3-1987

Conversion of the HP3577A Network Analyzer to a Spectrum Analyzer Mode of Operation with Low Noise and Cross-Talk

John E. Crowley
University of Tennessee - Knoxville

To the Graduate Council:

I am submitting herewith a thesis written by John E. Crowley entitled "Conversion of the HP3577A Network Analyzer to a Spectrum Analyzer Mode of Operation with Low Noise and Cross-Talk." I have examined the final electronic copy of this thesis for form and content and recommend that it be accepted in partial fulfillment of the requirements for the degree of Master of Science, with a major in Electrical Engineering.

David Rosenberg, Major Professor

We have read this thesis and recommend its acceptance:

J. R. Roth, T. V. Blalock

Accepted for the Council:
Carolyn R. Hodges

Vice Provost and Dean of the Graduate School

(Original signatures are on file with official student records.)

To the Graduate Council:

I am submitting herewith a thesis written by John E. Crowley entitled "Conversion of the HP3577A Network Analyzer to a Spectrum Analyzer Mode of Operation with Low Noise and Cross-Talk." I have examined the final copy of this thesis for form and content and recommend that it be accepted in partial fulfillment of the requirements for the degree of Master of Science, with a major in Electrical Engineering.

David Rosenberg, Major Professor

We have read this thesis
and recommend its acceptance:

Accepted for the Council:

Cw Minkel

Vice Provost
and Dean of the Graduate School

CONVERSION OF THE HP3577A NETWORK ANALYZER TO A SPECTRUM ANALYZER MODE OF OPERATION WITH LOW NOISE AND CROSS-TALK

A Thesis

Presented for the

Master of Science

Degree

The University of Tennessee, Knoxville

John E. Crowley

March 1987

ACKNOWLEDGEMENT

The author is deeply indebted to Professor J. R. Roth for his infinite patience, financial support, excellent working conditions, and good advice.

The author also extends his gratitude to the remaining members of his thesis committee: Professor T. V. Blalock and Associate Professor D. Rosenberg.

Special thanks goes to Professor J. Googe who had answers to problems at a particularly difficult time.

Finally, the author wishes to express his appreciation to his parents for their support and encouragement throughout his academic career.

This work was supported by the Air Force Office of Scientific Research under Contract #86-0100 (ROTH).

ABSTRACT

A system has been designed which modifies the HP3577A Network Analyzer so that arbitrary signals in the range from 5 Hz to 10 MHz can be compared in amplitude and phase. The system will be used to sample density or potential fluctuations at two points in a turbulent plasma, and measure the frequency, amplitude, and phase of signals detected by the probes. This modification is totally external to the HP3577A Network Analyzer, and has been designed to minimize the cross-talk between the two channels, so that nonlinear mode coupling processes in the plasma can be distinguished from mode coupling (cross-talk) originating within the instrumentation itself. The composite has a dynamic range from 0 dBm to -60 dBm, crosstalk of no more than -40 dB between channels, a maximum spurious response -40 dB below the maximum input signal, and an amplitude and phase accuracy of 1 dB and 5 degrees, respectively.

TABLE OF CONTENTS

LIST OF FIGURES

FIGURE PAGE

INTRODUCTION

This paper reports the modification of a piece of Hewlett-Packard test equipment. It is desired to modify the HP3577A network analyzer such that the unit will accurately measure the amplitude and phase of arbitrary signals. The frequency range for the signals of interest is between 5 Hz and 10 MHz. The amplitude range for the signals of interest is between 0 dBm and -80 dBm. The modification is to be totally external to the HP3577A and have two-channel capability.

The need for a two-channel device, capable of measuring the amplitude and phase relationship between frequency components of broad-band spectra, results from the study of turbulence in plasmas. Plasmas are hot gases which contain electrons and ions. These electrons and ions move about in the plasma creating varying electrostatic potentials. While studying the way in which the charged particles gain and lose energy, it is very desirable to be able to determine the fluctuating electrostatic potential at different points in the plasma. The frequency components of the plasma potential at two different points helps determine the way in which electrons, ions, and propagating waves move in a plasma. For example, if two points of the plasma have a strong frequency component at a particular frequency, coherent structures or waves could be moving from one point to the second as a function of time. The phase relationship between the spectral components determines if fluctuation-induced transport exists between the two points.

If the phase relationship of the spectral components between the two points is coherent, mode coupling is implied. If the phase relation between the two is random, then there is probably no coupling at that frequency.

In addition to finding out if there is mode coupling in a plasma, the engineer must be aware of the coupling between the channels in his test equipment. If the crosstalk between channels in the test equipment is -20 dB, then it will be hard to determine if any observed phase coherency below this level is due to mode coupling in the plasma or to the crosstalk between channels. The crosstalk in the HP3577A is -100 dB between channels. It is desired that the crosstalk between channels in the modification be greater than -40 dB.

The need to measure the phase relationship between spectral components is the reason why a two-channel data-handling system must be modified to measure mode coupling in plasmas. There is no off-the-shelf test equipment capable of phase measurements over the frequency range of interest. There are several devices available that measure phase between input signals. A dual channel spectrum analyzer capable of phase measurements will accomplish this measurement task. However, most of these units work at low frequencies. The HP3562A, for example, has a high frequency of 3 dB point of 100 kHz (and a cost of $23,100), which is below the region of interest in the current research program in the UTK Plasma Science Laboratory. Vector volt-meters, which measure the

amplitude and phase difference between two input signals could be used. However, these units operate on one frequency and cannot produce useful measurements on broad-band spectra. A signal digitizer with two channels could take accurate measurements over a very short period of time, but not on a steady state basis. Since it is desirable to move the measurement point at will, a digitizer is not the answer. Indeed, without modification, the HP3577A, a network analyzer, will not produce accurate results. After careful study it was determined that with proper signal processing, the HP3577A could be used for this task.

A note on the literature is now in order. Much has been written in the plasma field on mode coupling. However, because of its recent appearance on the market, apparently no one has ever attempted to modify the HP3577A to measure mode coupling. Therefore no literature was found on this subject. In addition, no literature was found that discussed the modification of the HP3577A for accurate measurement of arbitrary signals in two channels, with low crosstalk.

The modification, which is the subject of this thesis, is a narrow band preselector which tracks the source frequency of the HP3577A. The modification consists of a moving filter or tracking filter. The literature contains much information dealing with filters. Two main types of tracking filters are state variable and yttrium-iron-garnet (YIG) (ref. 1). State variable filters are filters containing operational amplifiers. In their operation, some

amplifier characteristic can be varied, affecting some critical frequency in the filter's operation. State variable filters have little to do with the operation of the moving filter contained in the HP3577A modification. YIG filters, on the other hand, are very analogous to the situation present in the modification. These filters operate on the principle that by changing the DC magnetic field in a ferrite, the associated resonance of the ferrite can be adjusted. Bandpass filters in the range of 2 to 18 GHz can be constructed with Q's of up to 500. Preselectors that help minimize aliasing in microwave spectrum analyzers are YIG filters. Since the region of operation for the YIG filter is much higher than the filter needed in the modification, YIG filtering could not be considered.

The filter implemented in the modification is a composite filter which contains crystal filters and mixers. An extensive computer search was done at the U. T. library. Sixty references were found which contained one of the key words. However, none of the abstracts showed any work discussing adjustable filters containing crystal filters and mixers. The key words used included: filter, variable, mixer, state variable, controllable, crystal, and composite.

The thesis is divided into seven sections. Section I discusses the HP3577A network analyzer's normal operation and the problems associated with the measurement of arbitrary signals. Section II discusses the approach taken in this thesis for solution

and estimated system performance. The second section also contains some experimental results to aid understanding the operation of the modification. Section III discusses the design and operation of tracking filter. Section IV details the design and operation of the preamplifier. Section V discusses the way in which the local oscillator that drives the filter board is produced. This part of the system will be called the local oscillator generation system. The spectral purity of the local oscillator and its effect on system performance will be discussed. Also in section V, the concept of composite filters will be discussed. Composite filters are contained in both the preamplifier and the local oscillator generation system, but their operation was particularly important to the local oscillator generation system. Sections III, IV, and V will also contain performance data. Section VI discusses experimental results and the calibration of the entire system. Section VIII, the appendix, contains schematic diagrams and data sheets not appropriate elsewhere.

I. PROBLEM STATEMENT

To completely describe the task of converting the HP3577A network analyzer, the operation of the unmodified unit must first be discussed. The HP3577A is a signal/response instrument designed to measure the magnitude and phase transfer characteristics of linear systems. A signal/response instrument measures a transfer function by comparing the signal fed to the device under test (DUT) with the signal returning from the DUT. This is accomplished in the following manner: The HP3577A unit, in the normal mode of operation (Figure 1), generates a signal of known frequency, magnitude, and phase. This signal is fed into the device under test and the output of the DUT is fed back to the HP3577A. With proper calibration and extensive digital processing, the HP3577A measures the magnitude and phase characteristics of the DUT to within 0.01 dB and 0.005 degree, respectively.

Note that the receiver samples only 401 frequency points across the frequency range of interest. The effects of this discrete frequency operation on the system conversion will be discussed in the performance section. The rated frequency response of the unit is between 5 Hz and 200 MHz.

The mode of operation into which it is desired to convert the unit is shown in Figure 2. Notice that with the modification of any arbitrary signal can be measured by this "spectrum analyzer" mode of operation.

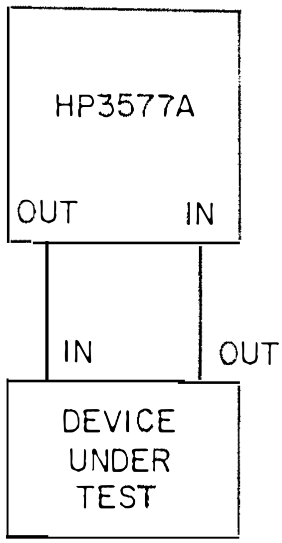


Figure 1. HP3577A in normal mode of operation.

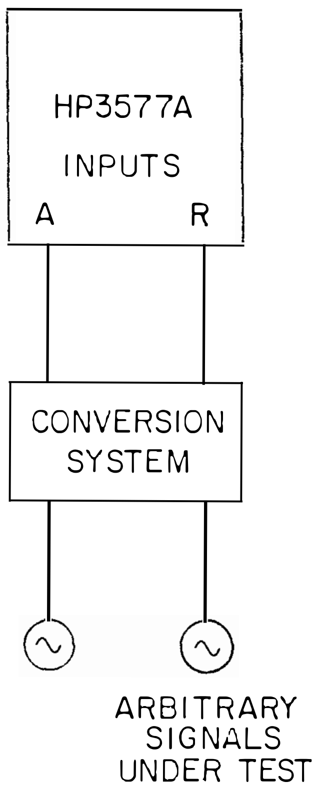


Figure 2. Desired configuration of the HP3577A.

At first glance, the user might feed an unknown signal directly into the HP3577A. When this is done, the frequency components of the signal being fed into the HP3577A are indeed measured (shown as spikes on the display screen). However, spurious internally-generated responses will be displayed at frequencies 500 kHz below each spectral component actually being fed into the HP3577A. The amplitude of each spurious signal will be indistinguishable from the corresponding real signal. Along with these obviously spurious signals, many other responses at unpredictable frequencies may be present within the band of interest.

These results are demonstrated with a simple experiment. A 4 MHz sine wave with amplitude of -25 dBm is fed directly into the HP3577A. Figure 3 shows the display screen in the narrow region around 4 MHz. The sine wave is shown at the correct frequency and amplitude (-25 dBm). Figure 4 is an expanded frequency sweep. Notice the second signal at a frequency of 3.5 MHz. The signal at 3.5 MHz is not present in the input, but is being spuriously generated by the HP3577A. Figures 5 and 6 show the display screen from 50 kHz to 625 kHz and 625 kHz to 1.25 MHz respectively. Notice the many spurious responses due to the one 4 MHz sine wave. The direct-input technique is useless for the study of broad-band signals associated with plasma turbulence. The generation of the spurious responses will now be explained.

Figure 7 is a block diagram for the input receiver of the HP3577A. The presence of the multiple spurious signals can be

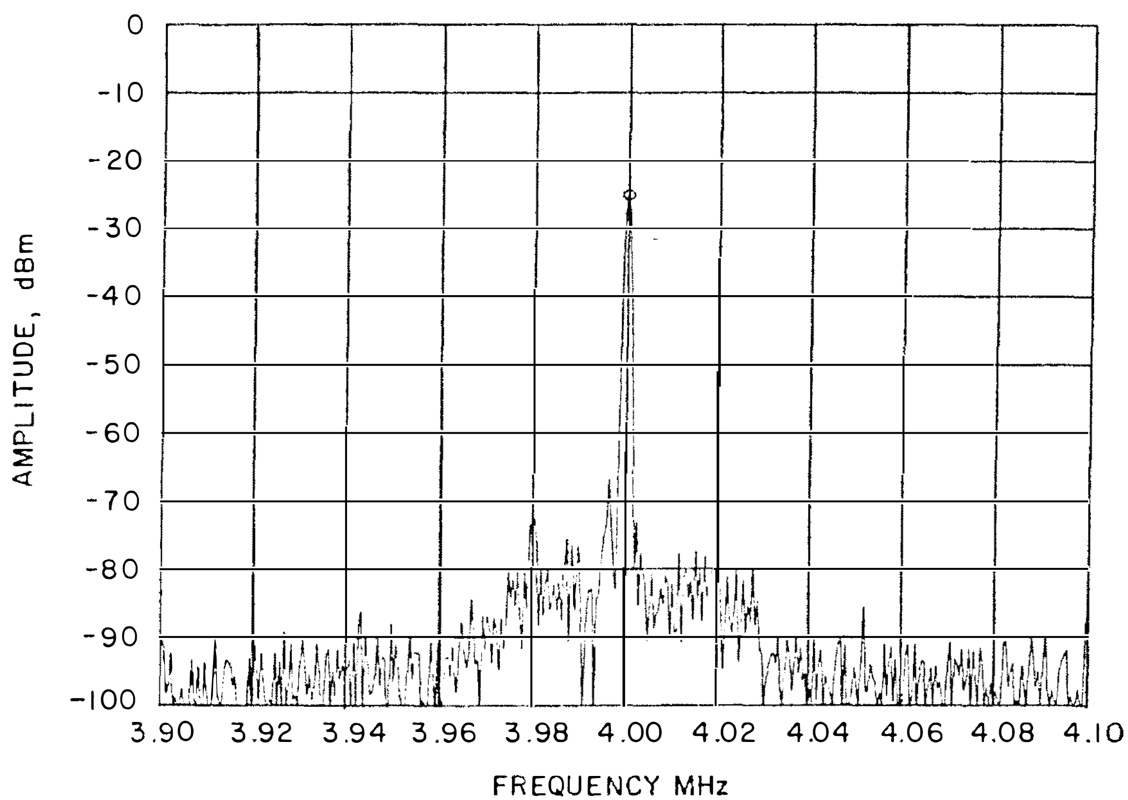


Figure 3. HP3577A display with 4 MHz sine wave input.

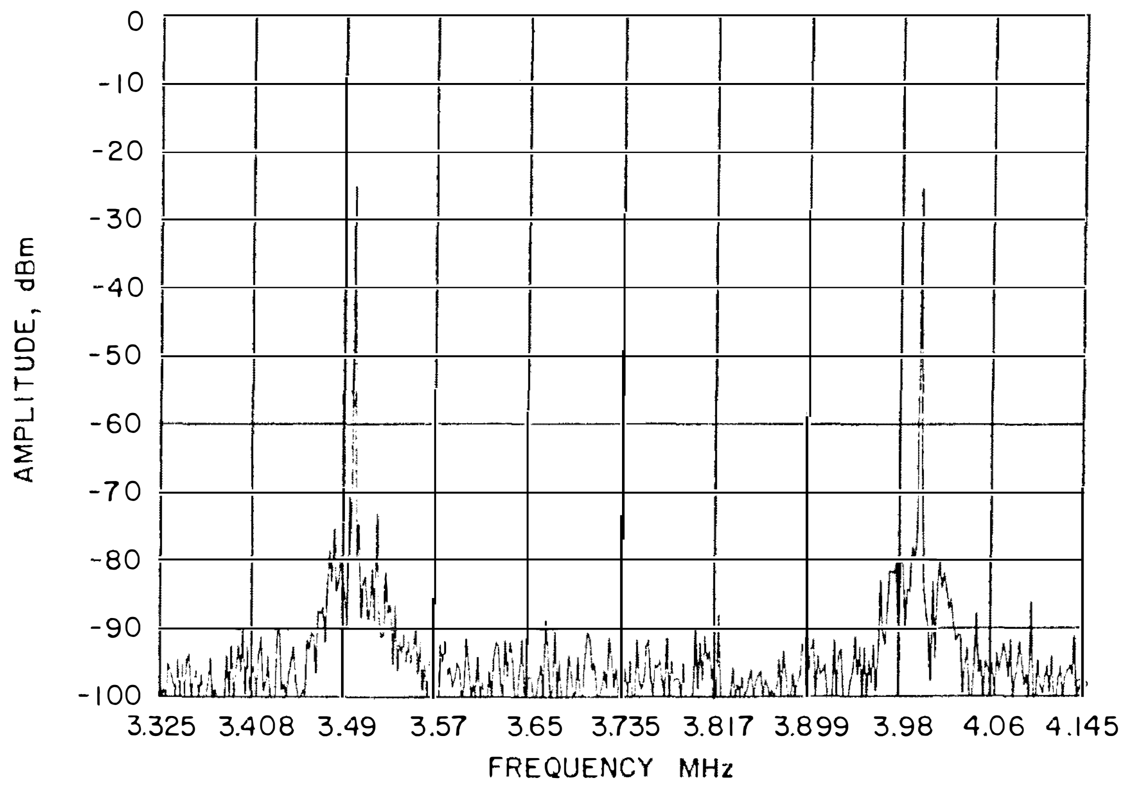


Figure 4. HP3577A display with 4 MHz since wave input and spurious response at 3.5 MHz.

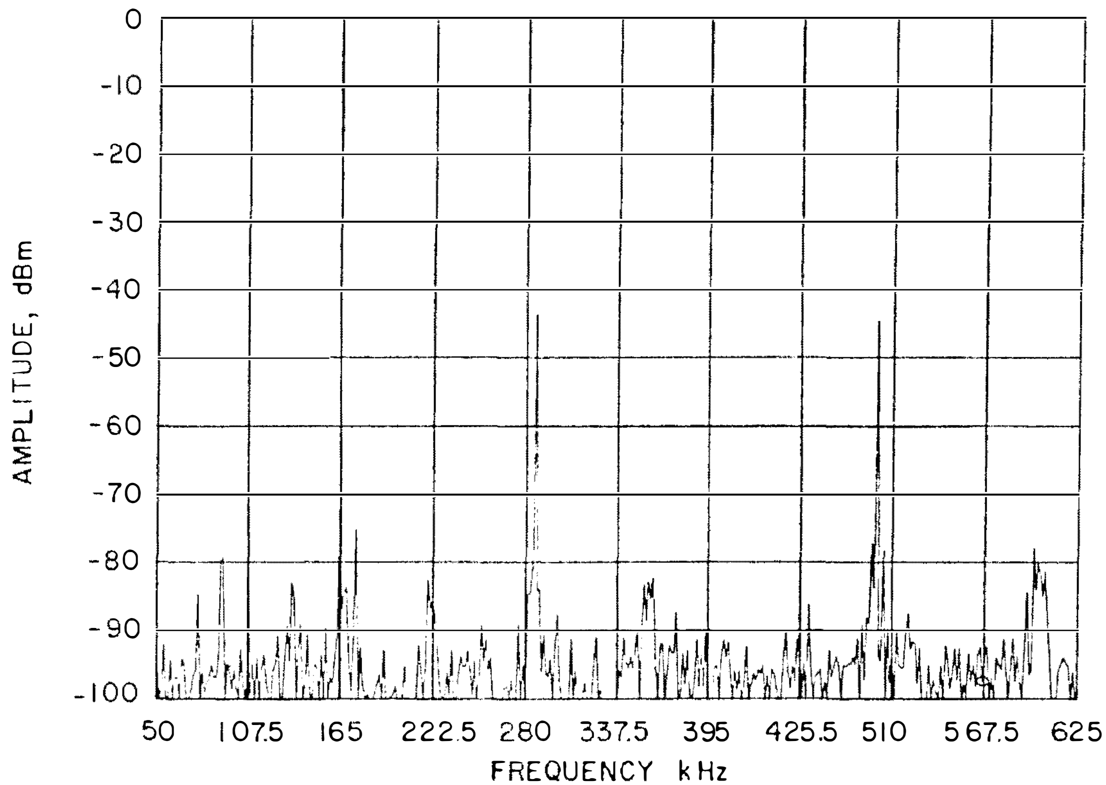


Figure 5. Demonstration of spurious signals in 50 kHz to 625 kHz bandwidth.

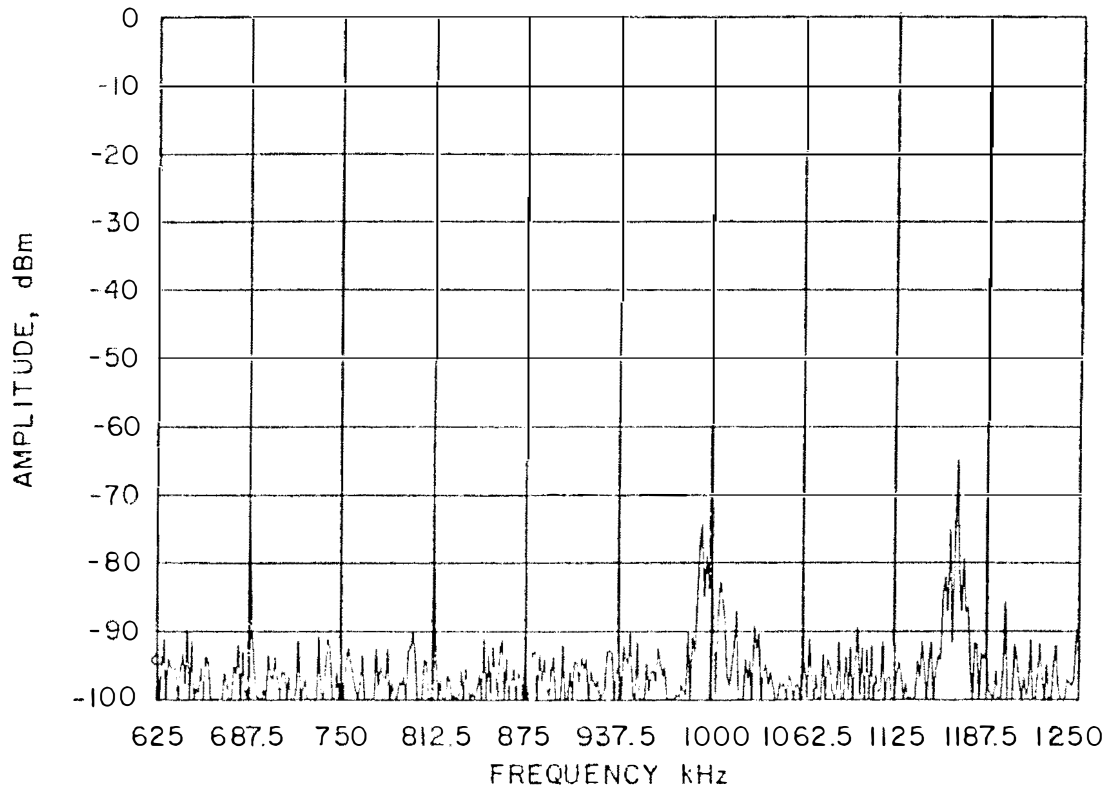


Figure 6. Demonstration of spurious signals in the 625 kHz to 1250 kHz bandwidth.

