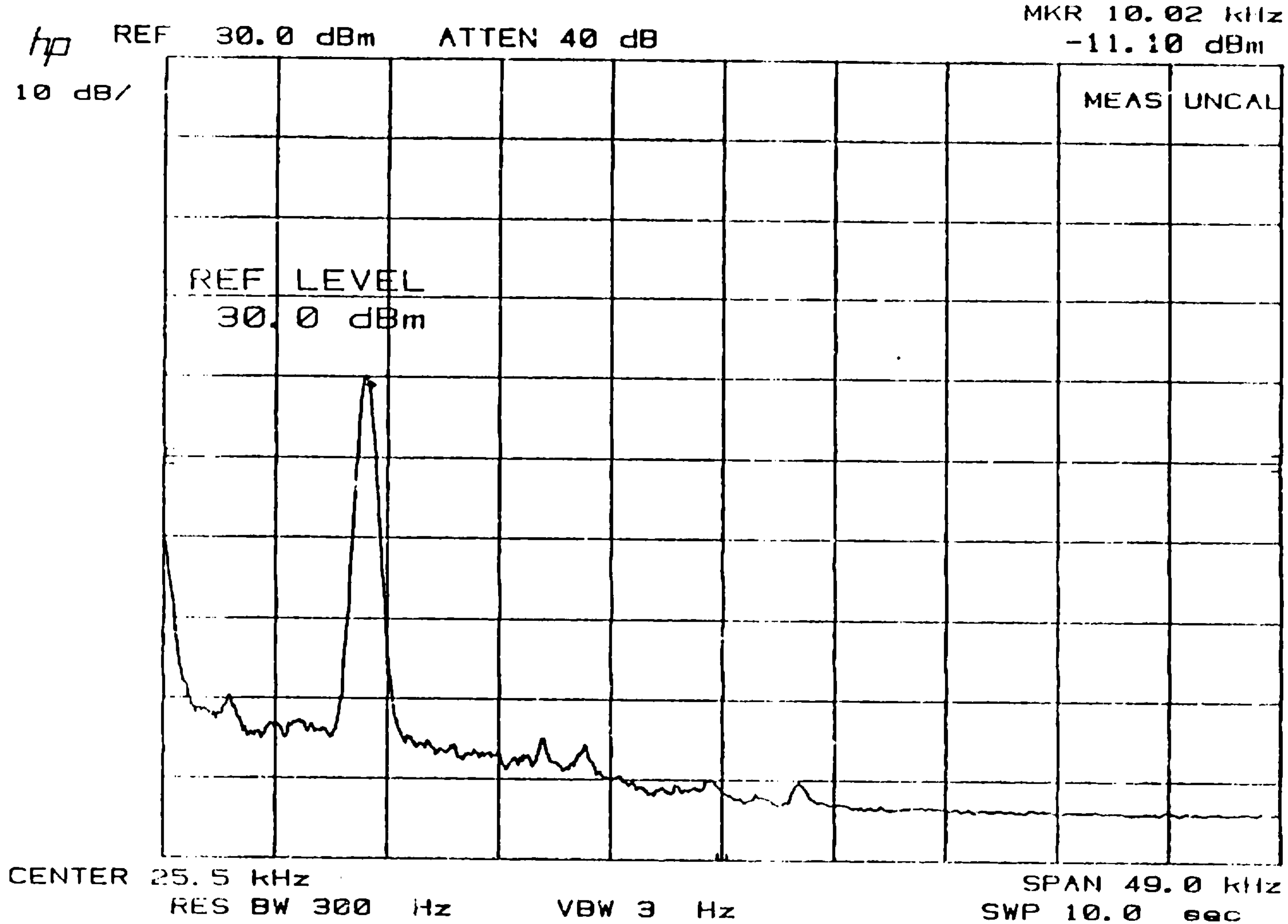


Figure 35. Harmonic Spectral Analysis, Channel 3

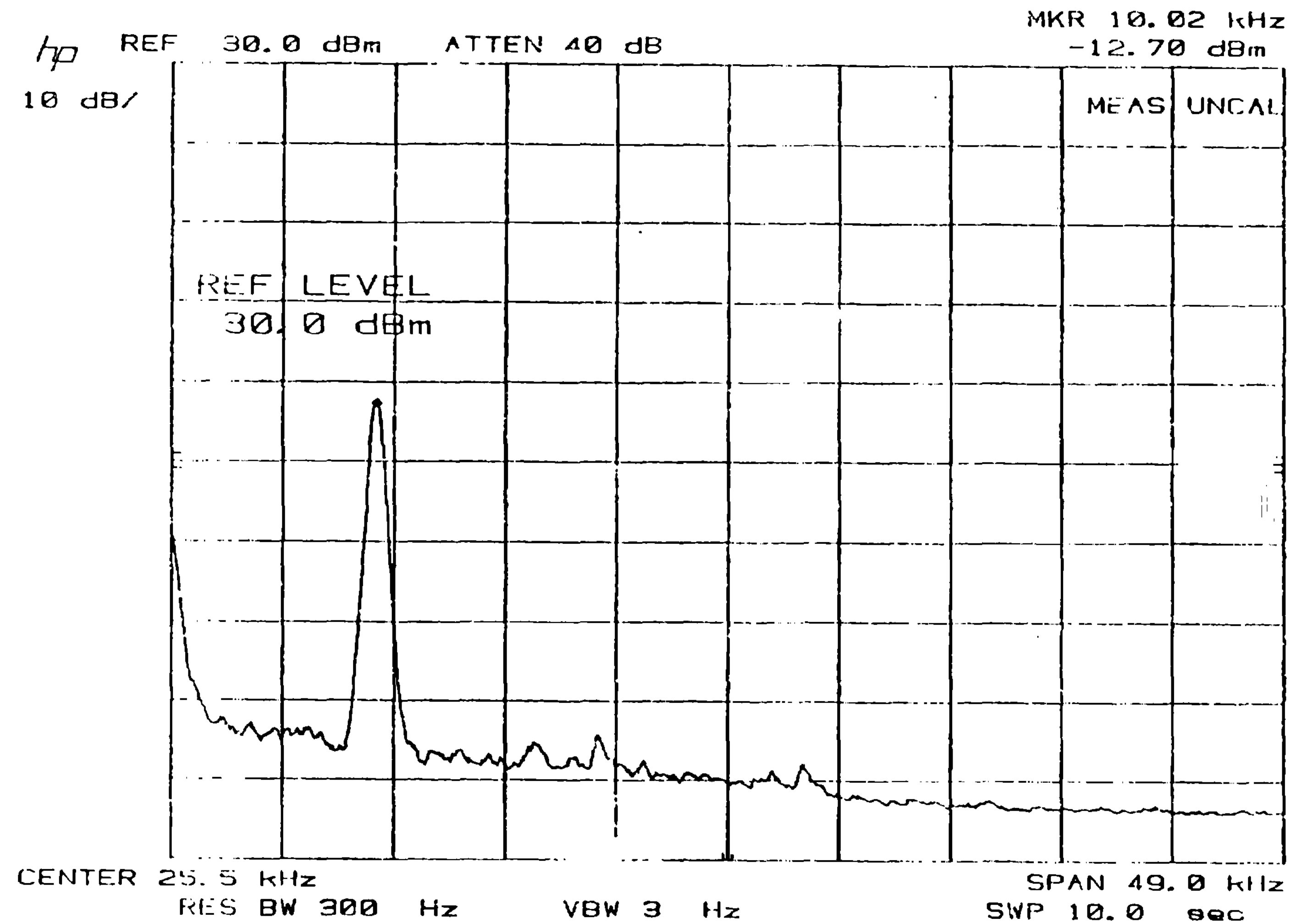


Figure 36. Harmonic Spectral Analysis, Channel 4

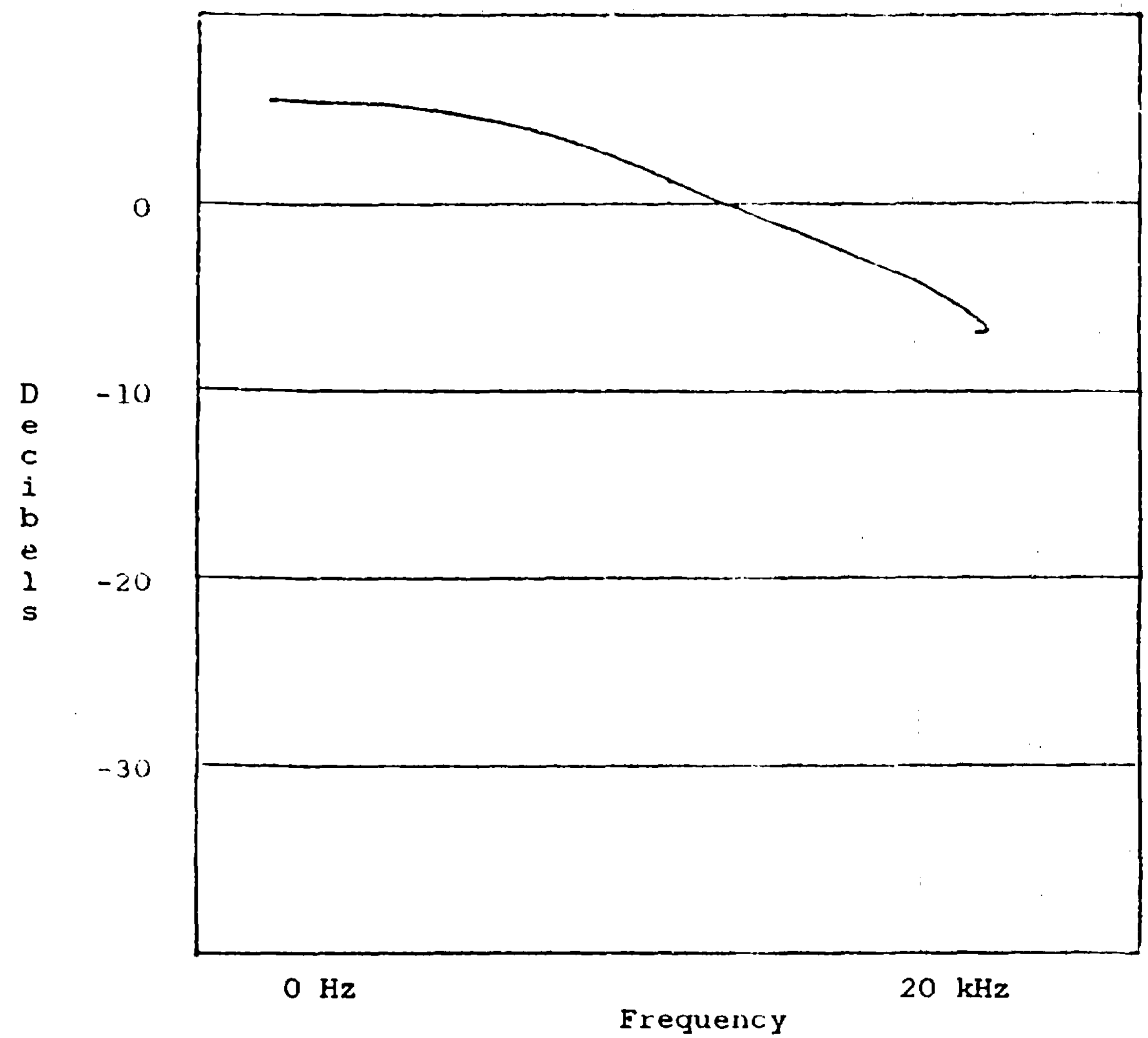


Figure 37. Gain characteristic, end to end, channel 1

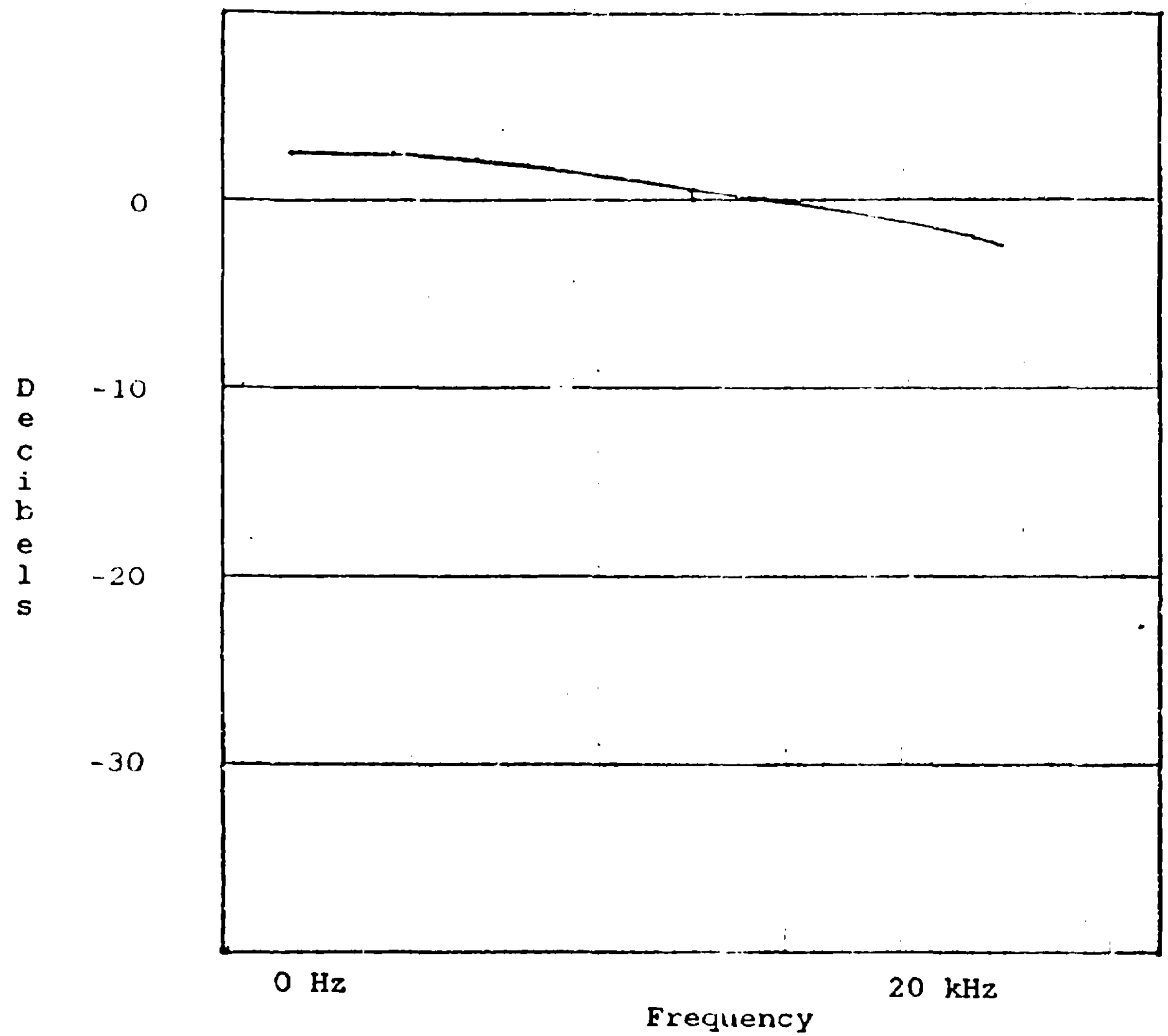


Figure 38. Gain characteristic, end to end, channel 2

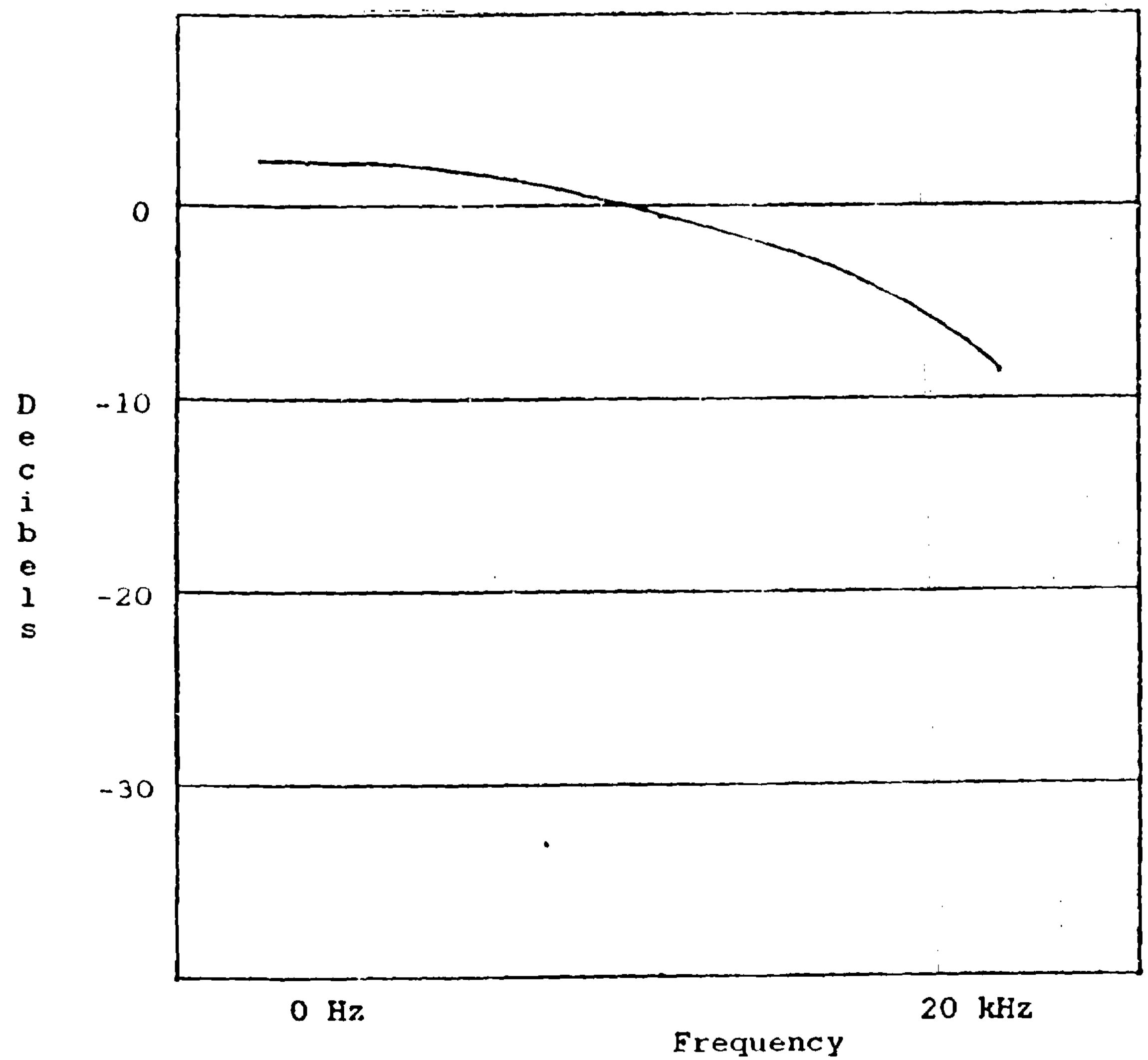