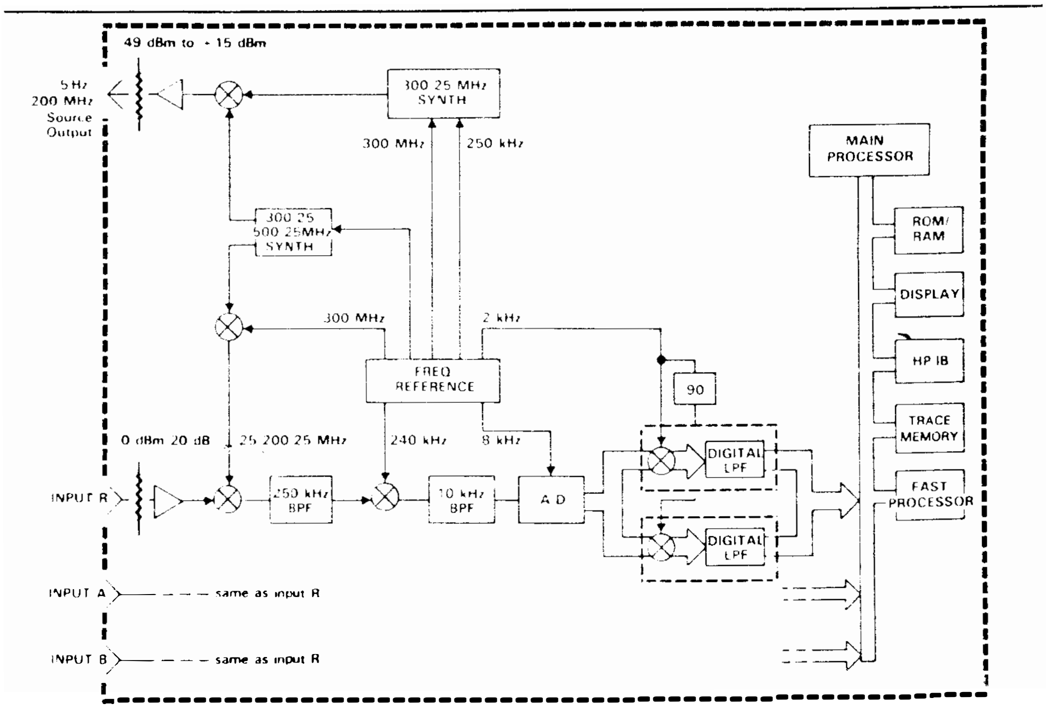


Figure 7. HP3577A receiver block diagram.

Source: Operating Manual Model 3577A Network Analyzer, Hewlett-Packard Company, P. O. Box 69, Marysville, Washington 98270, U. S. A., Microfiche Part No. 03577-90050, section 5, p. 8, November 1983.

In the normal mode of operation, the HP3577A receiver mixes the signal from the device under test to the intermediate frequency of 250 kHz. The local oscillator that drives the input mixer remains at a frequency 250 kHz above the signal fed into the DUT. Since the HP3577A was designed to measure transfer characteristics of linear systems, the signal to the HP3577A receiver should be at the same frequency as the signal from the HP3577A source. Herein lies the operational strength of the unit and the problem for

accurate measurement of arbitrary-input signals. When the HP3577A is measuring the magnitude and phase characteristics at a certain frequency, normal operation stipulates that only one frequency (the test frequency) be present.

The arbitrary signal is fed into the unit in the steady-state. Since mixers create sum and difference frequencies, 250 kHz signals will be fed into the intermediate filter bandpass any time the local oscillator is 250 kHz above or below all input spectral components. The unit will display two frequency components, only one of which is actually present. Also, mixers are not ideal linear devices. In addition to sum and difference frequencies, the nonlinear characteristics of the mixer will generate spectral components at frequencies of N*F (local osc.) $\pm$ M*F(RF) which will be fed into the bandpass filter. If these spurious components fall into the IF bandpass, they will be indistinguishable from actual input signals.

II. SYSTEM DESCRIPTION

In this section, a description of the signal processing necessary to eliminate the above mentioned problems will be discussed. As shown in the previous section, continued introduction of an arbitrary signal into the HP3577A, during the times the unit was not specifically tuned to the frequency of the signal, would lead to the display of internally-generated spurious signals. It is proposed to process the arbitrary input signals such that each individual frequency component will be fed into the HP3577A only during the time the unit is tuned to that particular frequency.

The concept can be visualized as a moving band-pass filter that tracks the HP3577A receiver as in Figure 8.

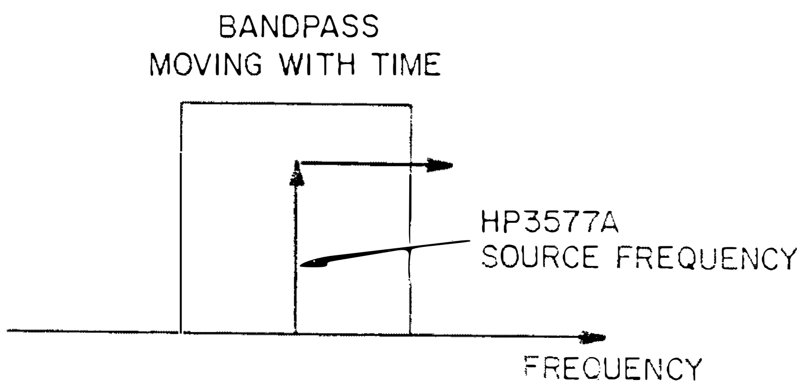


Figure 8. Tracking brickwall filter frequency response.

This filter, with moveable center frequency, will pass a particular frequency component when the HP3577A receiver is tuned to that frequency, allowing accurate measurement of the spectral component at each frequency. Otherwise, the component will be attenuated.

In order to illustrate the concept of the moving filter more clearly, Figures 9, 10, and 11, which contain actual performance data, are shown. Figure 9 shows, at one frequency, the type of filter that will be used. Notice the high selectivity and high out-band attenuation of the filter centered at 10 MHz. Figure 10 shows 10 superimposed traces in the range between 1 and 2 MHz. Each trace was taken with the effective center frequency 100 kHz above the previous trace. Figure 11 is of the same form as Figure 10 except that the range is between 1 and 10 MHz and the spacing is 1 MHz. The filter should follow the source frequency of the HP3577A, passing only the component in the 25 kHz band around the instantaneous source frequency.

The input signals will be divided up into two frequency ranges of interest: 5 Hz to 450 kHz and 450 kHz to 10 MHz. The design of the local oscillator generation system limits the filter response to about 450 kHz on the low end. Implementing the moving filter alone will not cover the frequency range from 5 Hz to 450 kHz. This problem is avoided by the fact that in the lower frequency range, the signal processing mentioned above, is not necessary. The design of the HP3577A is such that if there are no

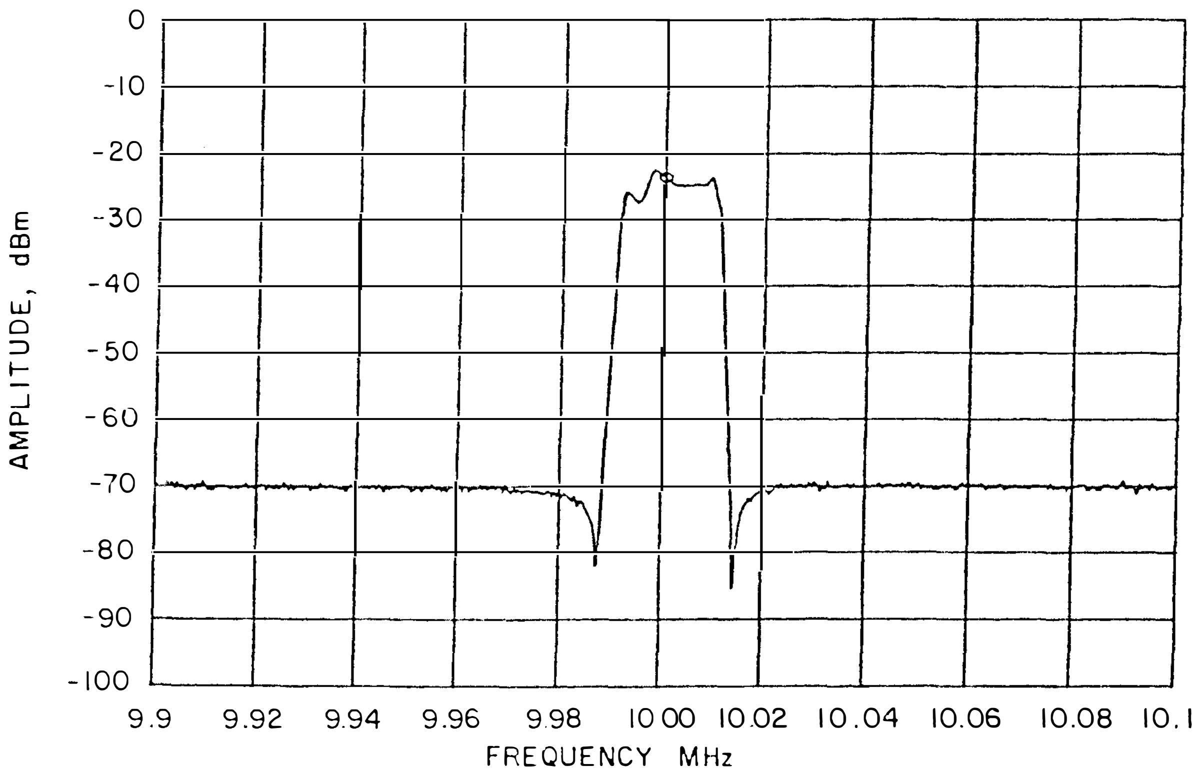


Figure 9. "Moving filter" with 10 MHz effective center frequency.

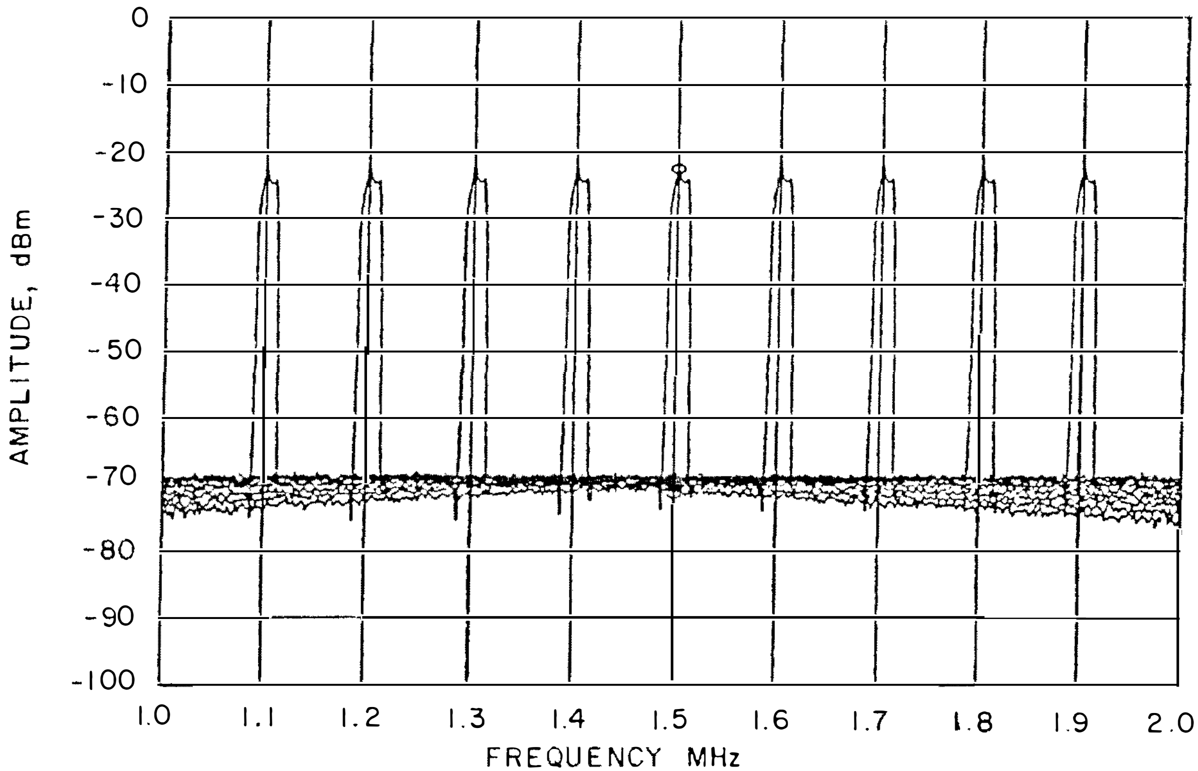


Figure 10. Ten superimposed frequency Bode plots from 1 to 2 MHz.

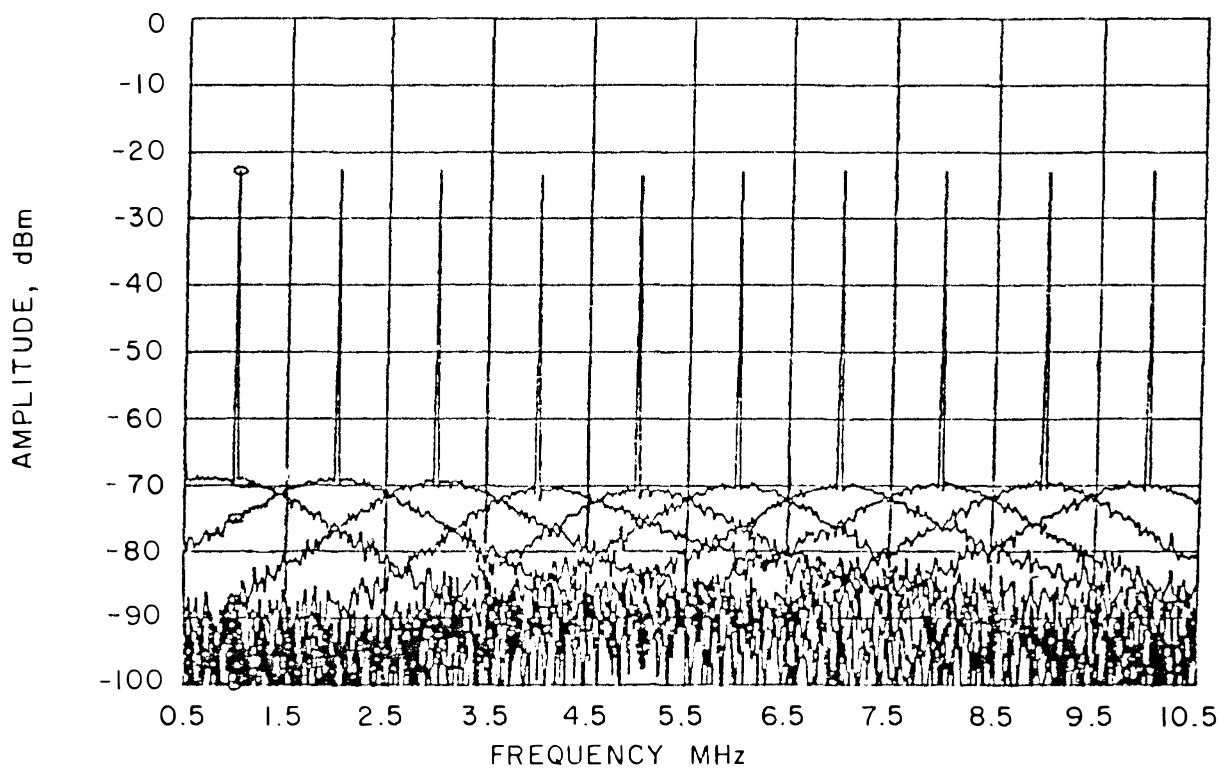


Figure 11. Ten Superimposed Frequency Bode Plots from 1 to 10 MHz.

frequency components in the arbitrary input signal above 500 kHz, the unit will operate alone with the same accuracy as the unit with the modification. In order to eliminate the frequencies above 500 kHz in the low frequency section of each channel, special filters were designed.

The block diagram of the modification is presented in Figure 12. Again, notice the three major system blocks. The function of each major system block is explained in the next three sections.

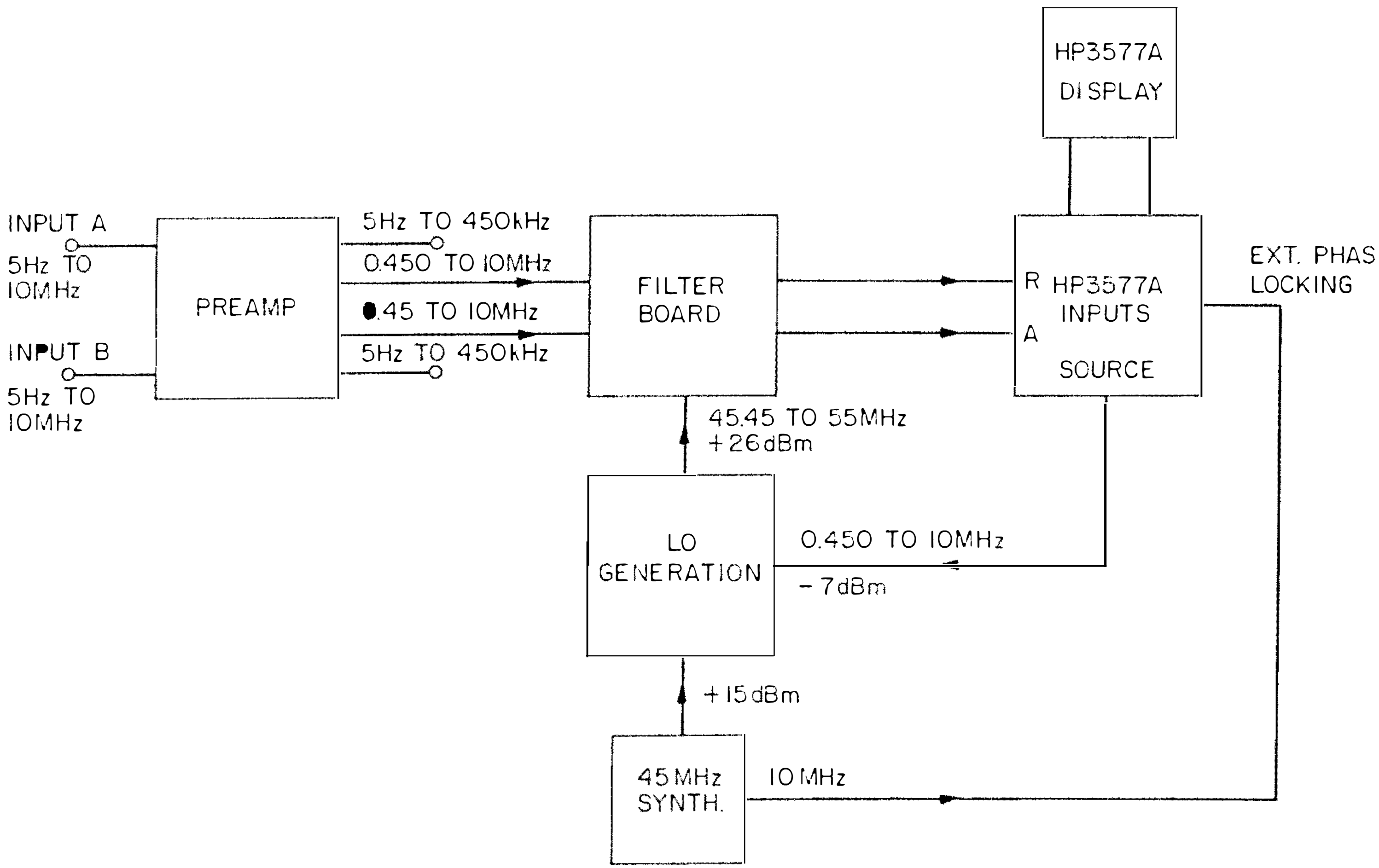


Figure 12. Conversion system block diagram.

III. MOVING FILTER

Introduction

The moving filter board is the heart of the conversion system. The functional purpose of the moving filter is to selectively filter the arbitrary signal input and pass only the desired components. A list of important specifications that the filter board should approach is given in Table 1. The moving filter block diagram is shown in Figure 13.

TABLE 1

FILTER BOARD SPECIFICATIONS

Frequency Range of Operation	450 kHz-10 MHz
Passband	$\pm$ 15 kHz F(instantaneous)
Attenuation	< 10-15 dB
Dynamic Range	0 dBm to -60 dBm
Impedance	50 ohms
Crosstalk	-50 dB
Out of Band Attenuation	> 45 dB
Output Spurious Response	> 40 dBc

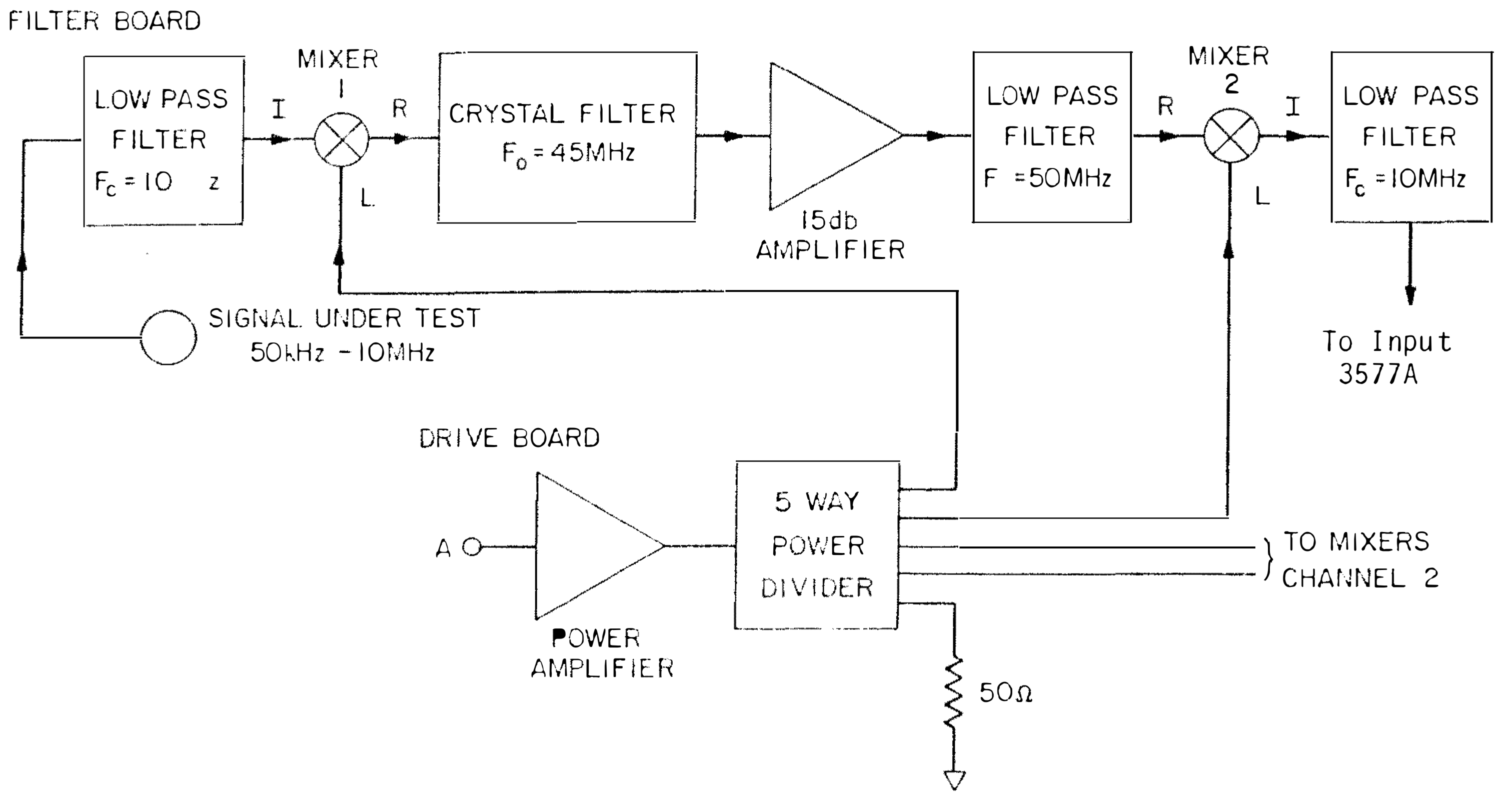


Figure 13. Filter board block diagram (1 channel).

Design

The moving filter can best be visualized as a bandpass-brickwall filter with variable center frequency. When one thinks of a bandpass-brickwall filter, a crystal filter comes to mind. However, a crystal filter has a set center frequency and cannot be varied. Ixers therefore are used to convert information from the frequency of interest to 45 MHz and back to the original frequency.

The principle of operation of the system is straightforward. All signals are mixed to 45 MHz, filtered by a crystal filter with a 25 kHz bandpass, then mixed back to the original frequency. This process passes the frequency components which satisfy the following equation:

$$F(\text{local oscillator}) - F(\text{spectral component}) = 45 \text{ MHz} . \qquad \text{Eq. 1}$$

It is easily seen by the preceding equation that by varying the local oscillator frequency, the effective passband of the system is controlled. During actual system operation the local oscillator remains at a frequency 45 MHz above the HP3577A source frequency.

The operation of the moving filter will now be discussed. The output of the preamplifier is fed into the input mixer's intermediate frequency port. This signal can be visualized as any spectrum that has been bandlimited to 10 MHz. When the signal is mixed up, it is centered around the local oscillator frequency. The local oscillator is set so that its frequency will remain 45 MHz above the instantaneous HP3577A source frequency. While the source

frequency sweeps from 450 kHz to 10 MHz, the local oscillator sweeps from 45.450 MHz to 55 MHz. The output from the input mixer's radio frequency port is fed into the crystal filter configuration with center frequency of 45 MHz. Any signal that passes the crystal filter configuration will be amplified and mixed back down to its original frequency.

An example of the moving filter's operation is now illustrated. Suppose a sine wave at a frequency of 2 MHz is fed from the preamplifier into the input mixer. Also, assume the HP3577A source frequency is sweeping from 1 MHz to 3 MHz. When the source frequency is at 1 MHz the local oscillator is at 46 MHz. The input mixer will produce sum and difference frequencies of 44 and 48 MHz. Neither of these signals will pass the crystal filter and no signal will be present to drive the second mixer. Indeed, no signal is exactly what the system should produce when the source frequency is at 1 MHz. The input signal is at 2 MHz and should be fed into the HP3577A only when the source frequency is at 2 MHz. Next, assume the source frequency is at 2 MHz. The local oscillator frequency will be 47 MHz. The input mixer will produce signals at 49 and 45 MHz. The 45 MHz signal will pass the crystal filter, be amplified, and then be mixed again. Since the local oscillator of the second mixer is still at 47 MHz, the output of the second mixer should have signals at 102 and 2 MHz. The HP3577A receiver is tuned to 2 MHz and will correctly measure the input 2 MHz sine wave. It is necessary to filter the output of the

second mixer to eliminate local oscillator feed through and all other high frequency components.

In the following, each of the components in the filter board is described and the design specification that led to the selection of each discussed. To minimize costs, standard crystal filters, mixers, and amplifiers were used.

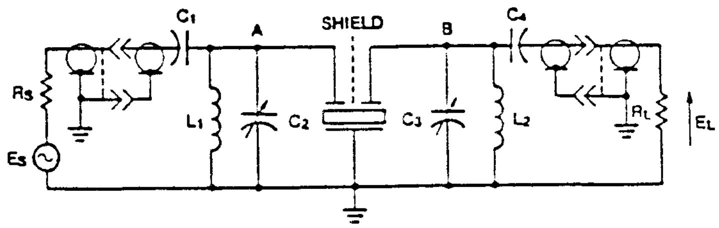
Crystal Filter

In the selection of a crystal filter, cost and frequency of operation were the main considerations. Crystal filters were readily available from Piezo Technology, Inc. (PTI) for $45 each. The standard values available were at 30 MHz, 45 MHz, and 75 MHz. The 45 MHz filter was chosen for two reasons: Fewer spurious signals enter the 450 kHz-10 MHz passband from the 45 MHz filter than the 30 MHz filter (this remark will be better understood after the mixers are discussed). Further, amplification of the 45 MHz signal is easier than the amplification of the 75 MHz signal. Also, fewer second order frequency effects occur with the 45 MHz system than with the 75 MHz system. For these reasons, the Model 4371-45 MHz crystal filters were used.

"The filter design utilizes four resonators on a single quartz wafter to provide four-pole response. These models offer wider passband values than two-pole inductorless tandem-set units" (ref. 3).

The input impedance of the filter is 7000 ohms, therefore, a matching network is necessary. This matching network, the design

of which was supplied by PTI, is shown in Figure 14.



C1, C4 = 5 pF

L1, L2 = 0.88 uH

C2, C3 = 1-10 pF.

Figure 14. PTI supplied matching network.

Source: Single-Wafer Four-Pole Monolithic Crystal Filters (Brochure), Piezo Technology, Inc., P. O. Box 7859, Orland, Florida 32854-7859, p. 4.

Figure 15 shows the PTI supplied attenuation-frequency plot of the Model 4371-45 MHz crystal filter. While this plot is very impressive, it does not tell the whole story. Note that the frequency bounds of the plot are 60 kHz below and 140 kHz above the center frequency. At greater deviations from the center frequency the attenuation decreases to about 25 dB. Also reducing the filter's out-band attenuation is the varying load presented to the filters by the mixers. The varying local oscillator frequency driving the mixers changes their output impedance. Changing the load on the crystal filter reduces the out-band attenuation, since the matching networks are tuned for maximum out-band attenuation at a single impedance. Taking both these characteristics into consideration,

it was necessary to cascade three crystal filters to obtain 45 dB out-band attenuation for the entire frequency range of interest. The matching network was modified and the final configuration is shown in the filter board schematic (Appendix A).

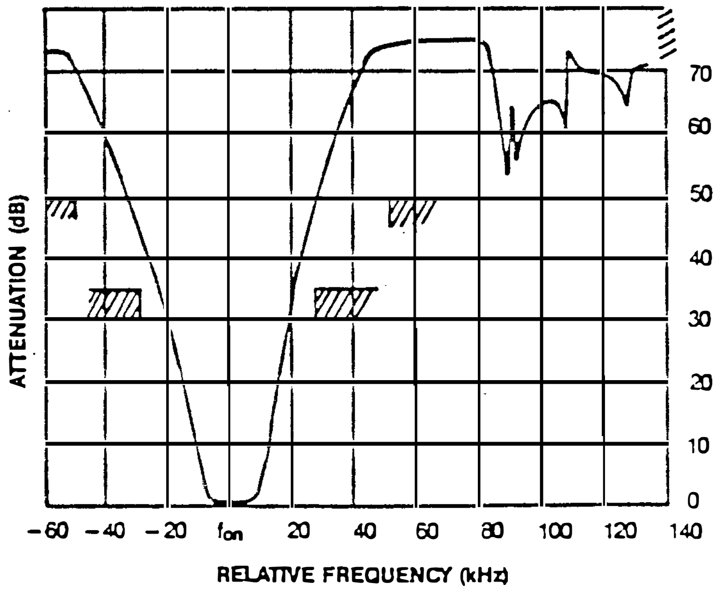


Figure 15. Amplitude characteristics of 4371 crystal filter.

Source: Single-Wafer Four-Pole Monolithic Crystal Filters (Brochure), Piezo Technology, Inc., P. O. Box 7859, Orlando, Florida 332845-7859, p. 3.

The frequency response achieved with this system is shown in Figure 16. The out-of-band attenuation of this system varied between 45-85 dB across the frequency band. This out-of-band attenuation is a limiting factor for the dynamic range of the instrument.

A few practical points should now be discussed. The variation of the variable capacitors in the matching network has a marked

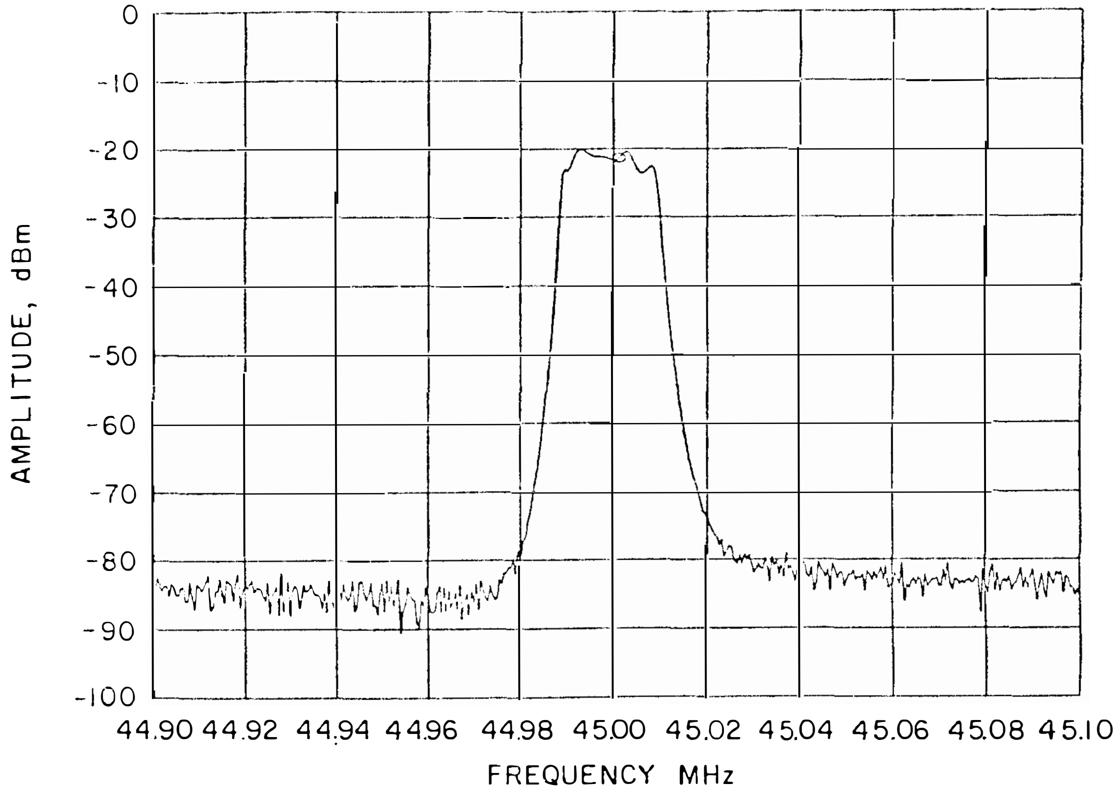


Figure 16. Actual cascaded-crystal filter response.

effect on the performance of the filter. It is possible, while covering the entire range of the capacitors, to transform the system from a workable filter with 10 dB loss and 2 dB in-band ripple to a system which will have 50 dB loss and 35 dB ripple. It is necessary to use high quality high-Q air dielectric capacitors.

To achieve the desired out-of-band rejection it was necessary to place a grounded metal plane 1/4 inch below the filter combination. This plane grounds out fringe fields that were produced by the inevitable impedance mismatch.

While the filter has flat amplitude characteristics, the phase shift across the bandpass is large (500 degrees). This phase shift and its effect on system performance will be discussed later.

Mixers

In each of the two channels of the filter board, two mixers are used. One mixer mixes the signal up in frequency, while the other mixes the signal down in frequency. The mixers should satisfy the following constraints: First, the frequency range of each of the three mixer ports must include the frequencies that will be introduced to that port (RF = 45 MHz, Local Oscillator = 45 MHz to 55 MHz, and IF = 450 kHz to 10 MHz). Second, the mixers must satisfy the desired amplitude dynamic range from 0 dBm to -80 dBm. Third, it is necessary that the spurious responses generated by the mixers be as low as possible. Fourth, the cost of the mixers must not be excessive.

A description of the operation of the mixers used is now in order. The mixers used in the conversion system are shown in Figure 17. This is a classic double-balanced mixer that suppresses the local oscillator frequency component from entering either the IF or RF port.

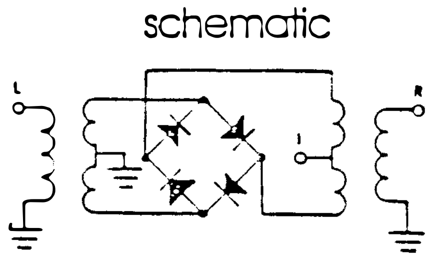


Figure 17. Double balanced mixer schematic.

Source: RF/IF Signal Processing Handbook, Mini-Circuits, Inc., P. O. Box 166, Brooklyn, New York 11235, p. 33, 1985-1986.

In Section II, the spurious generation properties of mixers were briefly discussed. However, the dependence of the mixer operation on the local oscillator was not mentioned. The maximum input amplitude and purity of the local oscillator signal are very important system parameters. The amplitude of the local oscillator determines the maximum signal that can be fed into the RF or IF ports while maintaining linearity. Also, the local oscillator has much to do with the spurious output components. For a given IF or RF signal level, the level of spuriously generated response will be inversely proportional to the local oscillator level. That is, the higher the local oscillator power level, the lower the spurious responses. This is best illustrated in Table 2.

TABLE 2

MIXER HARMONIC LEVELS FOR TWO MIXERS WITH DIFFERENT DRIVE LEVELS

mixer harmonic intermodulation

(relative to desired IF output)

RF ORDER HARMONIC (rows) / Harmonic LO Order (columns)

RF ORDER HARMONIC	RF CAL	0	1	2	3	4	5	6	7	8	9	10
0	-	-	4	30	16	36	25	41	48	57	53	69
1	-	16	-	30	13	36	32	57	61	58	51	68
2	75	38	32	42	29	45	36	53	45	59	57	68
3	95	44	42	48	43	60	41	56	63	66	64	67
4	96	72	57	51	48	60	48	53	47	60	70	73
5	94	73	69	65	59	53	54	59	59	64	69	80
6	>94	79	>83	>86	79	74	59	61	57	64	67	77
7	>96	>84	83	>83	>84	84	>85	80	67	73	67	73
8	>94	>84	>84	>84	>82	>84	80	74	73	78	69	73
9	>94	>83	>83	>84	>83	>83	>84	>84	81	82	78	>85
10	>93	>83	>83	>84	86	>84	>82	>84	>85	>84	82	81

test conditions RF - IN FREQ 500 1 MHZ DRIVE - 4 06 DBM
LO - IN FREQ 470 01 MHZ DRIVE 7 DBM
IF - MEASURED FREQ 30 09 MHZ AMP - 10 46 DBM

RF ORDER HARMONIC	RF CAL	0	1	2	3	4	5	6	7	8	9	10
0	>90	-	10	20	8	51	17	29	37	41	44	49
1	>90	17	-	37	10	34	46	48	28	47	50	48
2	78	60	46	60	47	71	47	71	53	60	60	75
3	>90	73	68	68	65	70	57	77	72	70	60	77
4	>90	77	>82	81	>81	>81	>80	>81	>83	>82	77	77
5	>90	77	77	>82	>82	>81	>81	>80	>81	>83	>82	77
6	>90	76	78	77	>82	>82	>81	>81	>80	>81	>82	>82
7	>90	78	77	77	77	>82	>82	>81	>81	>80	>81	>82
8	>90	77	78	78	76	78	>82	>82	>81	>81	>81	>81
9	>90	77	78	78	77	78	77	>82	>82	>81	>80	>81
10	>89	78	77	77	78	77	77	76	>82	>82	>81	>80

test conditions RF 500 100 MHZ INPUT P -4 94DBM
LO 470 010 MHZ INPUT P + 16 96DBM
IF - 30 090MHZ IF AMPLIT - 11 88 DBM

SOURCE: Ibid., pp. 164-165.

Table 2 contains spurious response tables for two mixers that use different local oscillator levels but have the same RF level. Notice that the components of interest (inside the lines) for the mixer with a 17 dBm local oscillator (level 17) have much lower levels than the mixer with a 7 dBm local oscillator (level 7). Note that the frequency components of interest are the 2x2, 3x3, etc. These may fall in the output bandpass of interest.

Another form of distortion, lower in the level 17 mixers, is two-tone intermodulation distortion. Two-tone intermodulation occurs when two components of a spectrum (f1 and f2), fed into mixers RF or IF port, mix together to produce an output spectral component. The frequency of this spurious response will occur at a frequency defined by the relation $((2*f1 \pm f2) \pm f(\text{local osc}))$. All mixers present some degree of two-tone distortion. The two-tone distortion level in the level 17 mixers will be 50 dB below the amplitude of the smaller of the two spectral components mixing together, the level 7 only 30 dB.

The mixers chosen for the project were the Mini-circuits TFM-3. These mixers have the specifications shown in Table 3.

This component fits the described requirements well. The maximum input level is well above the desired input specifications. The frequency dynamic range is much greater than needed. The local oscillator to IF isolation should be sufficient, so that with additional filtering its effect on system performance should be small.

TABLE 3

MINI-CIRCUITS MODEL TFM-1 CHARACTERISTICS

Frequency Response:	
LO/RF	0.1 to 250 MHz
IF	DC to 250 MHz
LO Signal Level	17 dBm
Maximum Input Level	14 dBm
LO to IF Isolation	> 25 dB
LO to RF Isolation	> 30 dB
Impedance	50 ohms
Conversion Loss	6 dB
Cost	$45

SOURCE: RF/IF Signal Processing Handbook, Mini-Circuits, Inc., P. O. Box 166, Brooklyn, New York 11235, pp. 164-165, 1985-1986.

Mixer performance specifications are contingent on proper loading of all ports and adequate local oscillator level. If the load on the RF or IF ports is not the desired 50 ohms, the levels shown in Table 2 will go up. Even if the mismatch does not occur in the band of interest, it can affect in-band performance. For example, directly cascading a filter after a mixer can cause a system to display the above mentioned effect. Suppose the base band of interest extends to 10 MHz and signals above are filtered with a multiple pole low-pass filter. The in-band impedance of

the filter is probably close to 50 ohms, but the out-band impedance may be far away from 50 ohms. It was observed that the mixer's third harmonic suppression was greatly reduced when loaded directly with a filter. The out-band mismatch of the crystal filter configuration reduced harmonic suppression. However, the change in the impedance of the crystal filter configuration is not as radical as an LC filter. If the level of the local oscillator signal is not adequate, several undesirable effects arise. First, the conversion loss increases as the local oscillator signal level decreases from the recommended level. If the level of the local oscillator signal on the level 17 mixers drops from 17 to 10 dBm, the conversion loss increases from 6 to 11 dB. Further reduction in local oscillator level increases conversion loss exponentially. Second, reducing local oscillator level reduces the suppression of spurious outputs.

Finally, it is essential that the packaging of the mixer follows good practice. If the mixer is improperly grounded, the local oscillator to RF isolation and spurious generation specifications will be compromised. Correct grounding, for a mixer packaged in an 8-pin dip can, means that the case itself must be soldered to the ground plane. It was necessary to use a ground plane on both sides of the filter board, and then to solder the mixer case directly to the board.

Amplifier

The purpose of the amplifier in each channel is to prevent the signal to noise ratio reduction of the system from becoming too high. The sum of all losses in the channel is approximately 30 dB (excluding the amplifier). The amplifier has a gain of 15 dB. Cascading the amplifier in line with the rest of the system results in a system that has a 15 dB reduction in signal to noise level. The input noise level of the amplifier is very small (-110 dBm) and does not affect system performance.

The amplifier used was obtained from government surplus. The Avantek UTA-108 has the specifications listed in Table 4.

TABLE 4

AVANTEK MODEL UTA-108 AMPLIFIER

Gain	15 dB
Frequency Response	5 MHz to 500 MHz
1 DB Compression Point	3 dBm
Supply Voltage	15 Volts
Signal to Noise Ratio	13 dB
Impedance	50 ohms

The amplifier was driven very close to the 1 dB compression point. The 1 dB compression point is defined by the output level where the gain has been reduced by 1 dB. Due to nonlinearity, the

energy will go into the harmonics instead of the fundamental. As the output level approaches the 1 dB compression point, the level of the harmonics greatly increases. Fortunately, these harmonics are located around 90 and 135 MHz, far away from the band of interest.

Output Filters

The filter cascaded with the amplifier is a low-pass filter with a cut-off frequency of 50 MHz. This filter eliminates higher frequency components generated by the input mixer and the amplifier.

Since the output of mixer 2 will have many spurious components outside the 450 kHz to 10 MHz band of interest, it is necessary to filter out these components so that they do not create spurious signals in the final system. A four section standard low-pass filter was used. The attenuation at frequencies above 50 MHz is 60 dB, which effectively eliminates high frequency signals. The VSWR of this filter is typically 1.5 in band and 17 out of band.

The output filters have the same effect on mixer performance as described in the mixer section. In this case, the higher level of harmonics, generated by an improperly loaded output mixer, has little effect. Suppose a 1 MHz signal is filtered by the modification and fed into the HP3577A receiver when the receiver is tuned to 1 MHz (normal operation). The third harmonic will be at 3 MHz. Since the receiver is not tuned to 3 MHz, the spurious response

should not affect the measurement of the 1 MHz signal greatly. When the receiver frequency reaches 3 MHz the 45 MHz frequency driving the output mixer is attenuated so that its voltage level is negligible, so there will be no 3 MHz harmonic to measure.

Experimental Results

The most graphic experimental results for the filter board are shown previously in Section II. These show the effective filter passband being varied across the band of interest. The passband width is 25 kHz and the ripple is 4 dB. The frequency range of the filter is DC to 10 MHz. The amplitude dynamic range of the filter is 0 dBm to -100 dBm. The -100 dBm specification is for an ideal local oscillator and is not obtained with the current local oscillator generation system. The outband attenuation is greater than 45 dB across the 450 kHz to 10 MHz band. The harmonics are greater than 40 dB below the fundamental.

IV. PREAMPLIFIER

Introduction

The two-channel, four-output preamplifier is used to isolate the outside measurement devices from the mixers of the adjustable filter. Specific features include the following: The preamplifier should provide proper 50 ohm loading for outside measurement devices as well as mixers on the adjustable filter. The preamplifier should divide signals into two frequency bands of interest from 5 Hz to 450 kHz and 450 kHz to 10 MHz. The system should be band limited to 10 MHz and DC should be blocked on the high-frequency channel. The low-frequency section of each channel should have a sharp rolloff so that no signals of frequency greater than 500 kHz pass into the HP3577A. The noise level of the amplifier should be lower than the noise level of the filter board. The maximum input level should be greater than -6 dBm. Harmonic distortion for the system should result in harmonics with power levels at least 40 dB below the maximum power output over the entire range of interest.