Figure 39. Gain characteristic, end to end, channel 3

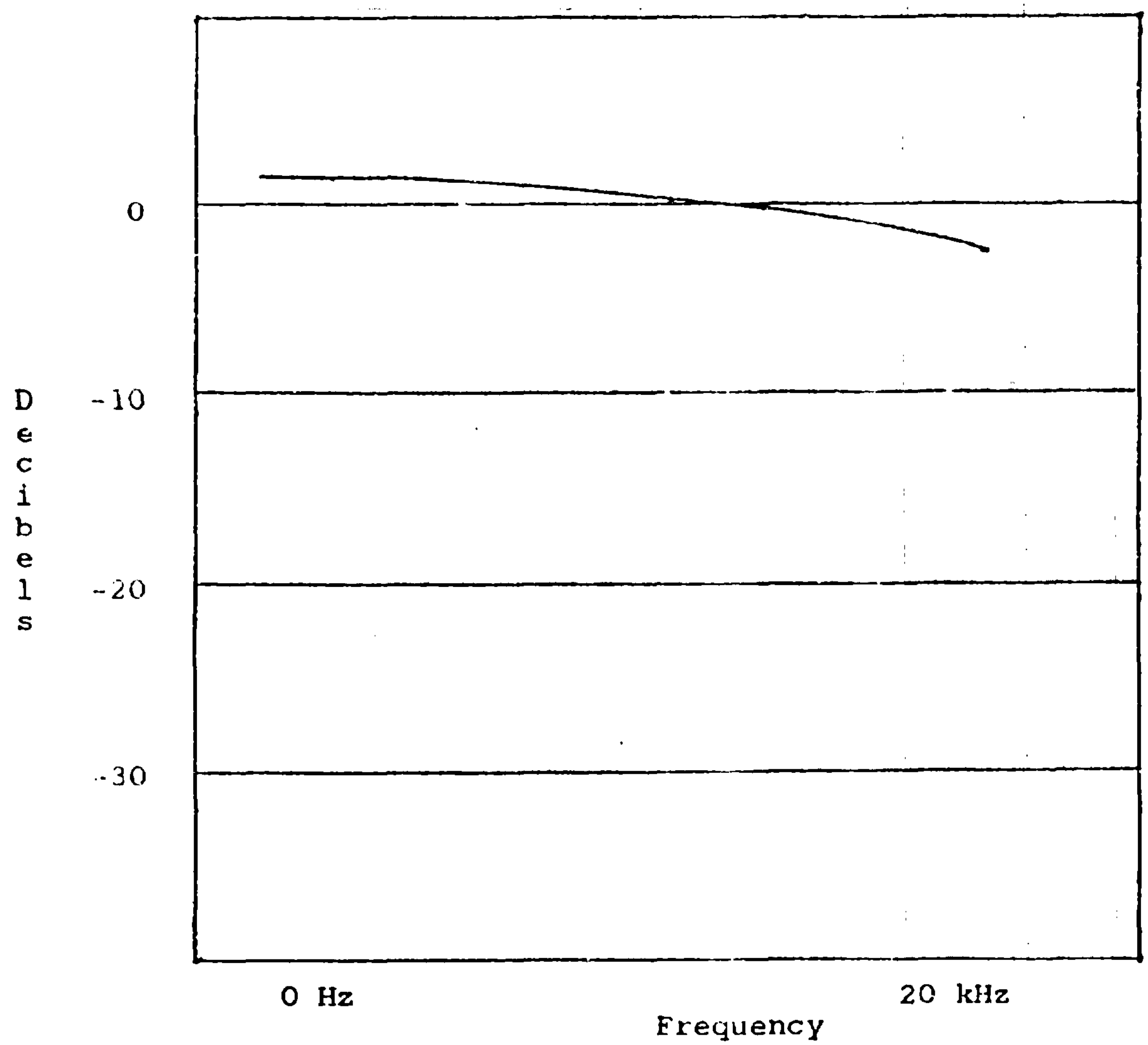


Figure 40. Gain characteristic, end to end, channel 4

where FM_R is the FM range, defined as the maximum frequency excursion of the carrier, and BW_I is the bandwidth of the applied information signal (Reference 8).

For complete transmission of an unattenuated signal, the receiver bandpass filters must pass, unattenuated, all frequency components of the FM signal. The quandry was that widening the passband of the filters on channels 1 and 3 to permit full transmission resulted in "user apparent crosstalk". This was due to exceeding the attentuation limits of the filters by too closely spacing the channels. If frequency components of adjacent channels exceed a threshold value of -40 dBv, interference (crosstalk) results. A decision was made, therefore, to optimize the "crosstalk" performance by limiting the passband of the filters at the expense of the high frequency (>16 kHz) capability of these two channels. In the frequency range 0-800 kHz allowed by the system electronics, three full frequency capable, low "crosstalk", channels could be maintained, but four could not.

D. PHASE LINEARITY

Unlike the gain response, the phase linearity of the transmitted signals on all four channels was excellent (Figures 41-44). This was expected as complex wave shapes (i.e., triangles) were easily reproduced after transmission through the link.

E. SYSTEM RANGE PERFORMANCE

Calculations of the minimum required received optical power were made using an insertable optical attenuator. As a first step, the insertion loss of the attenuator was measured and found to be 3.86 dB. At this point it was necessary to replace the HFBR-1402 optical transmitter with a HFBR-1404 due to this device's compatability with the