Design

The block diagram for the preamplifier is shown in Figure 18. The majority of the system is composed of operational amplifiers and standard components. A design using discrete devices was employed for the composite filters. The operational amplifiers

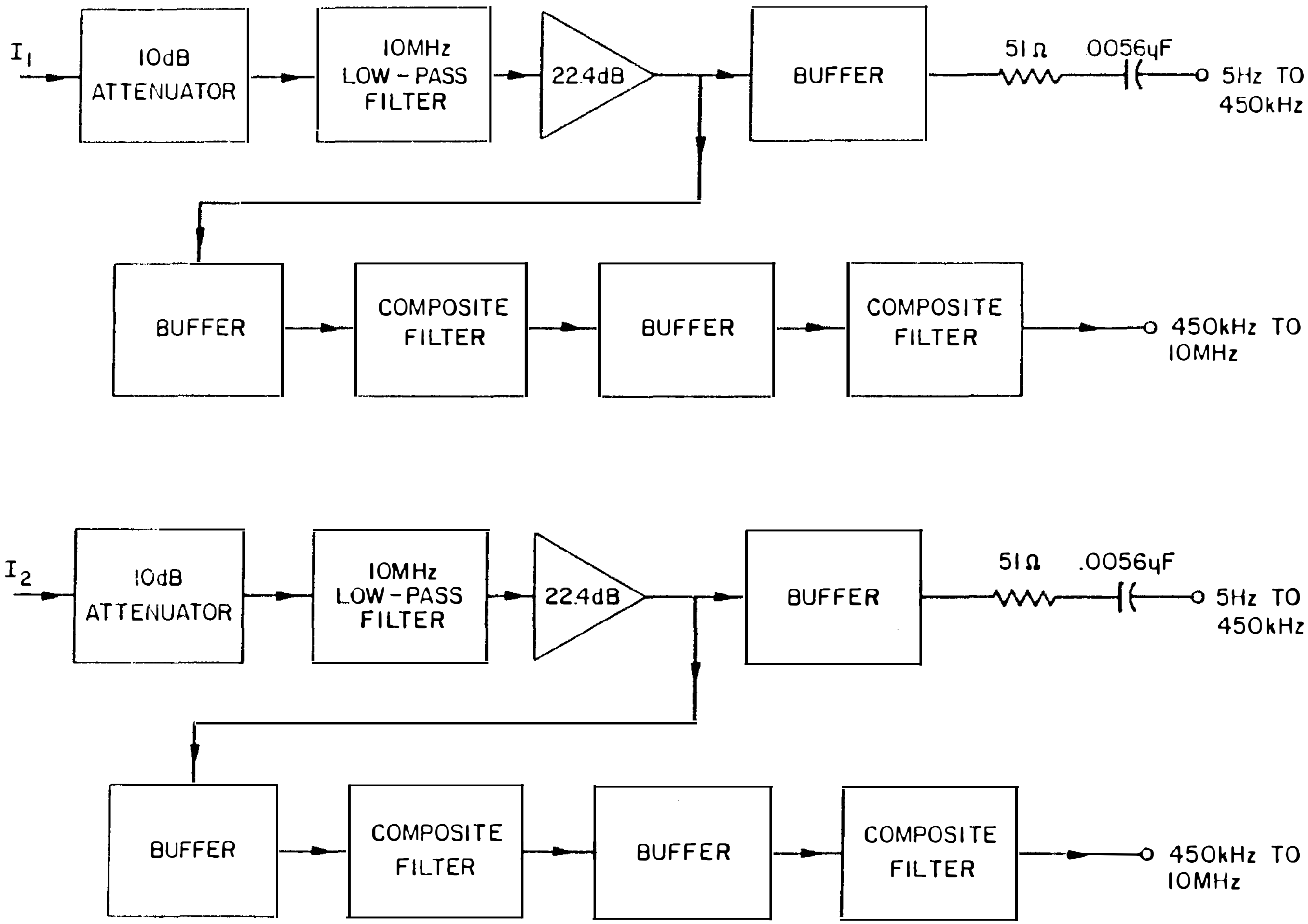


Figure 18. Preamplifier block diagram.

used were the Harris Semiconductor models 2539 and 5002.

The HA2539 is a device with a 600 MHz gain bandwidth product. One HA2539 was used in each channel with a linear gain of 13.3 or 22.5 dB. Since the gain bandwidth is 600 MHz, the 13.5 gain allows a bandwidth of 44 MHz. From a chart of phase shift versus loop transmission, the phase margin was determined to be approximately 75 degrees. This 75 degree phase margin was needed because of the susceptibility of the op-amp to oscillate. The slew rate was specified to be 600 V/uS. At 10 MHz maximum frequency the maximum amplitude should be 9.5 volts or 29.6 dBm. The noise level of the HA2539 is specified at -113 dBm in a 1 kHz bandwidth. Since the HA2539 was incapable of driving 50 ohm loads, the HA5002 was used as an impedance matching mechanism. The HA5002 is a 110 MHz buffer amplifier capable of driving 50 ohm loads. The maximum output current for the HA5002 is 200 mA, well above the 20 mA needed to drive 1 volt into 50 ohms. The noise level for the HA5002 was approximated at -100 dBm in a 1 kHz bandwidth.

Voltage regulators were used on each channel to isolate the amplifiers from power line noise. Many bypass capacitors were connected from power pins to ground to decouple the amplifiers from one another.

The input of the amplifier consists of a 10 dB attenuator and a 10 MHz low-pass filter. The attenuator provides a constant 50 ohm load with a USWR of less than 1.5. The filter has a sharp rolloff after 10.7 MHz which effectively attenuates high-frequency

signal components which might be misinterpreted. The output of the HA5002 contains a series RC network. The resistor is 50 ohms which provides an in-band match for the mixers of the adjustable filter. The capacitor was a 5600 pF component which effectively blocked DC without affecting in-band performance.

The filters which were designed to eliminate signals of frequencies greater than 450 kHz in the low-frequency section of each channel are "composite filters." These filters were used for several purposes in the thesis project. The design concept and design equations are presented in the local oscillator generation section.

Experimental Results

The preamplifier met all minimum specifications. However, the system was difficult to stabilize against oscillations. The HA2539 chip was inclined to oscillate when the circuit's physical layout was not perfect. It was necessary to connect bypass capacitors directly to power pins. The 9.5 MHz specification for the full power bandwidth of the HA2539 was very misleading. This specification was derived and not measured. Full power bandwidth should mean maximum frequency at which a maximum output sine wave can be fed into a load with some specified minimal distortion. Harris Semiconductor derived the full power bandwidth using the slewrate of a square wave in a calculation to obtain their 9.5 MHz for the full power bandwidth. The distortion of the waveform at

amplitudes greater than 0 dBm at 10 MHz was unacceptable. The distortion was in the form of slewing. At 10 MHz with an output of 0 dBm the harmonics were approximately 40 dB below the fundamental. Figure 19 contains data of harmonics at 5 MHz and 10 MHz.

Figure 20 contains a Bode plot of the low-frequency section of each channel. Notice the extremely sharp cut off above 400 kHz. The crosstalk between the low-frequency sections was very low.

Figure 21 contains a Bode plot of the high frequency section of each channel. The crosstalk between the channels is asymmetrical, with values of -43 dB from left to right and -50 dB from right to left.

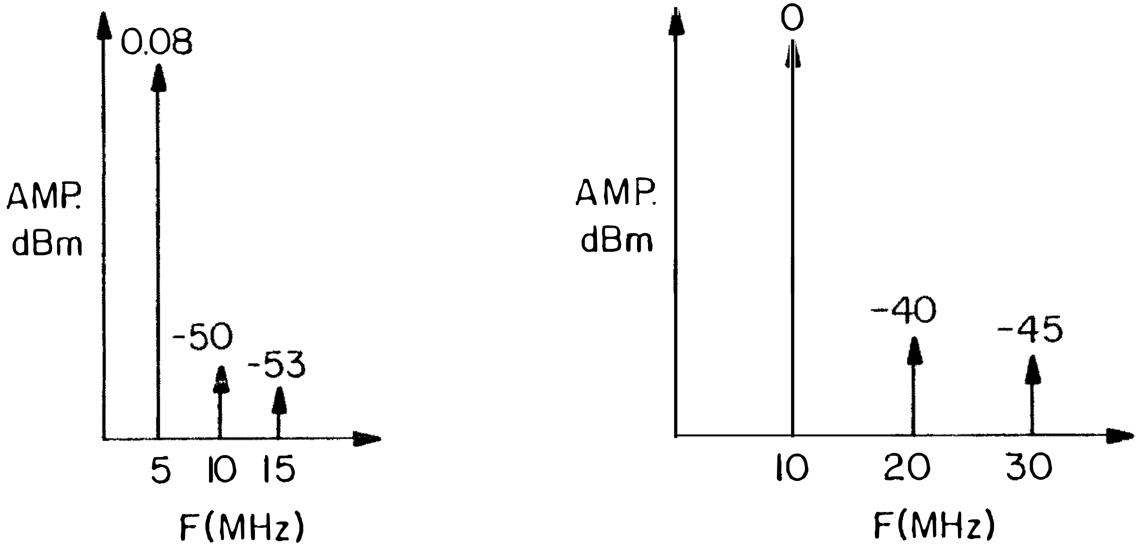


Figure 19. Preamplifier output for fundamentals at 5 MHz and 10 MHz.

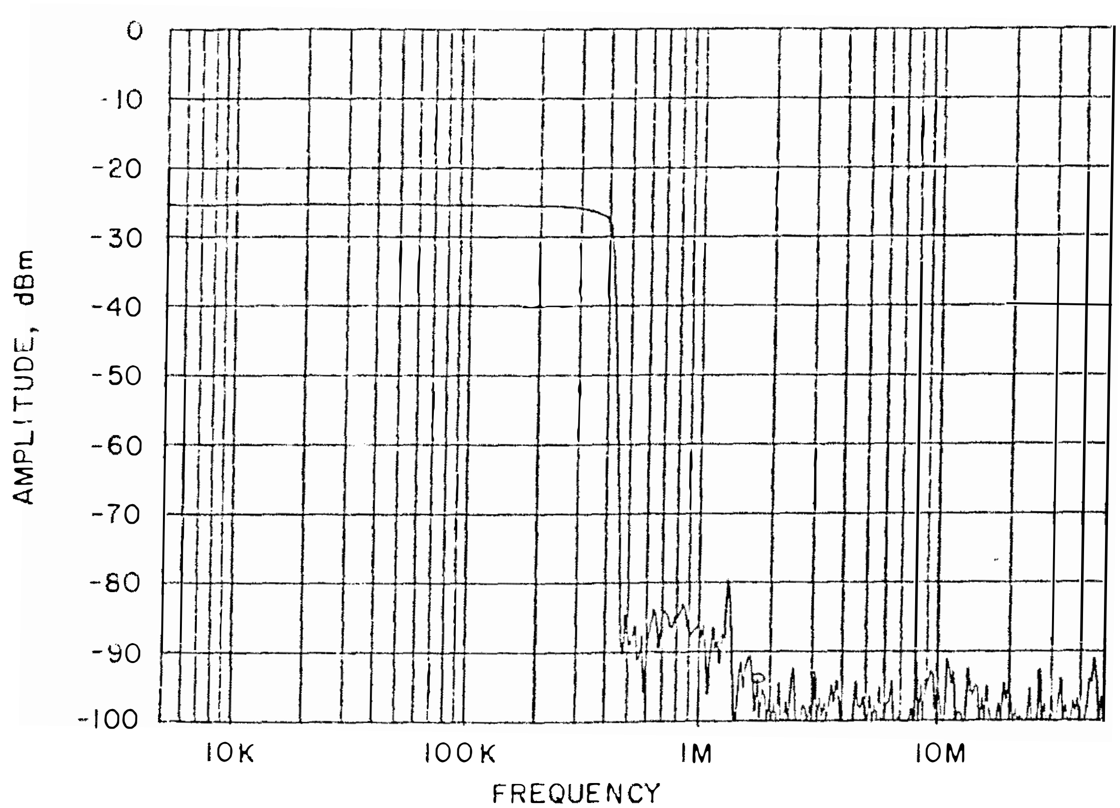


Figure 20. Bode plot of low-frequency section.

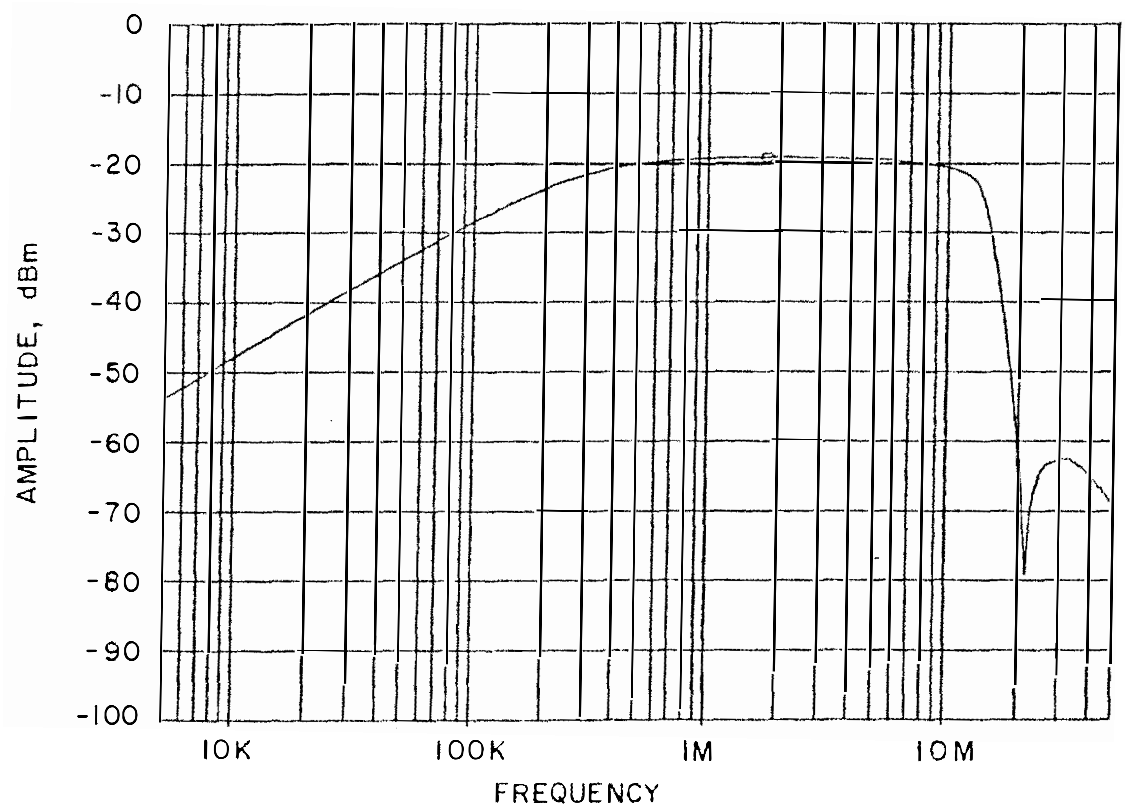


Figure 21. Bode plot of high-frequency section.

V. LOCAL OSCILLATOR GENERATION SYSTEM

Introduction

The final part of the HP3577A modification is the local oscillator generation system. In the preceding sections dealing with the operation of the moving filter, it was assumed that the local oscillator driving the filter board mixers would track the source frequency of the HP3577A with a 45 MHz offset. This is not completely straightforward in this application. Conceptually, it is desired to add 45 MHz to the source frequency. Since the addition of frequencies can be accomplished by mixing, mixing the HP3577A source frequency with a 45 MHz signal will generate the desired local oscillator. Unfortunately, the mixing process generates many other frequency components besides the sum frequency. The mixer output must be filtered to reduce these undesired frequency components to acceptable levels.

The local oscillator should meet the following specifications: The output signal should track the HP3577A source with a 45 MHz offset. The frequency bounds of the local oscillator should extend from 45.45 MHz to 55 MHz. Any spurious signals should have power levels at least 40 dB below the power in the fundamental. The ripple in the amplitude should be less than 5 dB. The nominal power output of the local oscillator generation system is 26 dBm. The power of the output signal must be 26 dBm

due to power division in the filter board. The power level at the local oscillator port of each mixer in the moving filter will be 7 dB less than the output of the local oscillator generation system.

The remainder of Section V contains the following: Local oscillator generation system design, experimental results, and the effects of undesirable spectral components on system performance. In addition, composite filtering, which played an important role in this thesis, will be discussed.

Design

The block diagram of the local oscillator generation system is shown in Figure 22. The local oscillator generation system mixes together the HP3577A source frequency and a synthesized 45 MHz signal. The mixer's output is filtered to pass only the sum frequency. The remaining signal is then amplified to the proper level.

The amplifier used for the power amplification to the 26 dBm level was a Mini-Circuit ZHL-32A. The maximum output power of the ZHL-32A is 30 dBm. The frequency range of this device is between 50 kHz and 130 MHz. The gain of the ZHL-32A is 25 dBm.

The mixer used was again Mini-Circuits's TFM-3 mixer. The reasons for this selection are the same as explained in the mixer

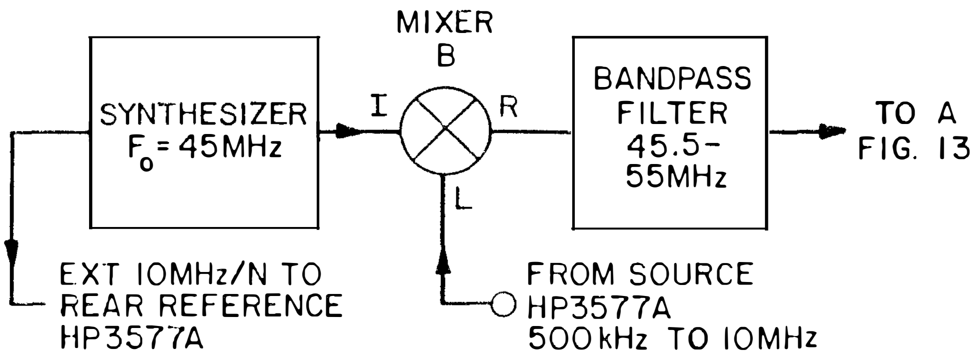


Figure 22. Local oscillator generation.

subsection of Section III. The higher level local oscillator results in lower harmonic content in the output. The TFM-3 mixer's output contains three undesired spectral components that must be eliminated. These include the difference frequency, the second harmonic, and the third harmonic of the local oscillator. The power level of the input signal driving the mixer will be +15 dBm for the 45 MHz local oscillator and -7 dBm for the HP3577A source.

The filters necessary to eliminate the difference frequency and the second harmonic of the local oscillator are not straightforward. At the lower limit of the system's operation (source frequency = 450 kHz), the difference frequency is only 900 kHz below the desired sum frequency. In the 45 MHz region, 900 kHz is a very short frequency space. Standard filters could not eliminate the difference frequency without affecting the sum frequency. The type of filter used for this operation is called a "composite filter." The composite filters for the local oscillator generation system were divided into two parts, cascaded high-pass and low-pass filters. An amplifier buffer was cascaded between the two filters to provide isolation and gain.

Composite Filters

Composite filters were developed many years ago to help eliminate crosstalk between channels on multiple carrier telephone lines. The basic concepts needed to understand the operation of

composite filters are these: Two-part networks can be designed such that loading the network's output with a particular impedance will result in that impedance being measured at the input. This impedance is called the image impedance of the network. Many such networks with the same image impedance can be cascaded without affecting the transfer function of any individual two-part component. Imaged matched LC circuits containing series and parallel tank circuits can be designed. Series LC circuits and parallel LC circuits have special characteristics at resonance. In the case of a series LC circuit, a short circuit is obtained at the resonance frequency; for the parallel, an open circuit.

In the design of composite filters, image matched high and low-pass two ports are cascaded with image matched networks containing LC tank circuits. The resonance frequencies of these tank circuits are outside the bandpass of the filter. At resonance frequencies, the signal is either blocked or shorted, preventing the signal from reaching the output. The high or low-pass networks eliminate frequencies far away from the band pass. The LC circuits are used to eliminate particular frequencies. Since resonance is a narrow band phenomenon, the rolloff in the resonance region is sharp. These tank circuits can be placed just outside the passband of interest to achieve rapid rolloffs. Since the response of the tank circuits is very narrow band, several tank circuits across the reject band may be necessary. These extra networks prevent the total filter's response from rising in the

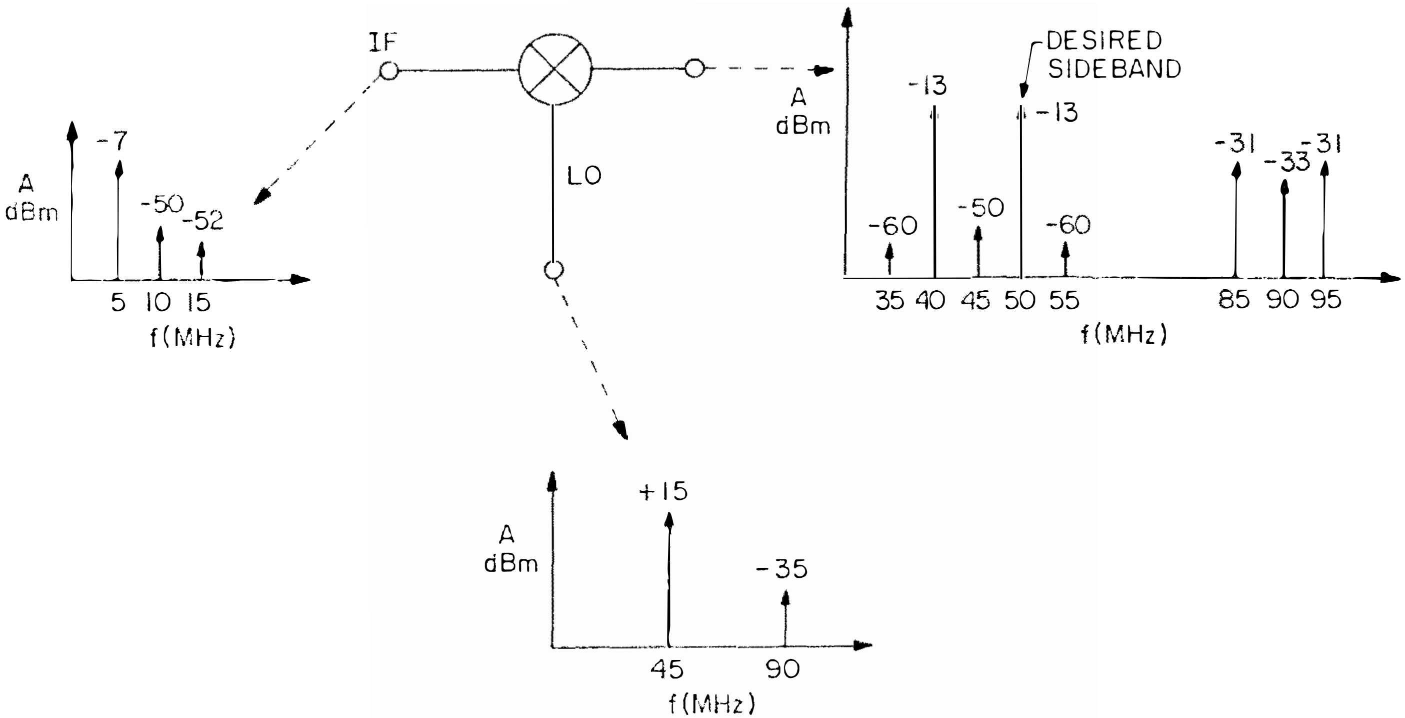
reject band. The design equations for the composite filters are shown in Appendix C.

Three composite filters were designed and integrated into the modification. The first, contained in the preamplifier, was designed to have an extremely sharp rolloff about 450 kHz. The results for this filter are shown in Figure 20, page 46. The second, is the high-pass filter in the local oscillator generation system that passes signals above 45 MHz and rejects signals below. The third is a low-pass filter which filters out unwanted higher harmonics generated by the mixer in the local oscillator generation system.

Experimental Results

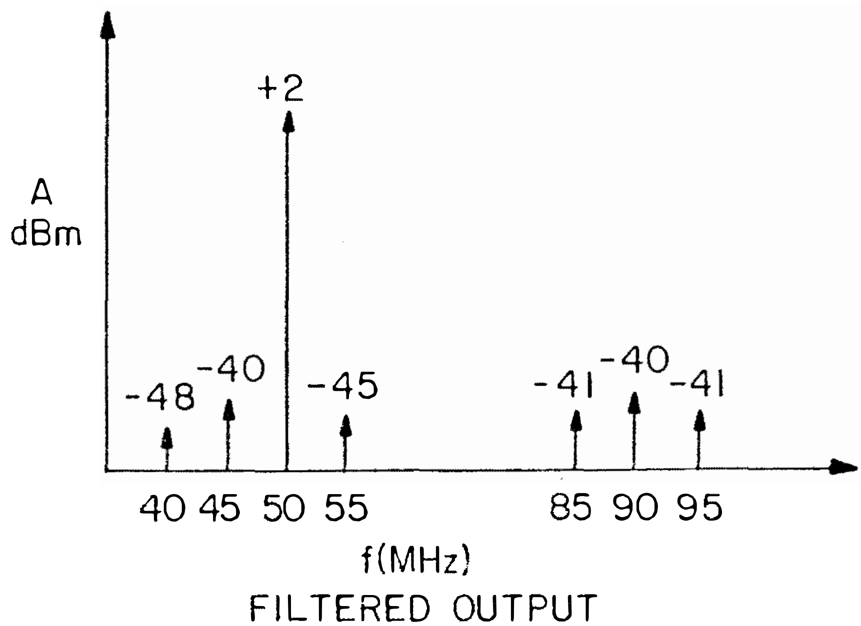
The mixer's input and output signals are shown in Figure 23a. Note the many spurious signals in the mixer's output. This spectrum is then fed into the bandpass filter. The high and low-pass filters contained in the local oscillator generation system are shown in Figure 24. The composite filter response for the high-pass low-pass configuration is shown in Figure 25.

The Bode plot in Figure 25 is the most important factor affecting the performance of the entire modification. Even though the rolloff at 45 MHz is sharp, it is not ideal. Therefore, the ideal goal of eliminating the difference frequency and not affecting the sum frequency is not obtained for frequencies very close to 45 MHz. Figure 23b is the spectrum obtained from the local



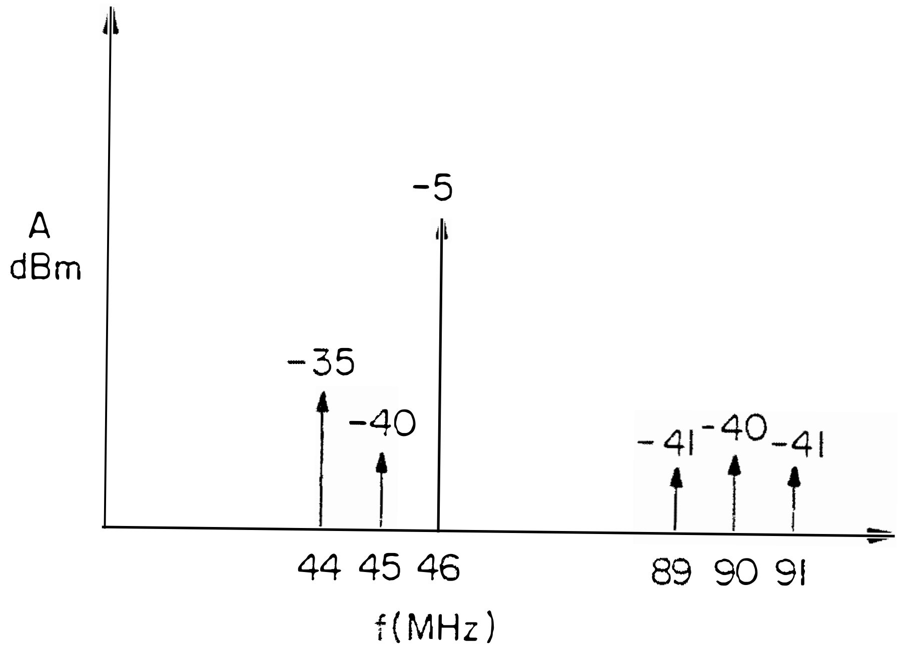
(a) 45 MHz Local Oscillator Spectrum

Figure 23. Local oscillator generation operation for 5 MHz.



(b) Local oscillator generation spectrum after band-pass filter.

Figure 23 (continued).



(c) Local oscillator generation spectrum after band-pass filter for 1 MHz source frequency.

Figure 23 (continued).

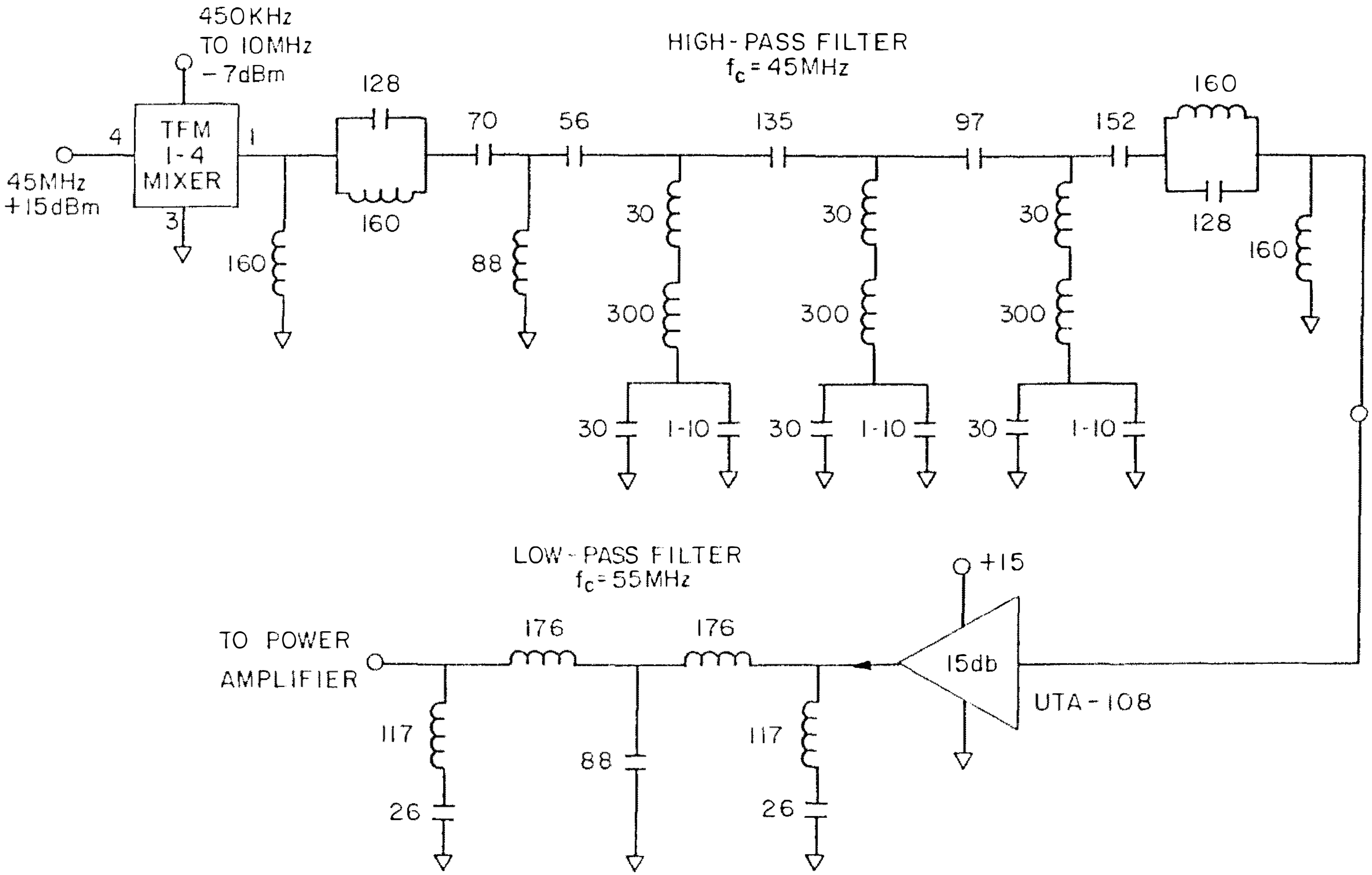


Figure 24. Local oscillator generation system schematic.

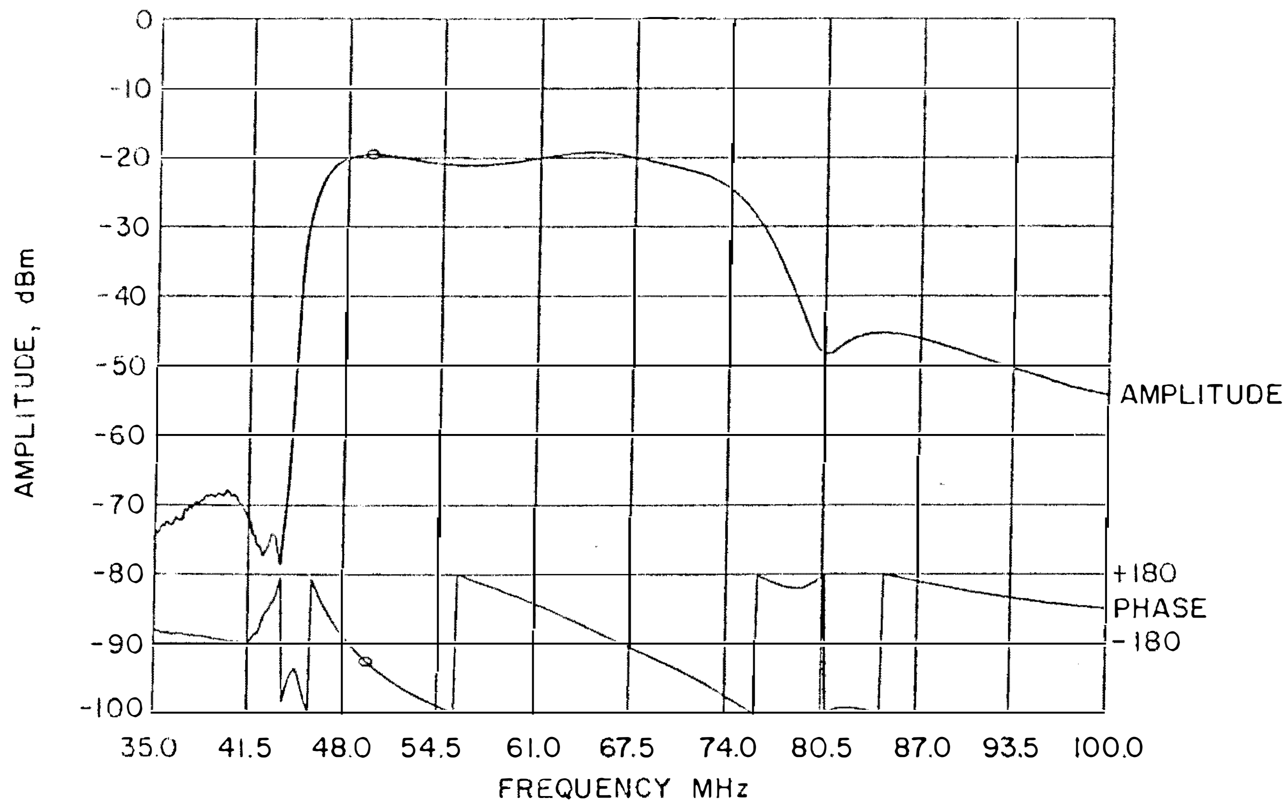


Figure 25. Bode plot of 45.45 MHz to 55 MHz bandpass filter.

oscillator generation system when the HP3577A source frequency is 5 MHz. Notice the excellent spectral purity. Figure 23c is the spectrum obtained when the source frequency is 1 MHz. Not only is the sum frequency attenuated, the difference frequency is at a higher level. This results in lower spectral purity.

Local Oscillator Specifications:

Nominal Amplitude	26 dBm
Frequency Range	450 kHz to 10 MHz
Spurious Signals (1.5 MHz to 10 MHz)	40 dB below fundamental
Spurious Signals (450 KHz to 1.5 MHz)	30 dB below fundamental

Local oscillator fundamental frequency tracks HP3577A source with 45 MHz offset.

The effect of the local oscillator's spectral purity on the performance of a mixer must now be discussed. The relationship between local oscillator spectral purity and the resulting mixer output is complex. A mixer is not a two part. Therefore, spurious signals and harmonics contained in the local oscillator spectrum do not linearly affect the output. This phenomonon is best demonstrated with an example. Suppose a square-wave local oscillator (frequency F1) is used to mix with a sine wave (frequency F2) in the mixing process. A square wave is composed of many odd harmonics in addition to the fundamental. However, when the mixer's output is examined, only the sum and difference frequencies (F1 $\pm$ F2) are present. Why don't the harmonics of the square wave produce an effect? The reason is because a diode mixer,

a bi-phase modulator, changes the phase of the IF signal 180 degrees each time the sign of the local oscillator changes. Since the operation of the local oscillator is dependent on sign change alone, the harmonics of the square wave, which add together, have no effect. As a rule of thumb, unless a local oscillator spurious component is a harmonic of the fundamental, a spurious response will be present in the mixer's output. The amplitude and frequency relationship between the desired mixer output and the spurious response will be the same as that of the actual local oscillator signal to the spurious signal in the local oscillator.

VI. EXPERIMENTAL RESULTS

Introduction

This section discusses the experimental results for the entire system. First, the results of the system are compared with the HP3577A alone. Second, accuracy, harmonic generation, linearity, and calibration of the system are examined. Third, the ability of the system to make phase coherency measurements is discussed. Fourth, the system dynamic range is discussed and limitations explained. Fifth, crosstalk between channels is presented. As a final example, the frequency spectrum of a square wave is measured and the results compared to a spectrum analyzer's results (Appendix B).

Figures 3-6, pages 10-13, show the results of directly feeding a 4 MHz sine wave into the HP3577A. Figures 26-29 show the HP3577A display when the 4 MHz sine wave is first processed by the modification. Notice the tremendous reduction in the image response in Figure 27 as compared to Figure 4, page 11. The level of the other spurious signals found in the low frequency range has also been greatly lowered.

The noise levels at the edges of the Figures 27-36 traces looks as if they rise. This is not the case. This phenomonon is due to calibration. When the system is calibrated over the entire 450 kHz to 10 MHz range, the video trace is stored in the HP3577A

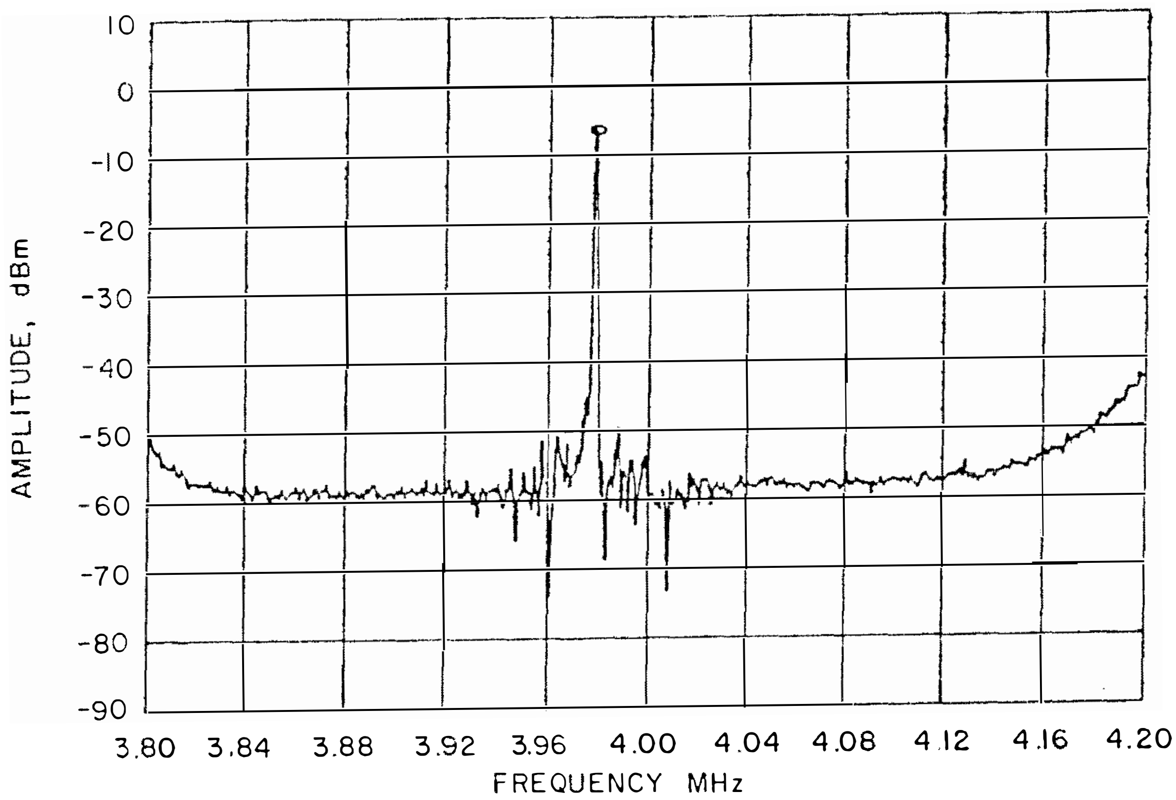


Figure 26. HP3577A display for a 4 MHz sine wave with system modification.

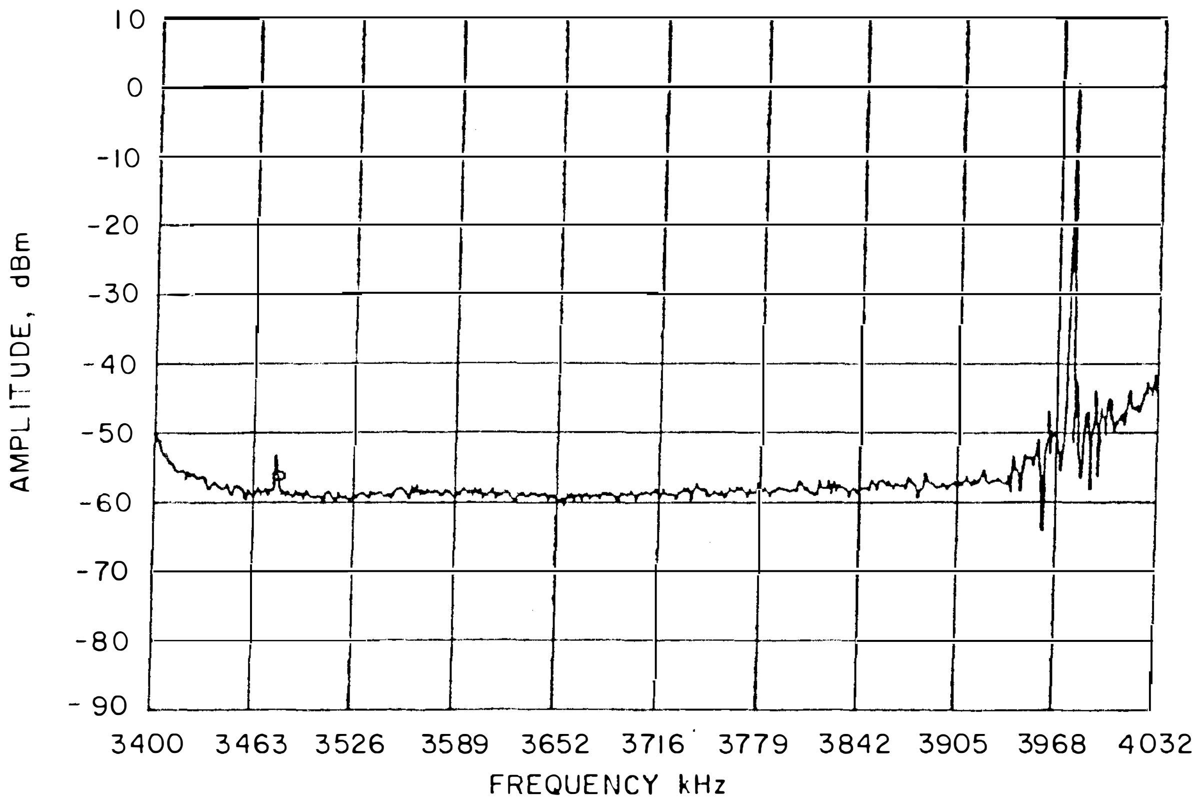


Figure 27. HP3577A display showing attenuation of image signal.

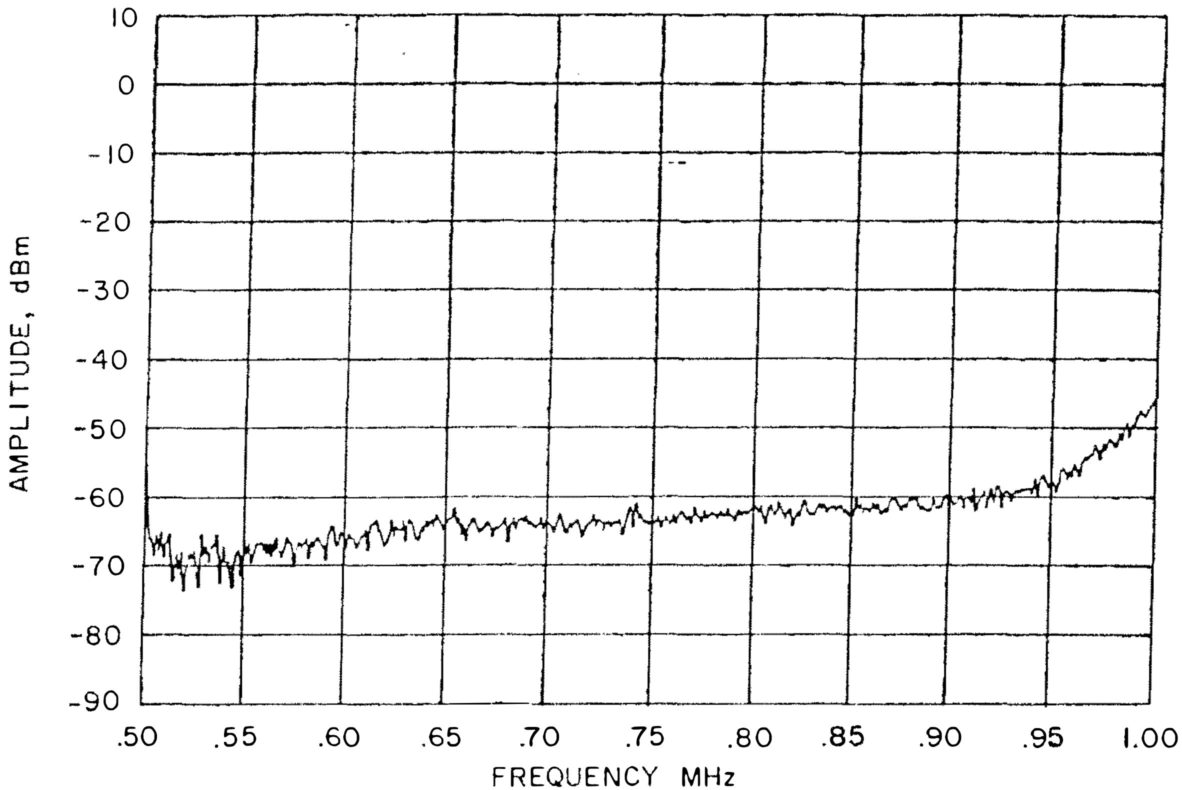


Figure 28. HP3577A display showing lowered spurious responses.

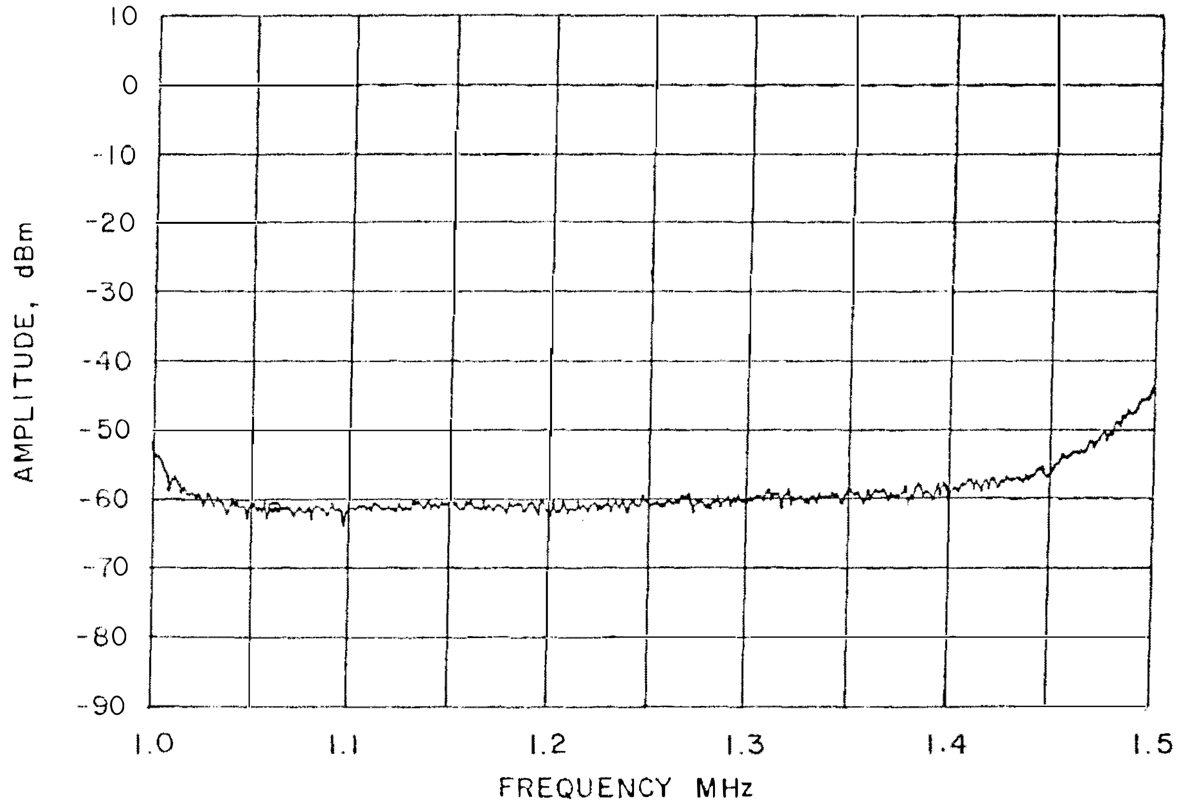


Figure 29. HP3577A display showing lower spurious responses.