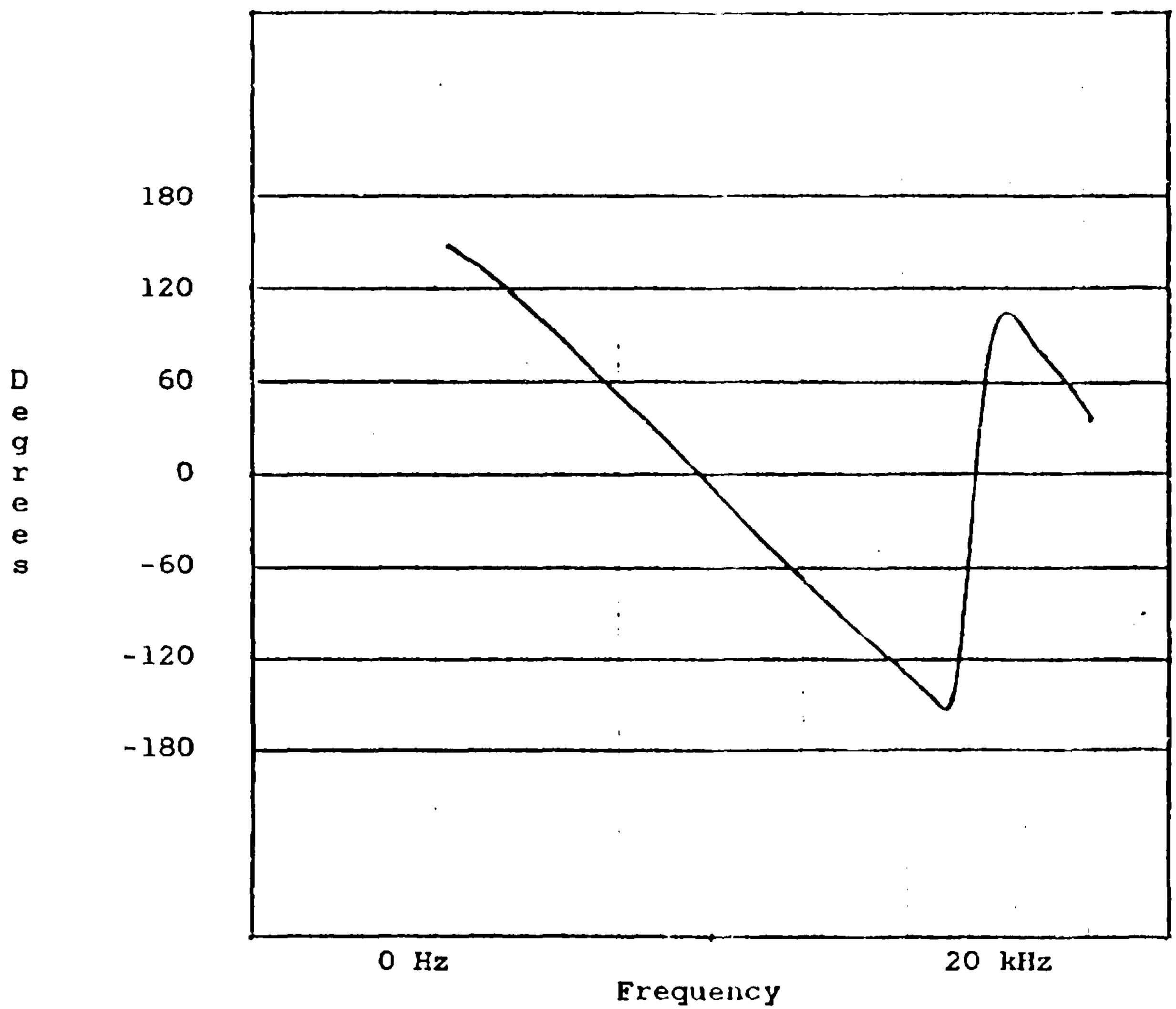


Figure 41. Phase characteristic, end to end, channel 1

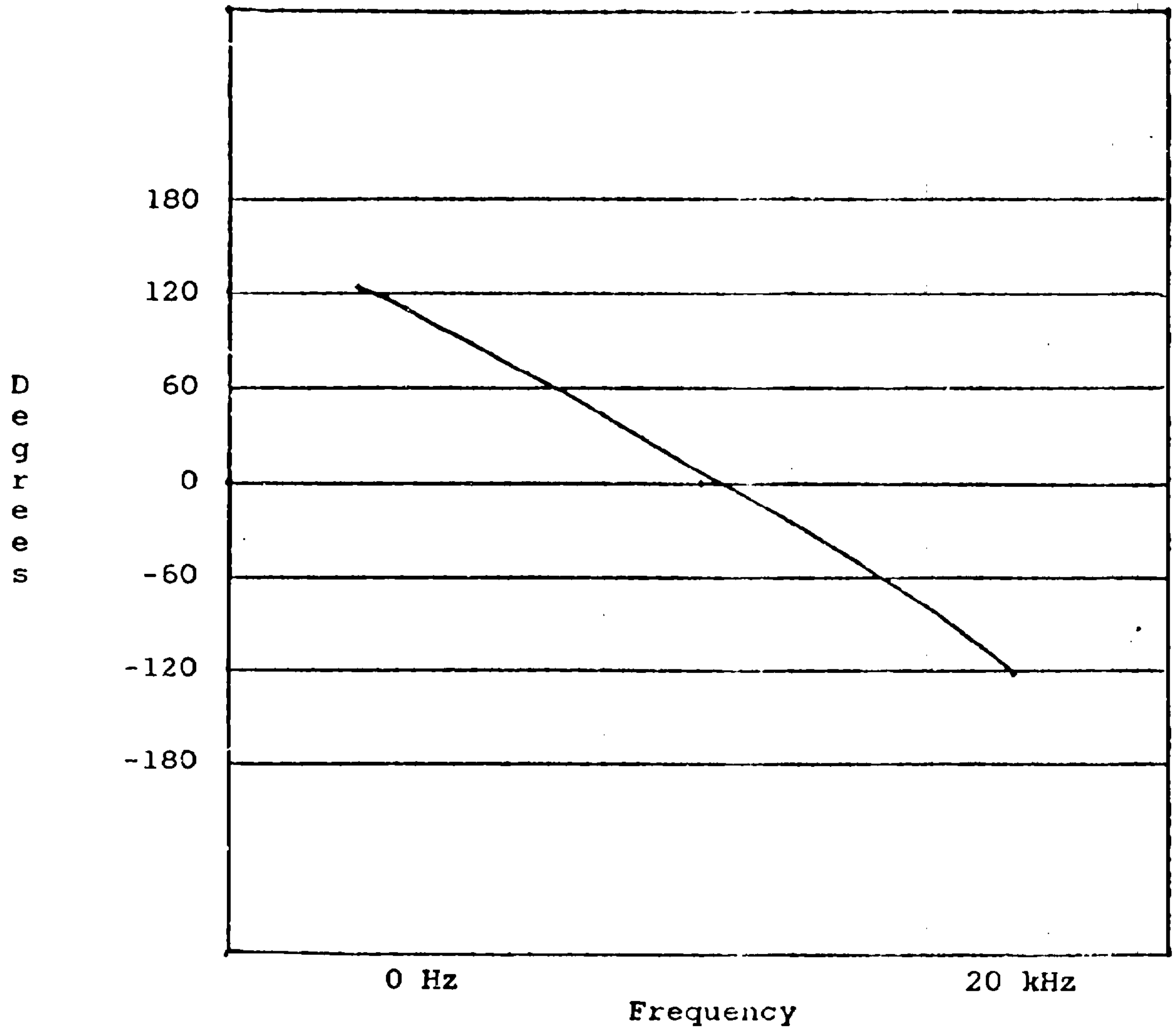


Figure 42. Phase characteristic, end to end, channel 2

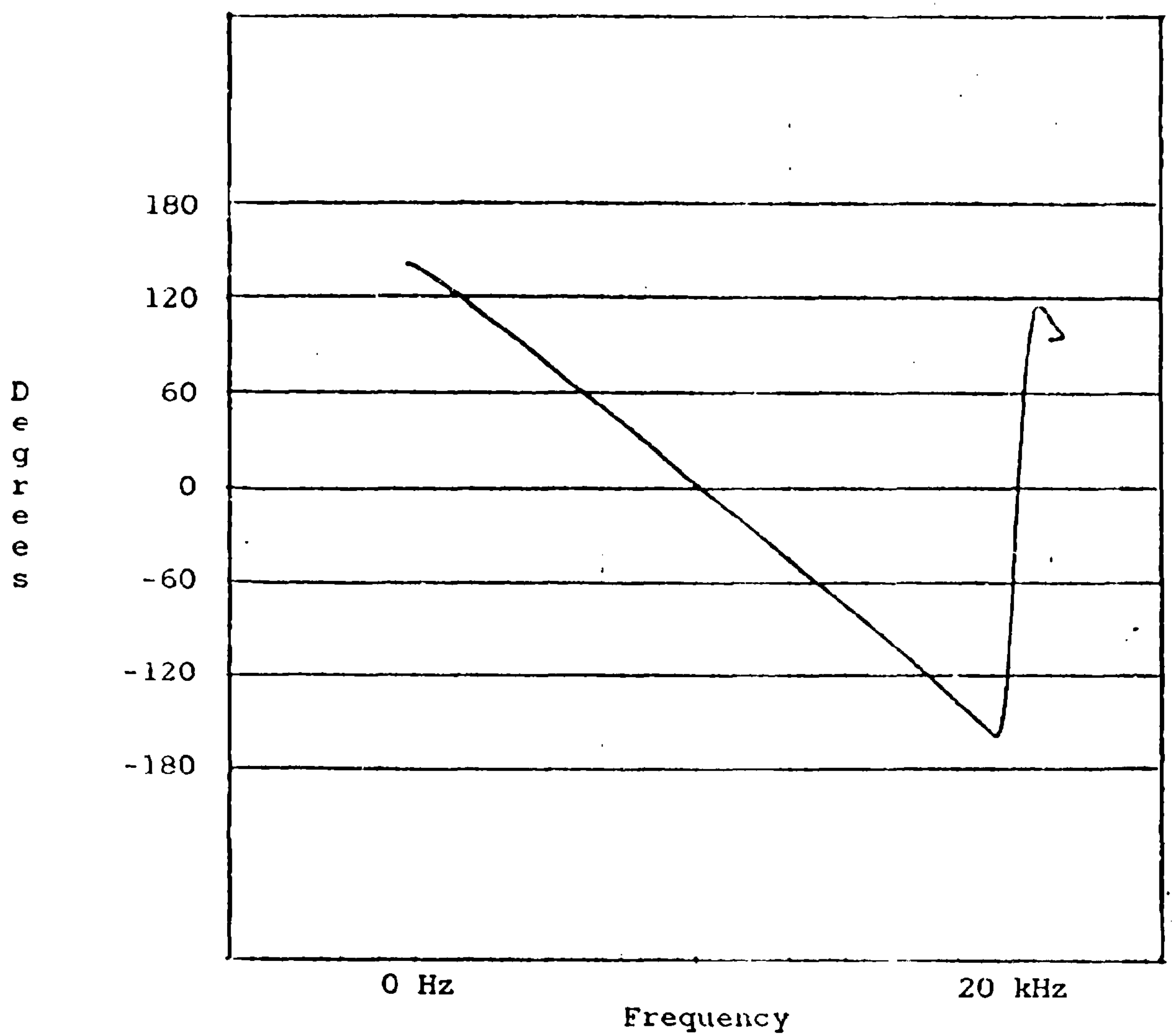


Figure 43. Phase characteristic, end to end, channel 3

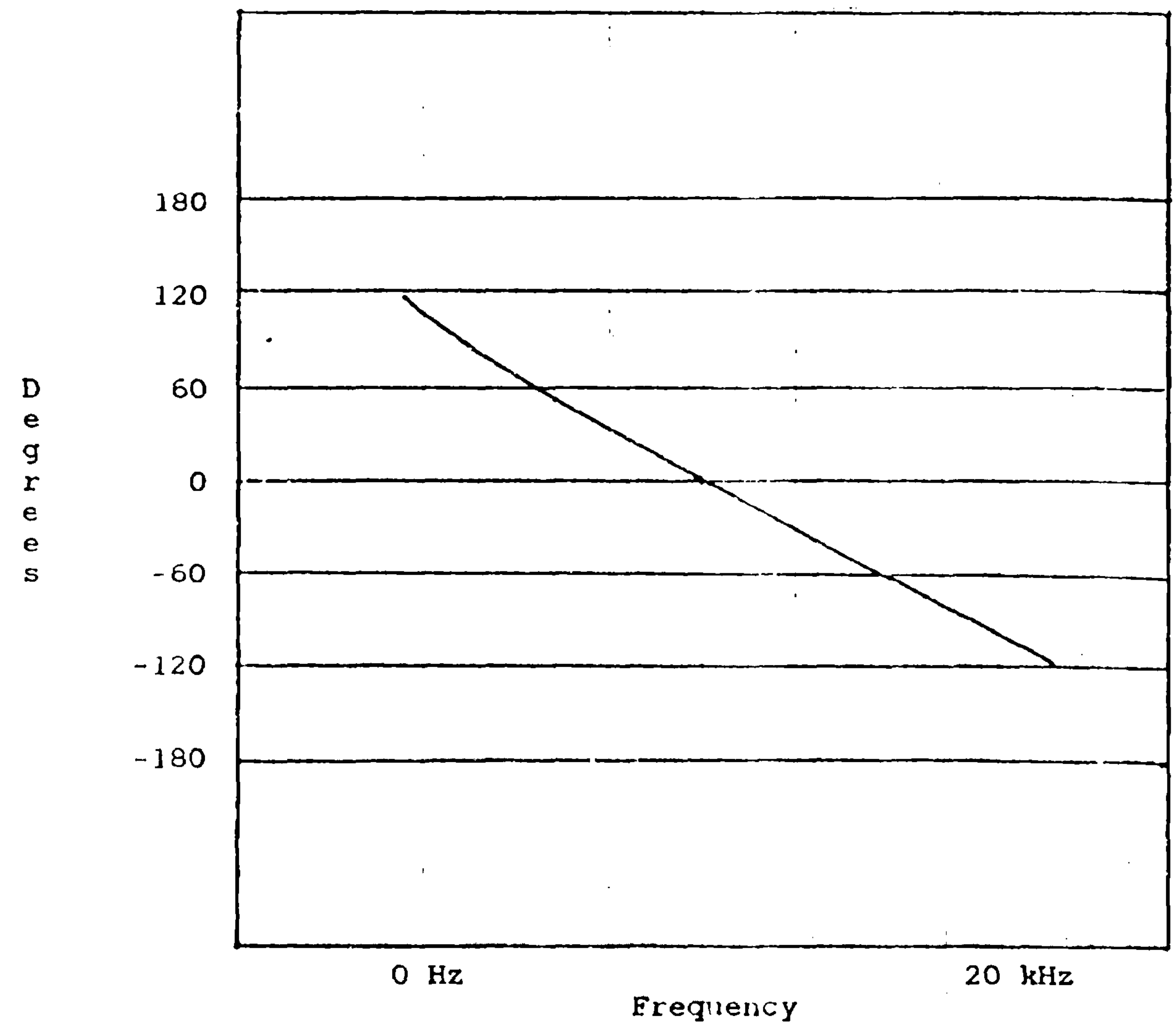


Figure 44. Phase characteristic, end to end, channel 4

50/125 um cable of the attenuator. This also had the advantage of providing more power, (-17.5 dBm), than did the 1402, albeit more costly. With this setup in place, the received signal was observed as the attenuation was increased. It was found that good quality on the weakest channel, number 4, was maintained to a total attenuation of -9.86 dBm. Admittedly, "good" is a subjective measurement and is somewhat situation dependent, however, an absolute benchmark is also available: the loss of receiver lock. This occurred, once again for the weakest channel, at a total attenuation of -11.86 dBm. Good signal quality requires, therefore, -27.36 dBm of received power while receipt of any signal at all requires -29.36 dBm. Ranges of several kilometers are readily available with these specifications. For example, given the parameters of this system as just discussed, the dynamic range (DR) of the system is