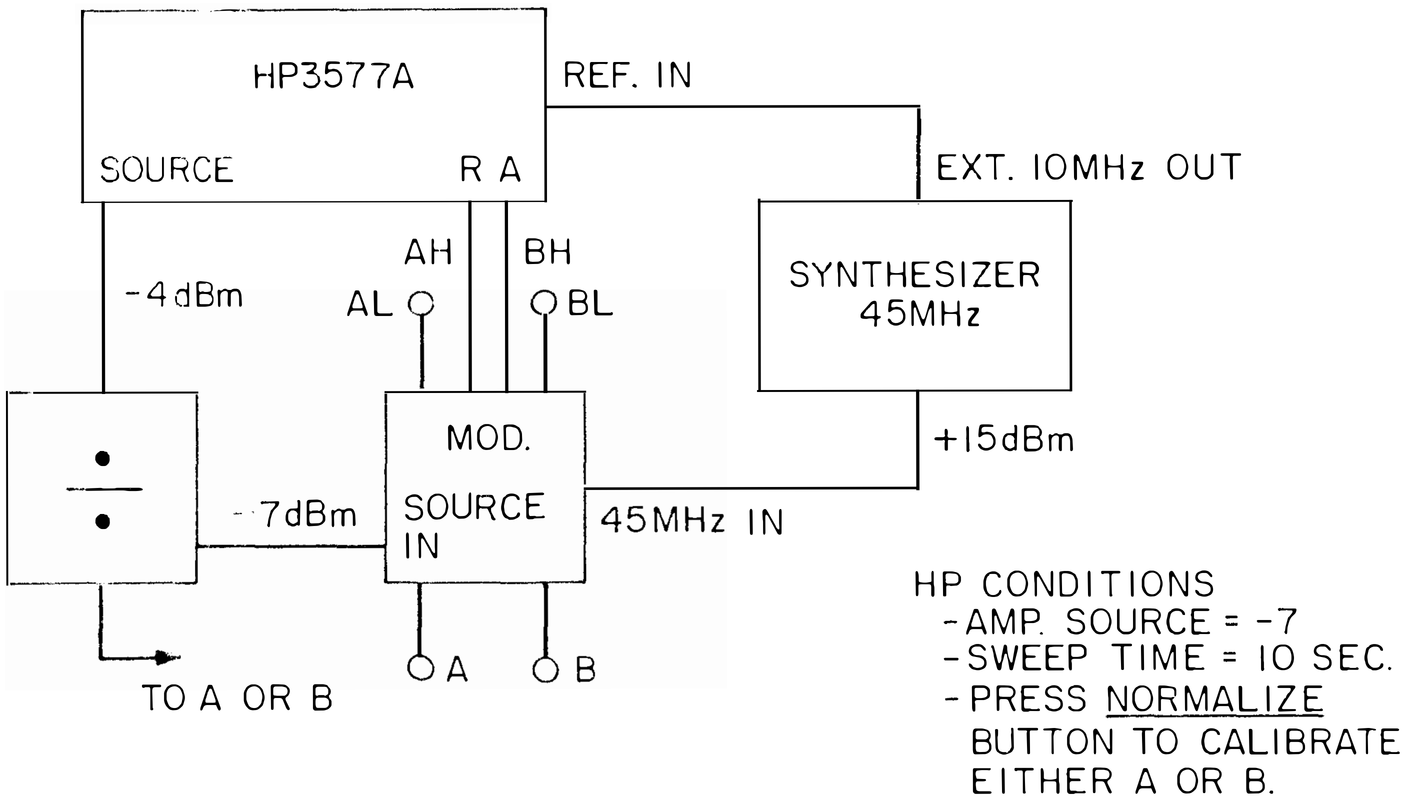


Figure 30. Calibration system block diagram.

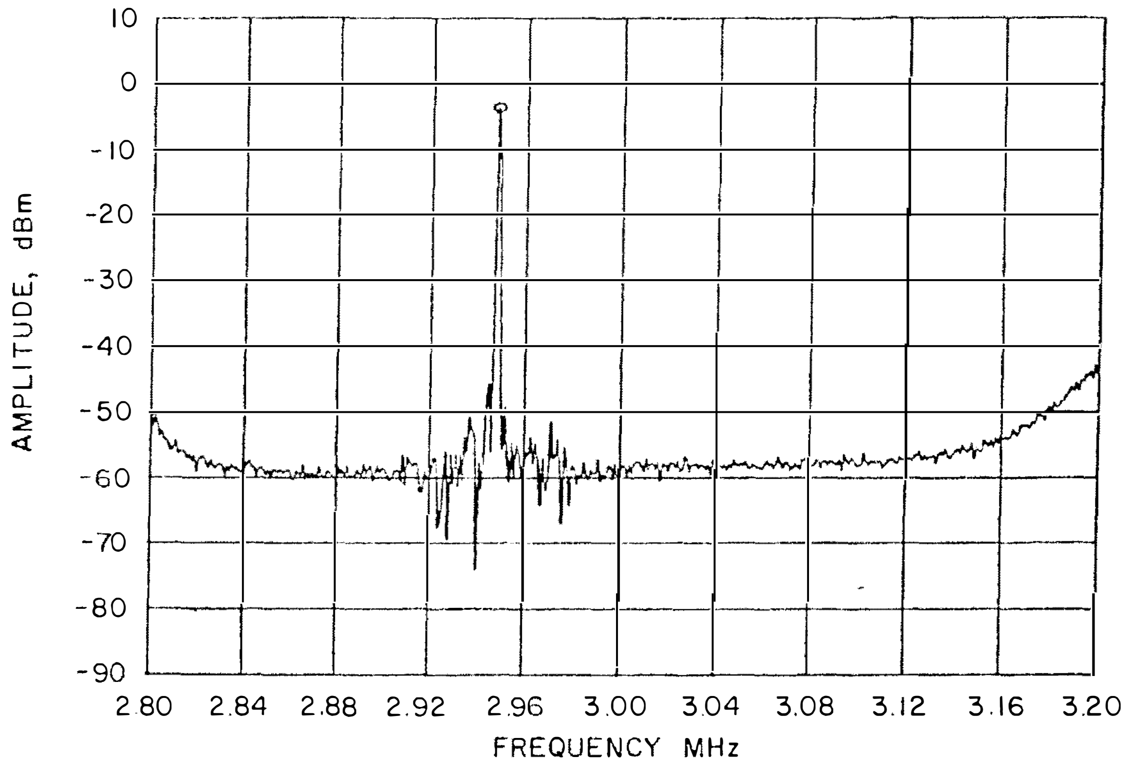


Figure 31. 2.954 MHz sine wave fundamental.

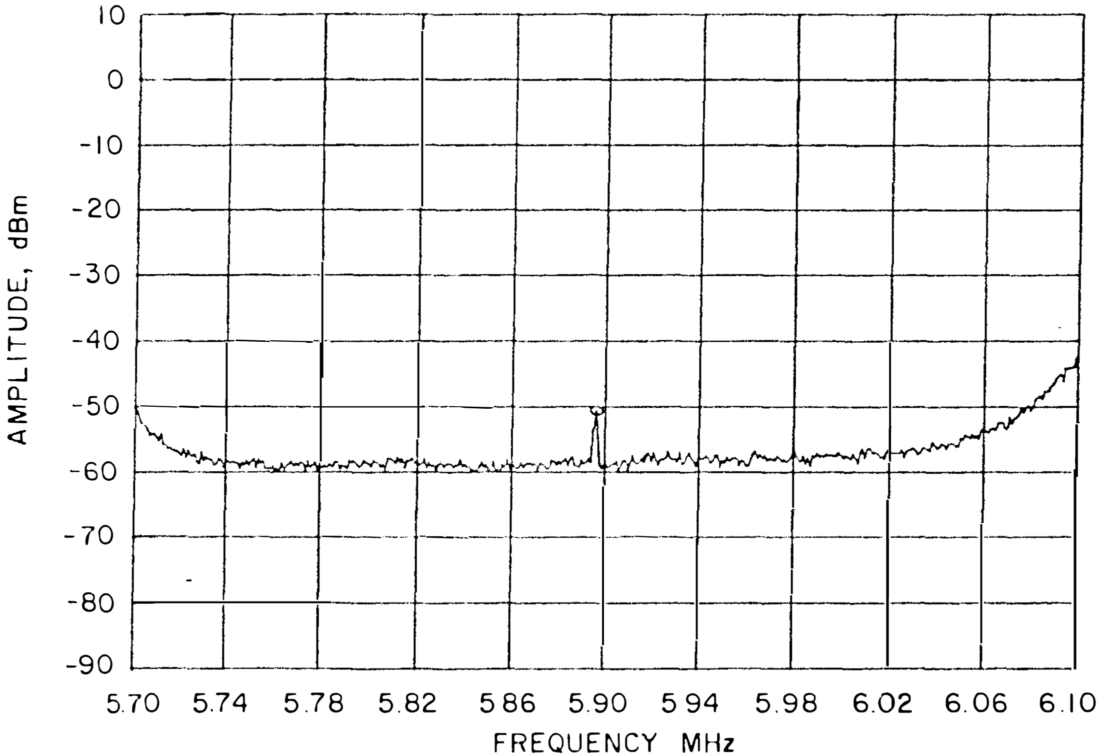


Figure 32. Example of second harmonic (5.89 MHz).

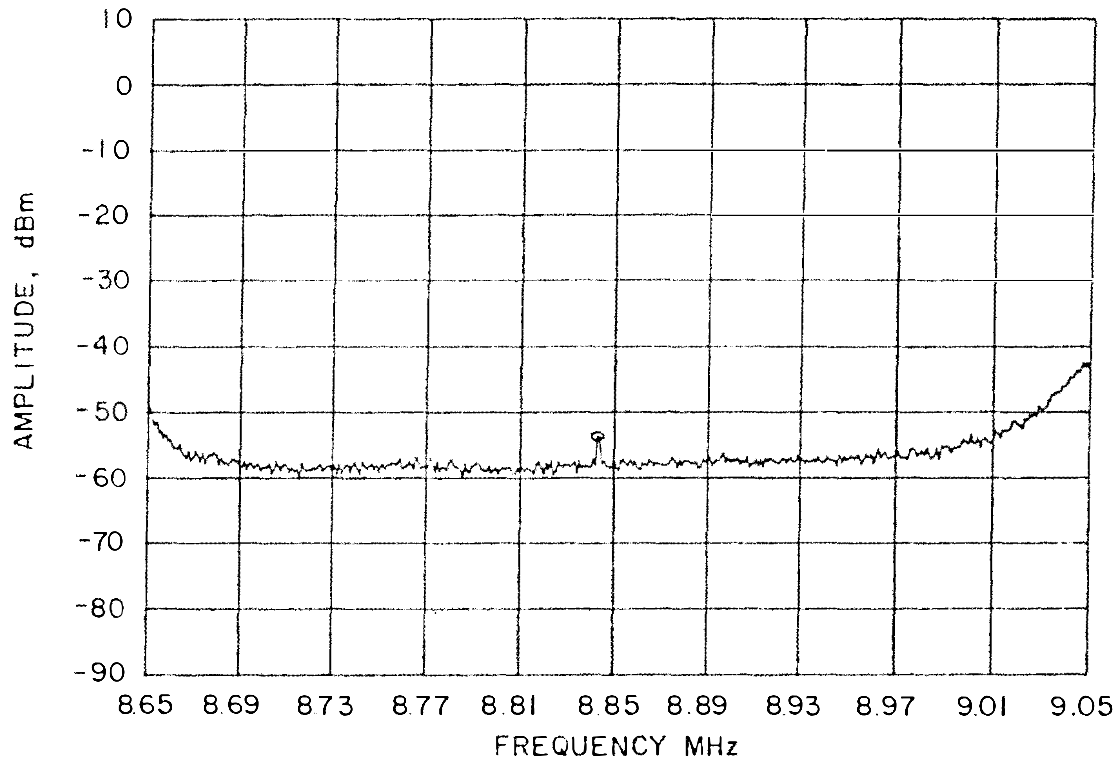


Figure 33. Example of third harmonic (8.84 MHz).

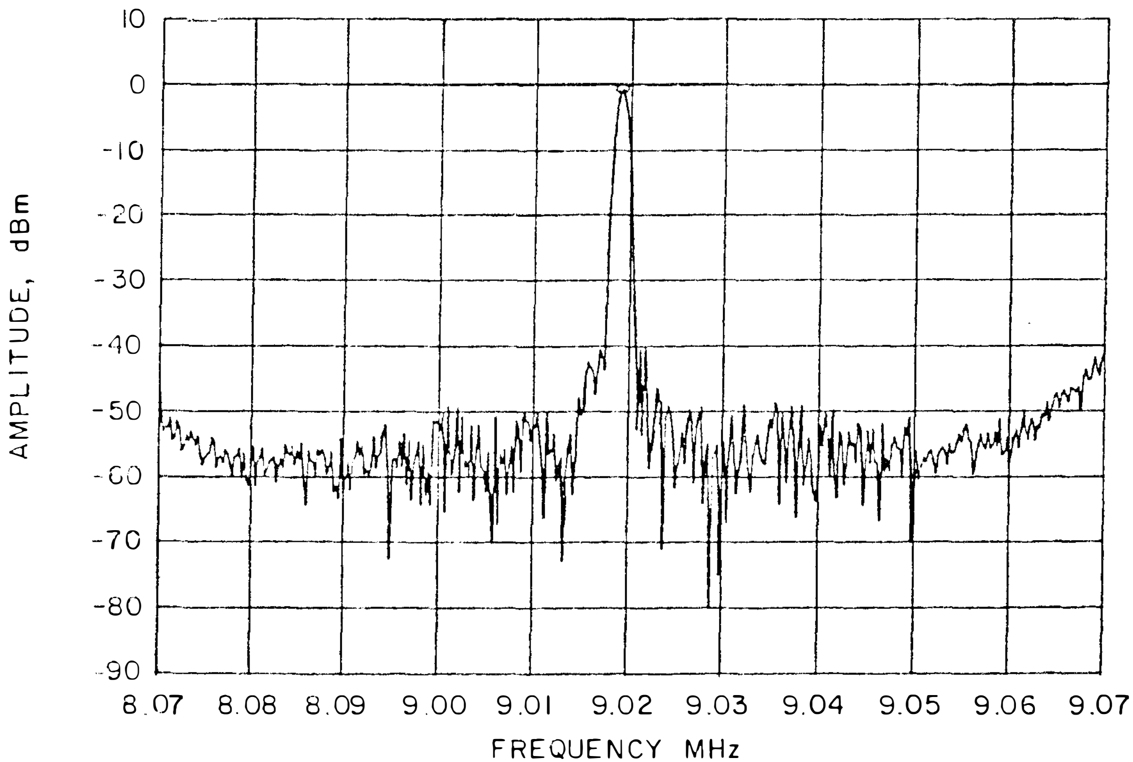


Figure 34. 9.015 MHz sine wave with 0 dB attenuation.

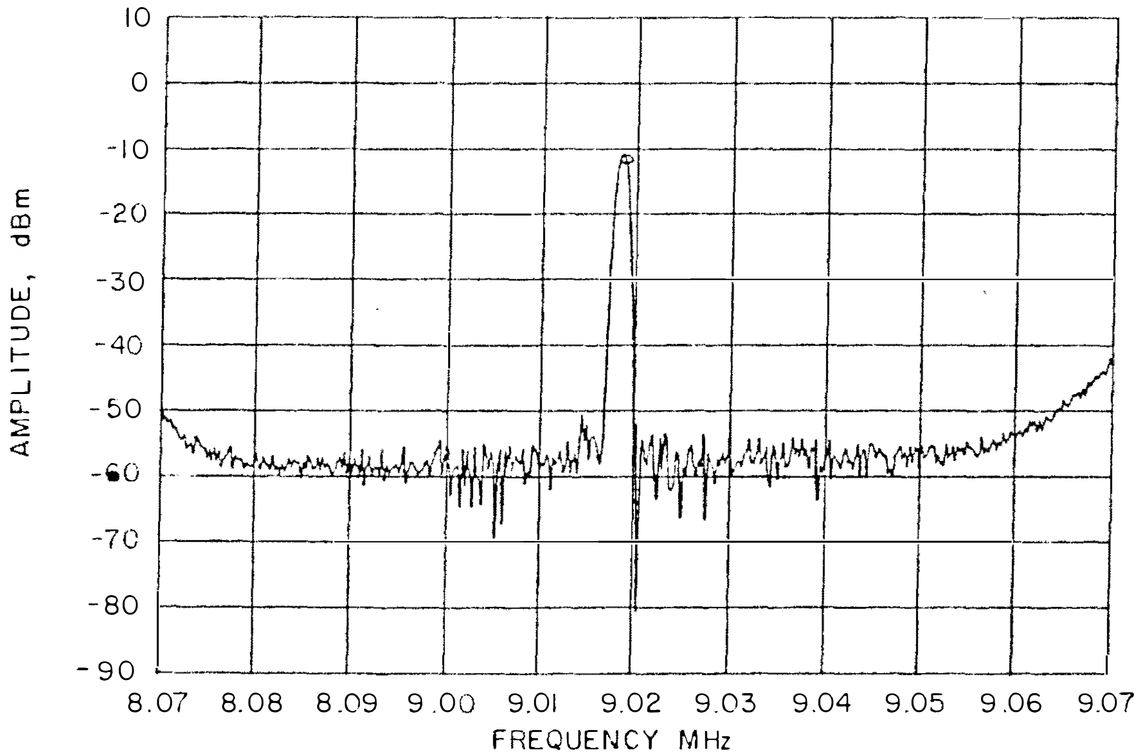


Figure 35. 9.015 MHz sine wave with 10 dB attenuation.

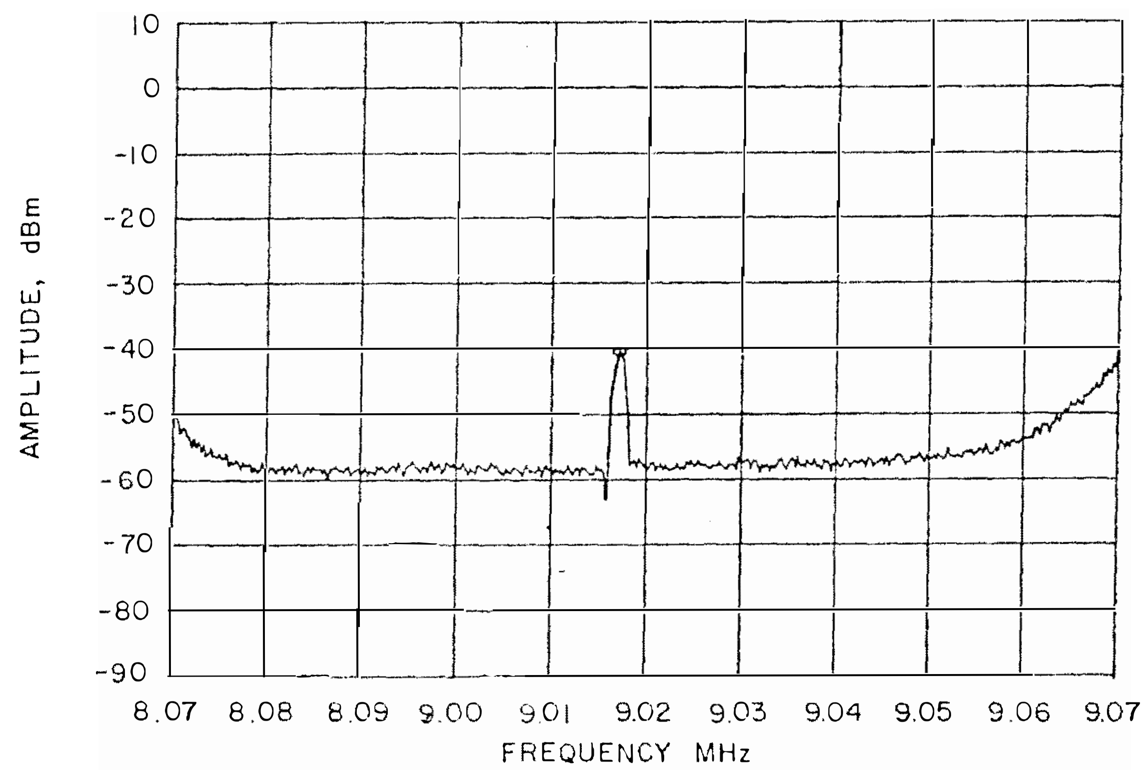


Figure 36. 9.015 MHz sine wave with 40 dB attenuation.

memory. When the frequency band is changed, the trace modification is maintained as before. Hence, it is necessary to recalibrate the system over each bandwidth to eliminate this problem.

Several points on system performance can now be discussed: The reader might have wondered why graphs of different frequency regions are separated. The reason for examining the spectrum in discrete bands is due to the limited number of data points the HP3577A takes over a frequency sweep and the limited resolution bandwidth of the HP3577A. The unit can take up to 401 data points across a frequency sweep, no matter the size of the frequency range being swept. If the system user is not careful, spectral components can fall into the region between data points and be missed. For good accuracy, 500 KHz regions must be examined separately.

Spurious frequency components seen around the spectral components may be system generated. These unit generated spurs arise from the fact that, since the moving filter has some finite bandwidth, the spectral component is still introduced into the HP3577A during the times the receiver is not exactly tuned to that frequency. So while the signal is in the system bandwidth, but not exactly at the source frequency, some spurious signals may be generated. These spurious frequency components will have power levels at least 40 dB below the fundamental.

Accuracy, Linearity, and Calibration

The modification has a system loss of 15 dB from input to output as well as high and low-frequency rolloff. These problems, which cause inaccuracy, can be calibrated out of the system's performance. The calibration procedure is quite simple and is shown on Figure 30. The source frequency from the HP3577A is split and one component is fed directly into the preamplifier. Since the source frequency will always stay at the system's tuned frequency, it can be used as a reference. Due to the way it was necessary to set the system up, the actual amplitude of any input signal will be 7 dB below the display value.

Figures 31-33 show a 3 MHz sine wave and the harmonics generated. The actual level of the input signal is -10 dBm. The system shows the correct -3.8 dBm level (allowing for the 7 dB offset and 1 dB accuracy). The power level of the harmonics is at least 40 dB below the fundamental.

Figures 34-36 show a 9 MHz sine wave at different amplitudes. Figure 34 is at a level of 0 dBm (-7 dBm). The signal is attenuated 10 dB in Figure 35 and 40 dB in Figure 36. Notice the linearity over the different amplitudes.

Phase Coherency Measurements

A very important part of the system performance is phase coherency measurements. Figure 37 shows the spectra of a split

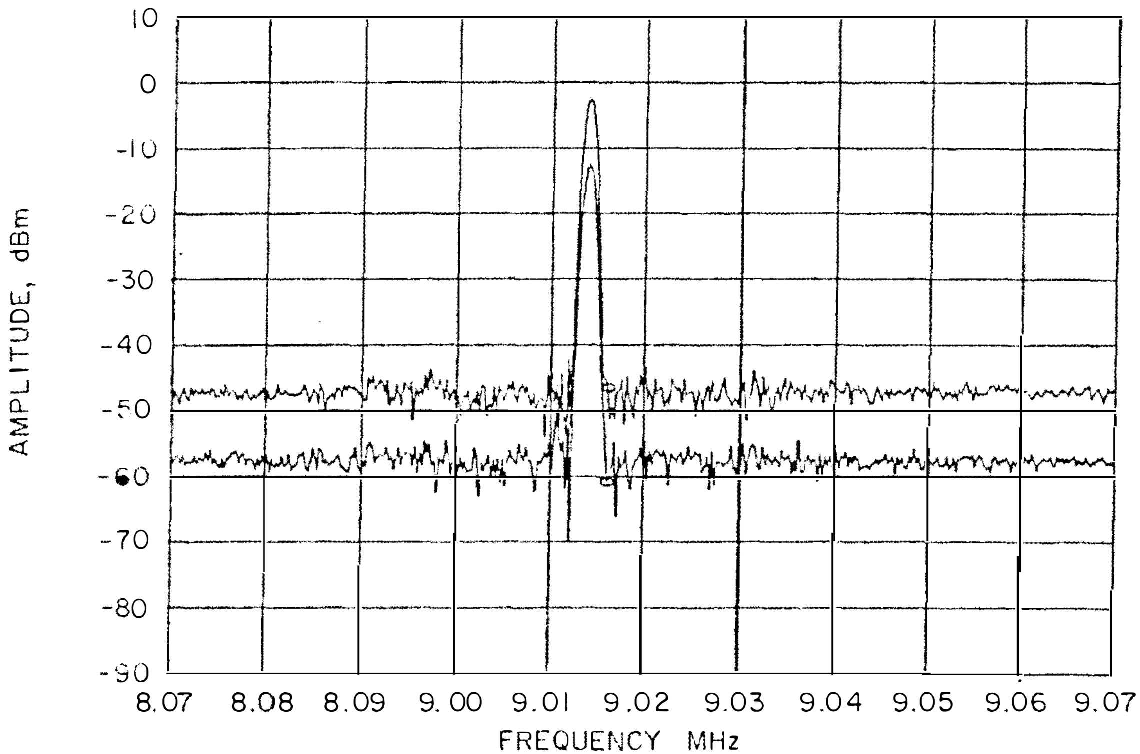


Figure 37. 9.013 MHz sine waves fed into each channel of system.

9 MHz sine wave fed into both channels of the modification. Notice the amplitude of the signal is identical on both channels, as it should be. Figure 38 is the ratio of the phase of the two signals. The phase ratio is constant, confirming the system's ability to make phase measurements.

System Dynamic Range

Figure 39 is the graph that outlines the dynamic range in amplitude and frequency for the system. The frequency rolloff on the low end is caused by the decreasing system local oscillator level. The frequency rolloff on the high end is caused by the band limiting 10 MHz filters. The maximum amplitude input is determined by the mixer's maximum input level and the preamplifier's slew rate. The noise level of the system is due to the direct feed through of the synthesized 45 MHz signal. The 45 MHz feeds through the input mixer, passes the crystal filter, and is mixed to the current receiver frequency.

Figure 40 shows the amplitude and phase of the system across the 450 kHz to 10 MHz band before calibration. Figure 41 shows the calibrated system response.

Figure 42 is the normalized crosstalk between channels. The crosstalk is less than -40 dB over the frequency change.

A measurement taken with the low frequency section of a channel is shown in Figure 43. Figure 43 is the HP3577A display with a -25 dBm sine wave at 300 kHz. A sine wave at 1 MHz was also

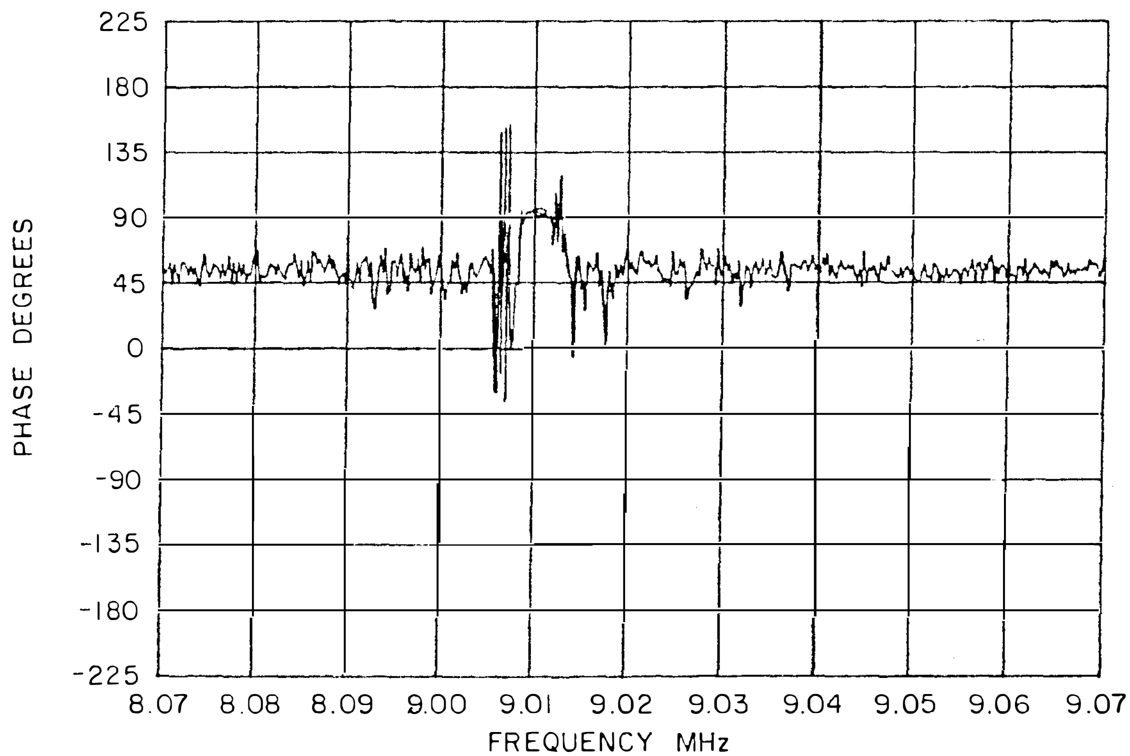


Figure 38. Phase relation between two 9.013 MHz sine waves.

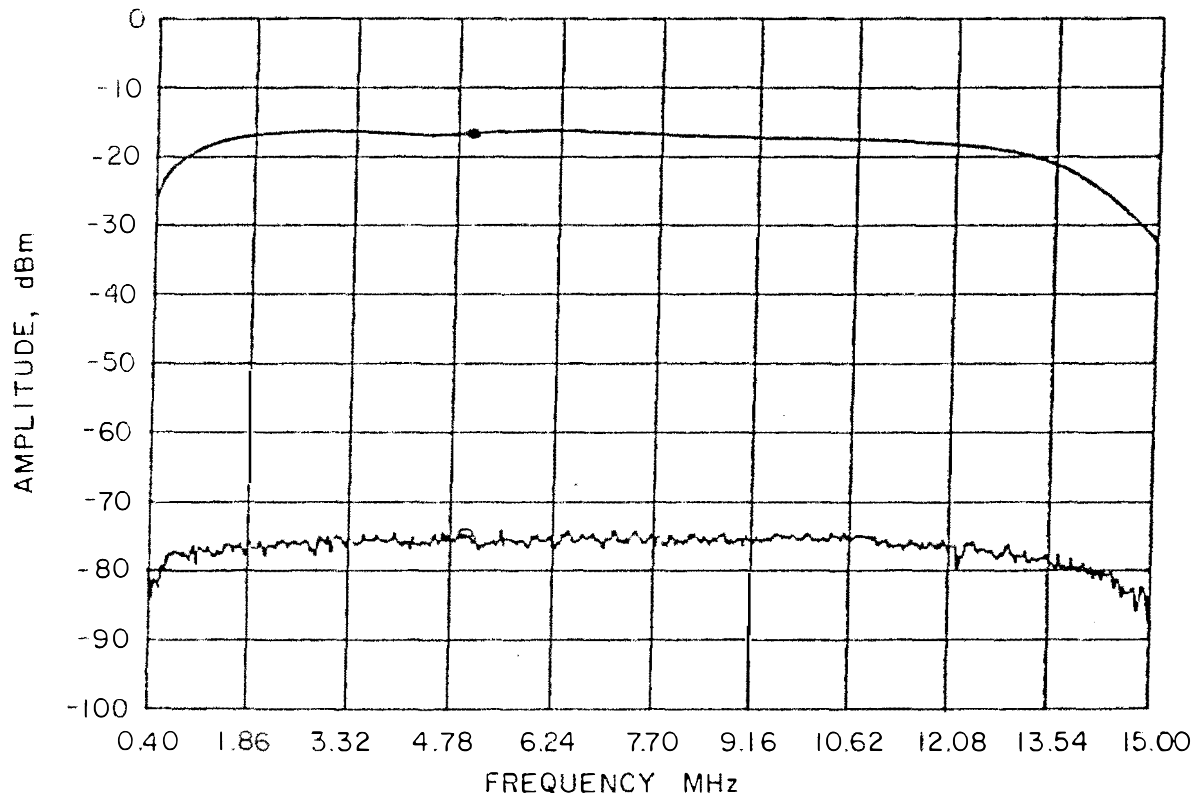


Figure 39. Dynamic range of high-frequency section.

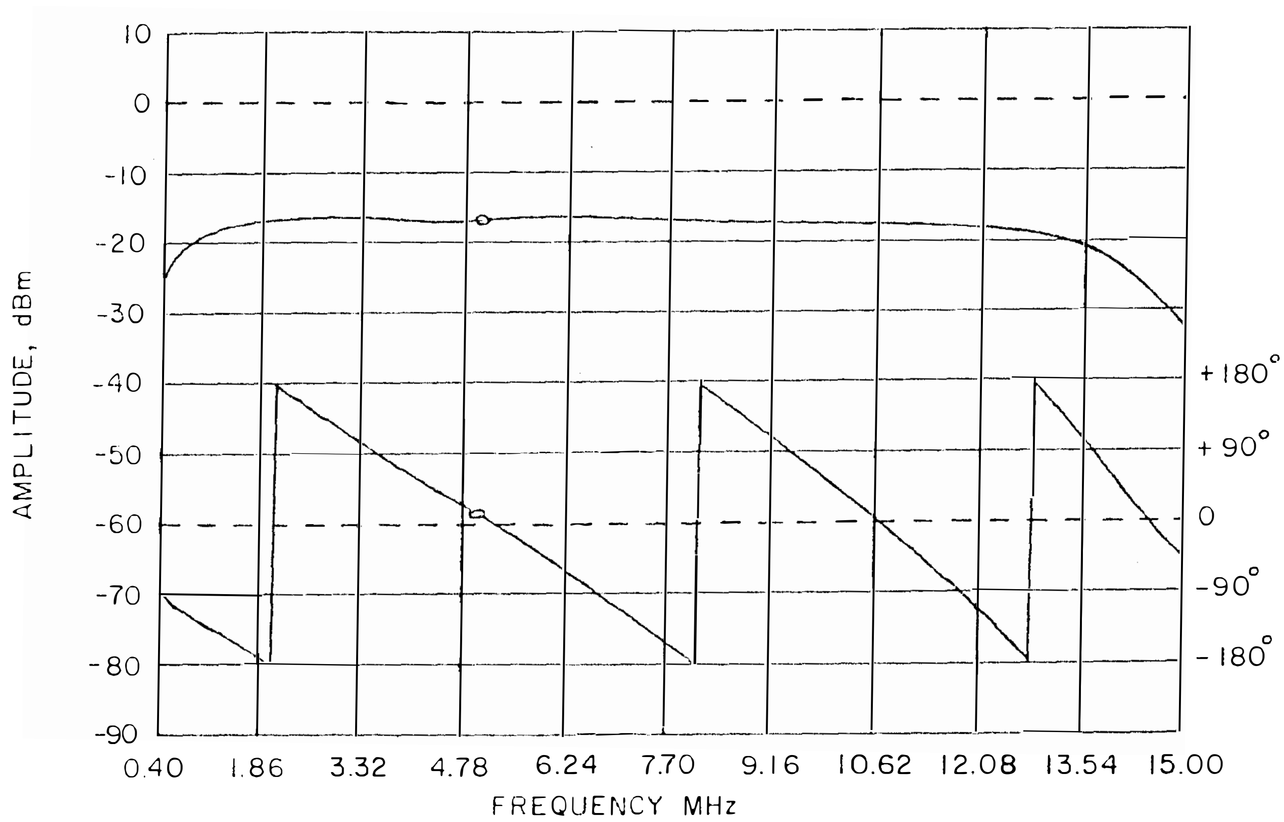


Figure 40. Amplitude and phase response of high-frequency section.

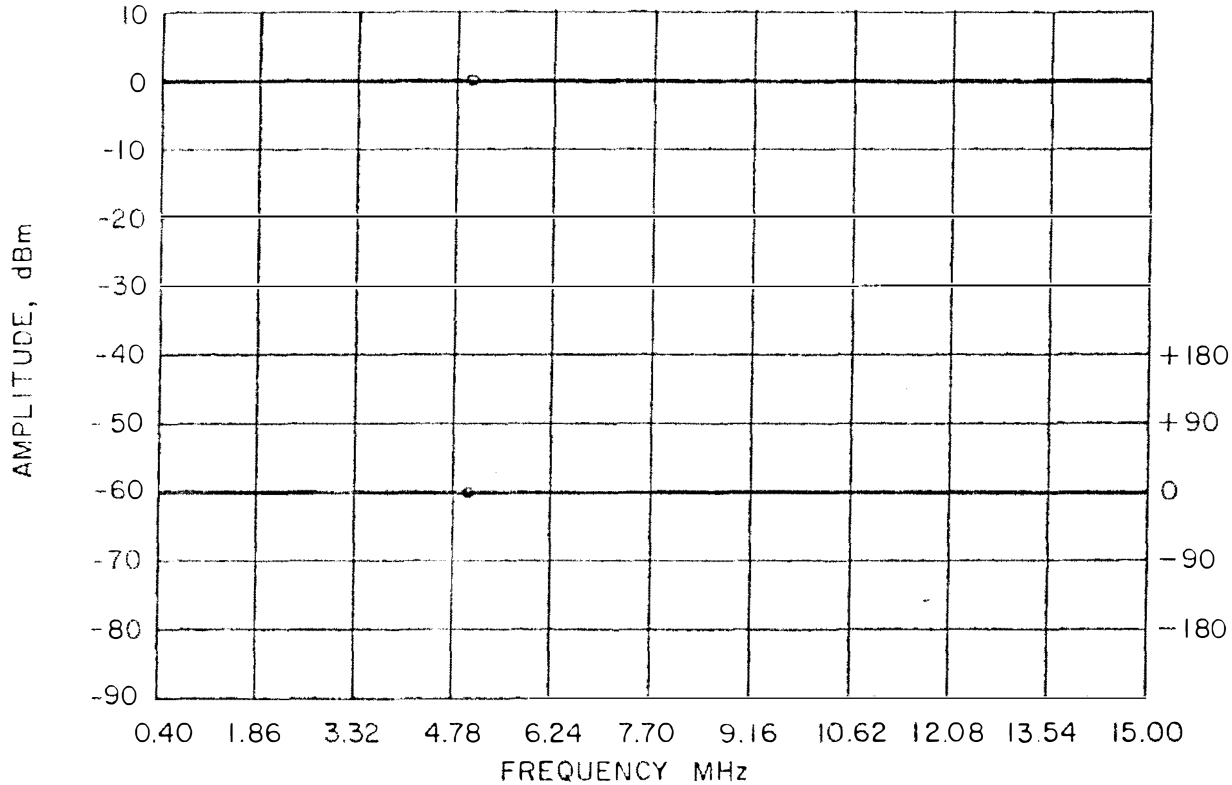


Figure 41. Normalized amplitude and phase response of high-frequency section.

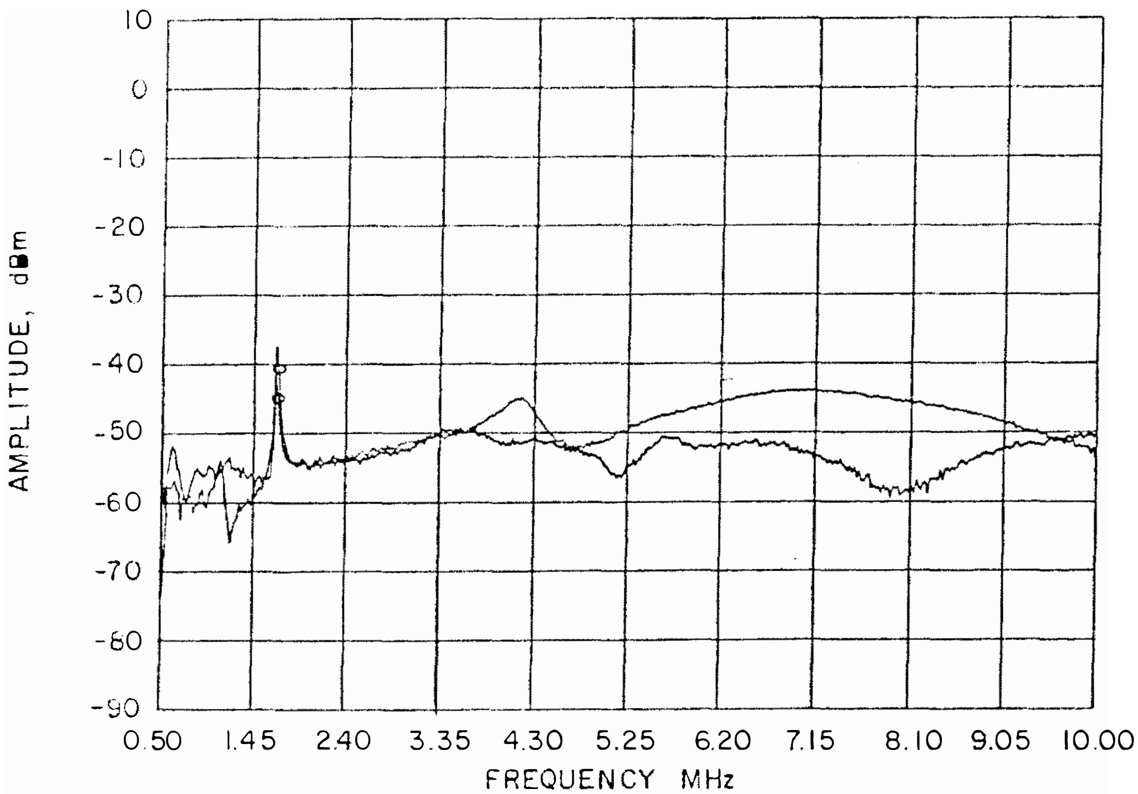


Figure 42. Normalized crosstalk between channels.

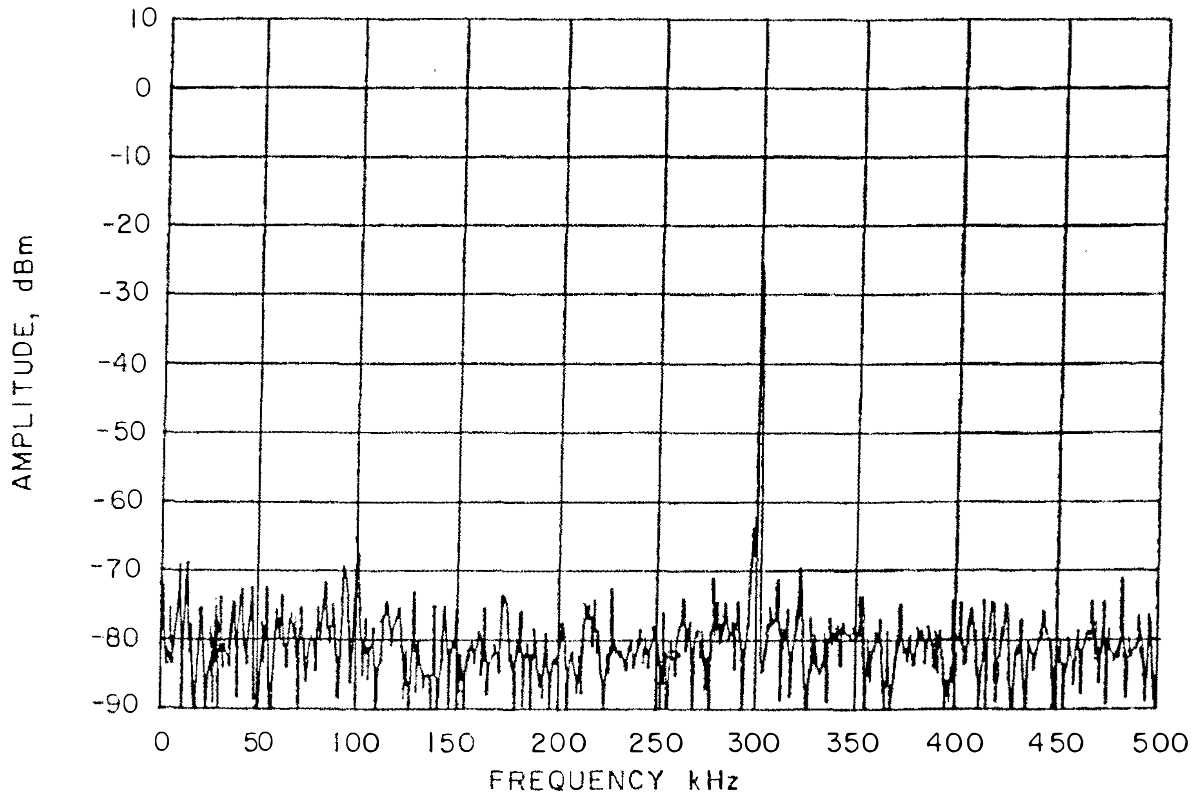


Figure 43. HP3577A display of 300 kHz sine wave in low-frequency channel.

fed into the low frequency; no spurious signals were displayed in the 5 Hz to 450 kHz range.

Conclusions

The system worked as expected. The unit specifications are shown again for clarity.

Frequenty Range	5 Hz to 450 kHz and 450 kHz to 10 MHz
Low Frequency Section Dynamic Range	0 to -90 dBm
High Frequency Section Dynamic Range	-6 to -65 dBm
Low Frequency Harmonics	>-40 dBc
High Frequency Harmonics	>-40 dBc
High Frequency Section Spurious Signals:	
450 to 1500 kHz	>-30 dBc
1.5 to 10 MHz	>-40 dBc
Accuracy:	
Low Frequency Section	0.001 dB magnitude and 1 degree phase
High Frequency Section	1 dB magnitude and 5 degrees phase
Crosstalk:	
Low Frequency Section	>-60 dB
High Frequency Section	>-40 dB

REFERENCES

REFERENCES

1. Microwave Solid-State Devices, Samuel Y. Liao, Prentice-Hall. Englewood Cliffs, New Jersey 07632. pp. 306-307, 1985.

2. Operating Manual Model 3577A Network Analyzer, Hewlett-Packard Company, P. O. Box 69, Marysville, Washington, 98270 U. S. A., Microfiche Part No. 03577-90050, Section 5, p. 8, November 1983.

3. Single-Wafer Four-Pole Monolithic Crystal Filters (Brochure), Piezo Technology, Inc., P. O. Box 7859, Orlando, Florida 32854-7859, pp. 1-4.

4. RF/IF Signal Processing Handbook, Mini-Circuits, Inc., P. O. Box 166, Brooklyn, New York 11235, pp. 33, 17-29, and 164-165, 1985-86.

5. 36 Frequently Asked Questions About Selecting Mixers, Application Note 4, Mini-Circuits Co., P. O. Box 166, Brooklyn, New York 11235, pp. 1-5, 1985-86.