$$DR = 27.36 \text{ dBm} - 17.5 \text{ dBm} = 9.86 \text{ db} \tag{17}$$

This is the amount of power which can be lost and still maintain receiver lock. The system at hand used Siecor Optical Cable with 7 db of loss/km and assummed connector losses of 1 db each. Assuming no splices, the maximum range of this system is therefore

$$R = \frac{9.86 \text{ dB} - (2 \times 1 \text{ db/conn.})}{7 \text{ dB/km}} = 1.12 \text{ km} \tag{18}$$

Of course, most cable is in lengths of 1 km and therefore the range of this system is 1 km. The use of a laser optical source and an avalanche diode detector could be expected to extend this range several fold if required.

V. CONCLUSION

In overall performance, this system far exceeded the author's expectations as to the fidelity of the received signal and the absence of "crosstalk". As it stands, the system is a very usable one for the transmission of signals in the entire audio range. The reduction of capability in channels 1 and 3 for frequencies greater than 16 kHz is not viewed as a serious one. However, further exploration could be done in that regard as an almost infinite combination of frequency spacings, filter window widths, and center frequencies are available. Additionally, since a great portion of this section of the receiver required trial and error methods, significant progress is not impossible in this area.

A system such as this one could be a serious competitor for digital systems in the arena of low multi-channel applications. Although time did not permit further exploration, it is believed that an AM version utilizing many of the same components would have a capability of up to 10 channels. Such a system would be somewhat more complex and more subject to noise, but nevertheless, viable.

Applications for the present link are envisioned to include the original goal of transmission of hydrophone data plus the capability for intermachine multi-channel low data rate networks (it will transmit square waves up to 5 kHz).

As a final comment, the author is pleased to have apparently made a minor discovery which appears to quadruple the usable frequency range of the GIC filter.

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