6. Practical Design and Evaluation of High Frequency Circuits, Applications Note 7, Hewlett-Packard Company, P. O. Box 69, Marysville, Washington, 98270 U. S. A., pp. 1-10.

7. Mixer Distortion Measurements, Application Bulletin 7, Hewlett-Packard Company, P. O. Box 69, Marysville, Washington, 98270 U. S. A., P. O. Box 166, Brooklyn, New York 11235, pp. 1-6.

8. Introduction to Modern Network Synthesis, Van Valkenberg, John Wiley and Sons, New York, Chapter 16, p. 447, 1960.

9. Transmission Networks and Wave Filters, T. E. Shea, M. S., D. Van Nostrand Company, Inc., New York, pp. 20-36, 1929.

10. Network Analysis and Synthesis, Louis Weinberg, McGraw-Hill Book Company, Inc., New York, Chapters 11-12, 1962.

11. Microwave Solid-State Devices, Samuel Y. Liao, Prentice-Hall, Englewood Cliffs, New Jersey 07632, pp. 306-307, 1985.

12. Reference Data for Radio Engineers, Fourth Edition, H. P. Westman, editor, International Telephone and Telegraph Corporation, pp. 166-169, 1956.

APPENDIXES

APPENDIX A

HP3577A CONVERSION SYSTEM SCHEMATICS

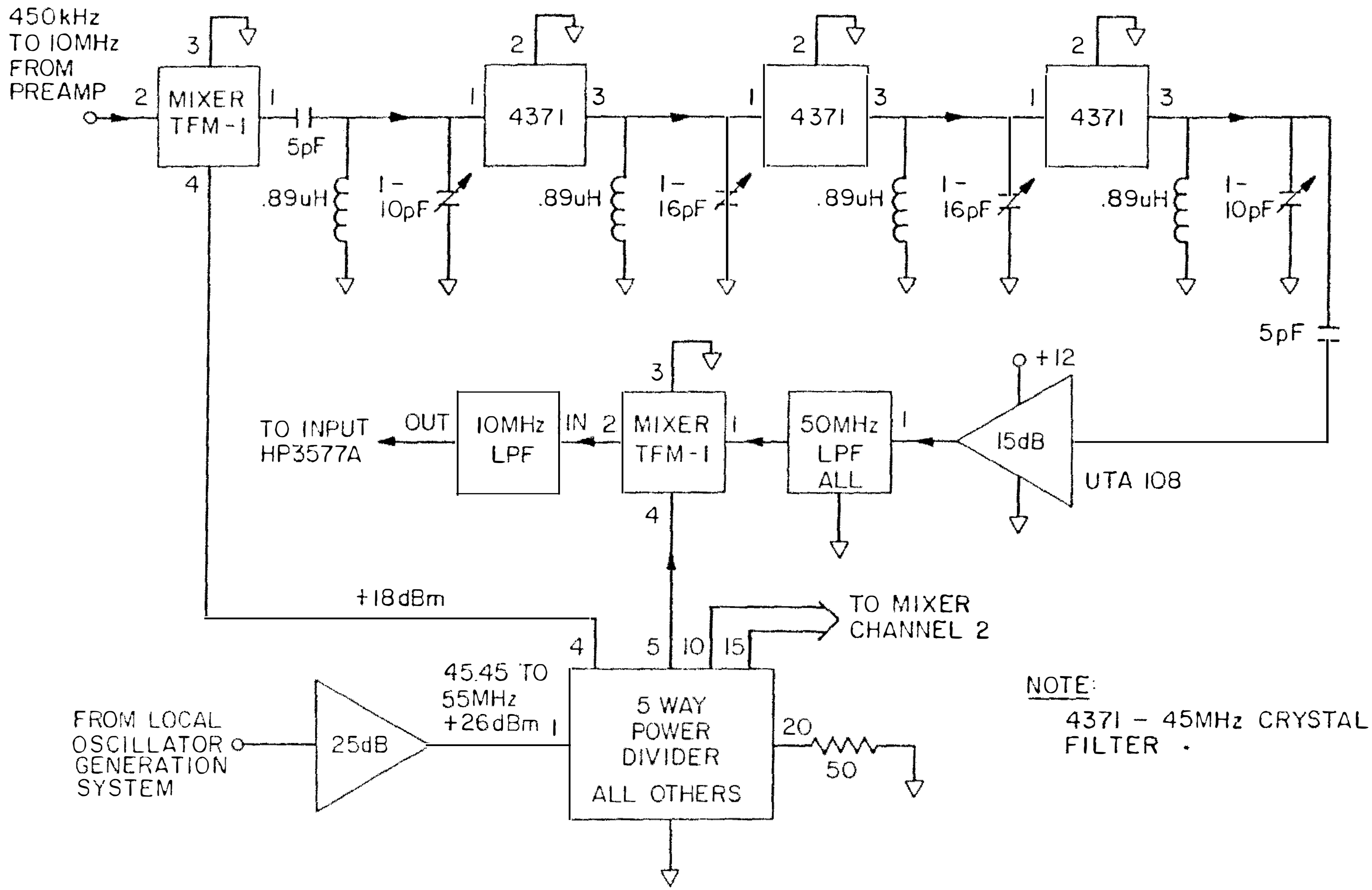


Figure A-1. Filter board schematic (1 channel).

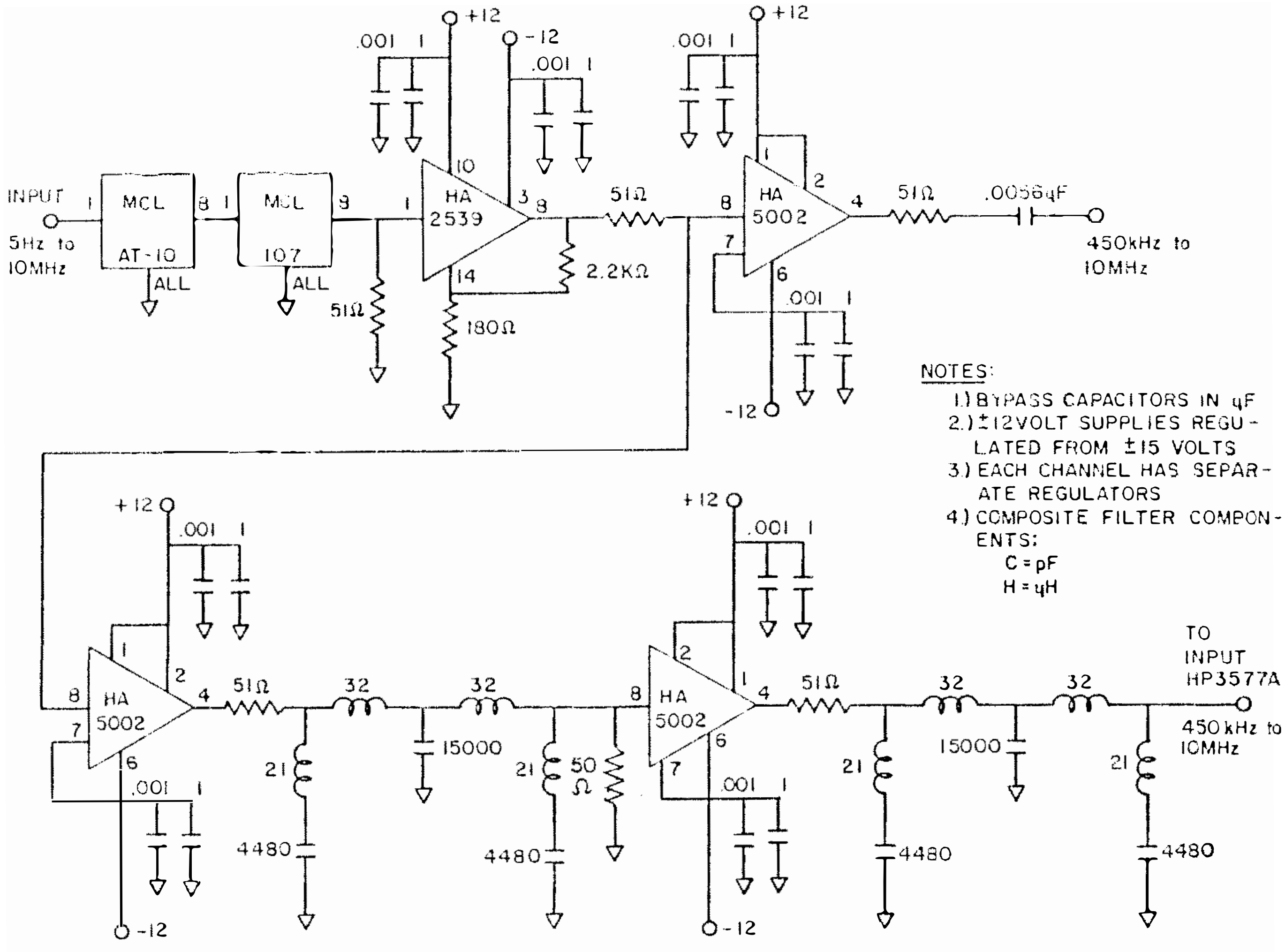


Figure A-2. Preamplifier schematic (1 channel).

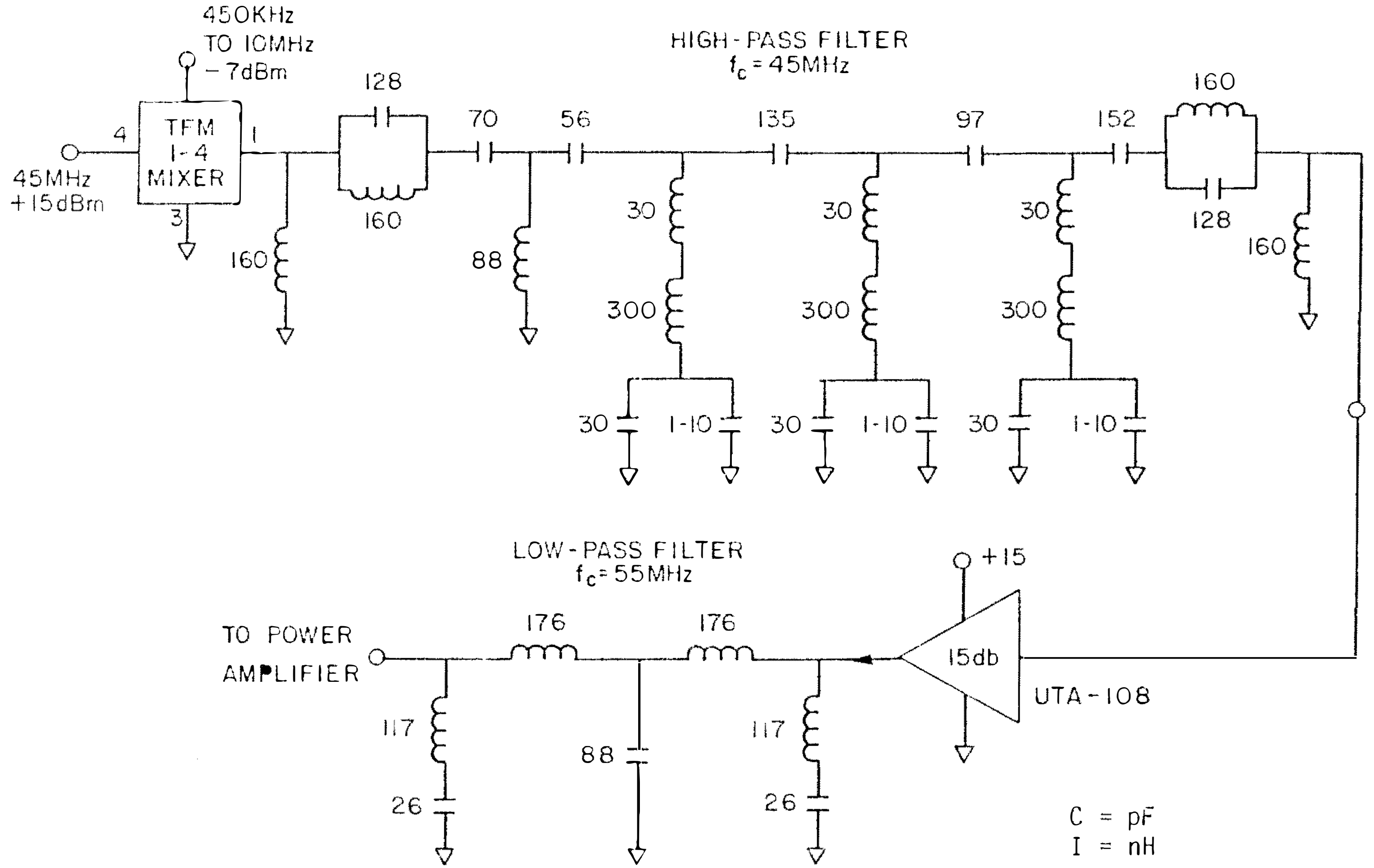


Figure A-3. Local oscillator generation system schematic.

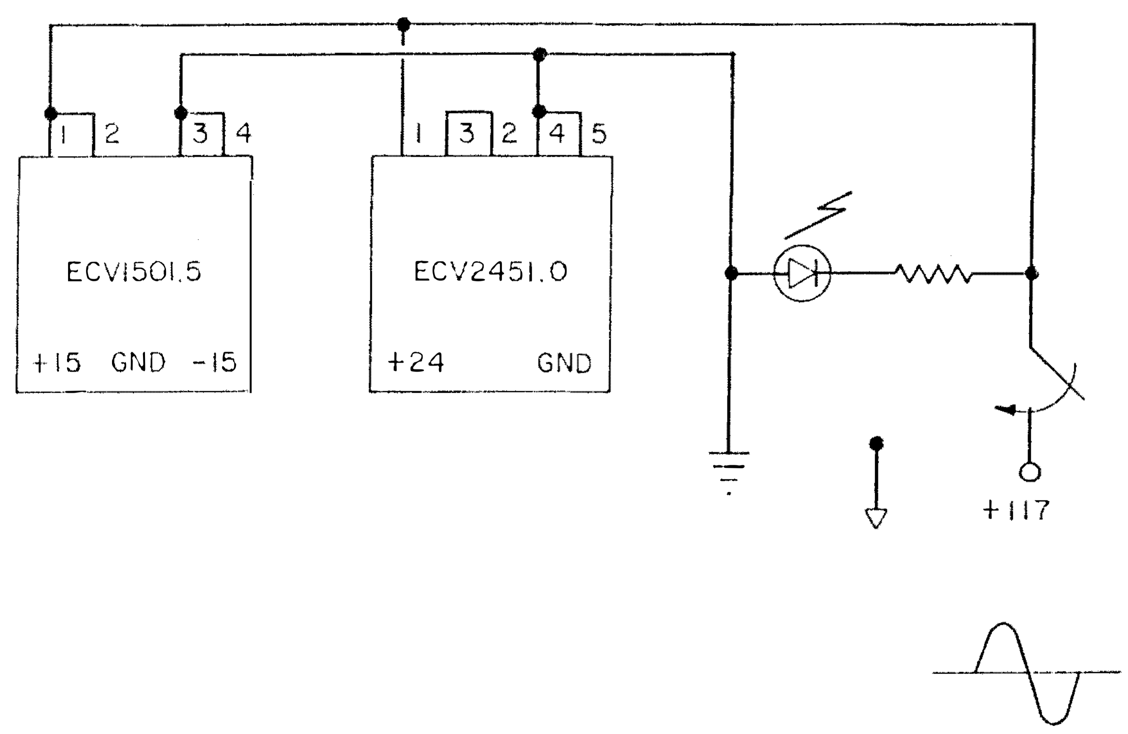


Figure A-4. Power supply block diagram.

APPENDIX B

THEORETICAL AND ACTUAL MEASUREMENTS OF THE HARMONICS OF A SQUARE WAVE

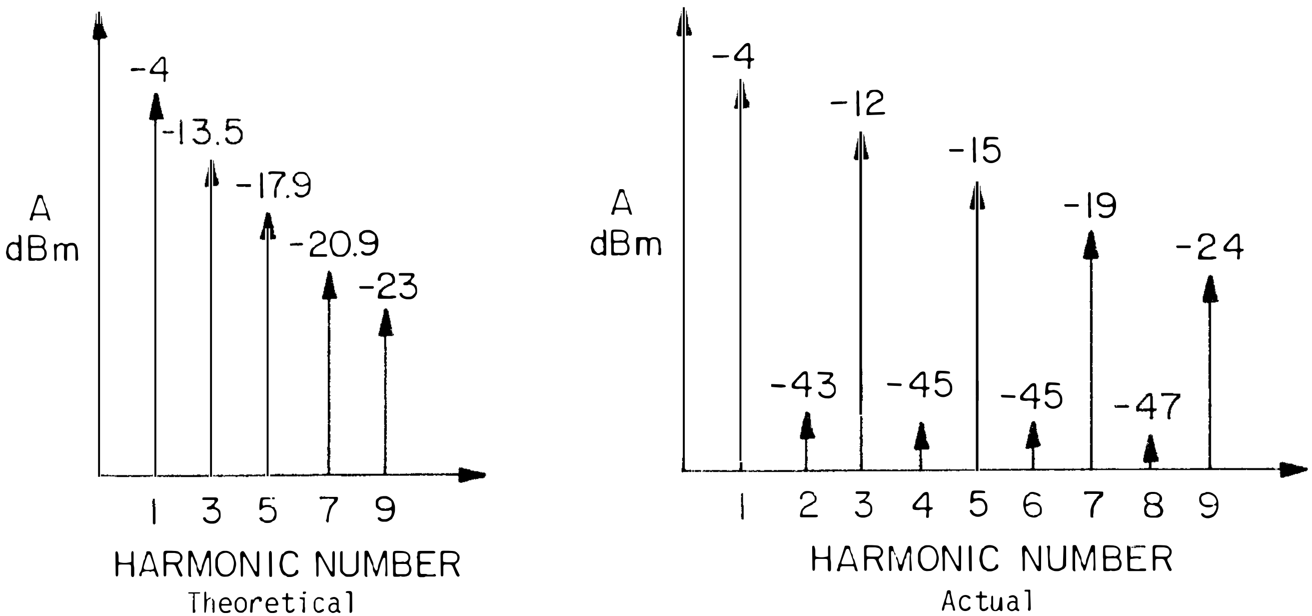


Figure B-1. Theoretical vs. actual spectral components of square wave.

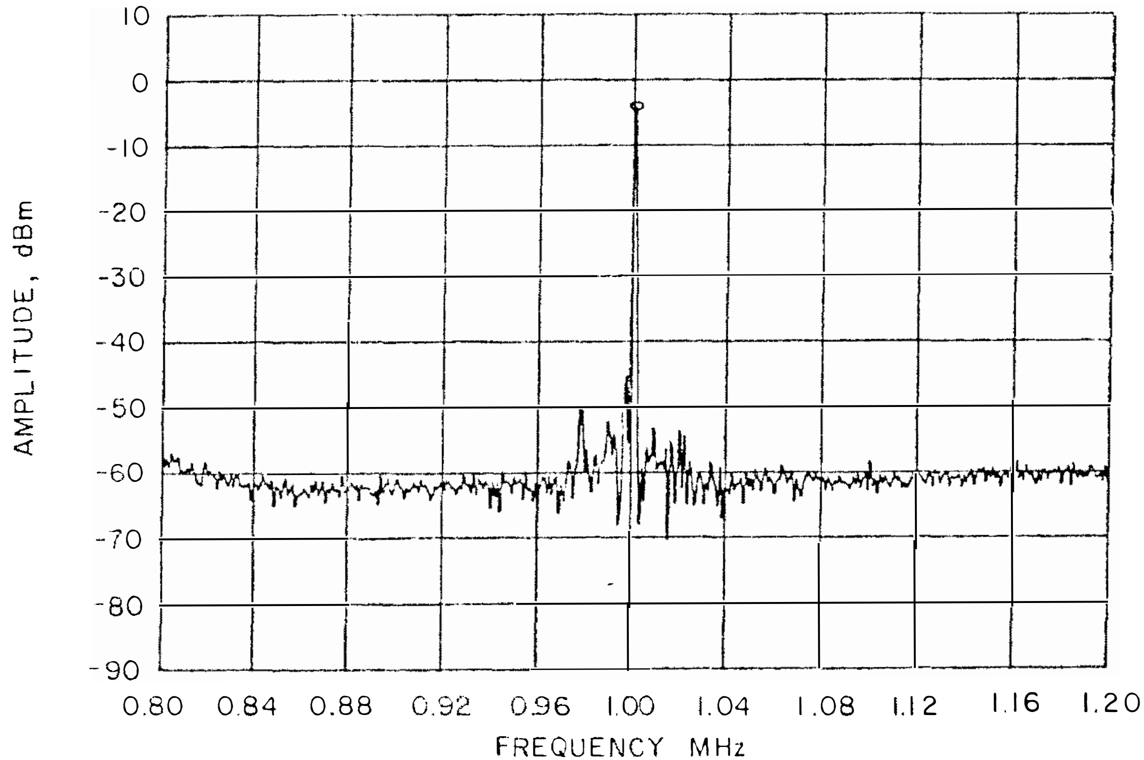


Figure B-2. Fundamental of square wave measured by conversion system.

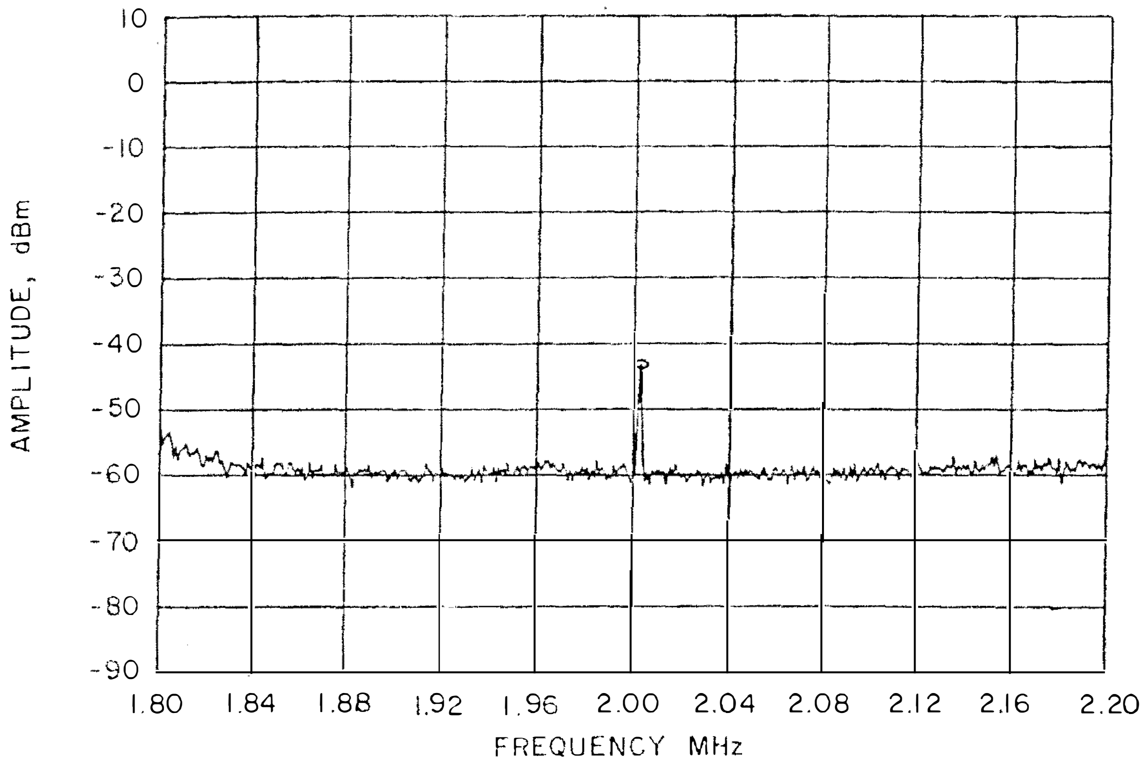


Figure B-3. Second harmonic of square wave.

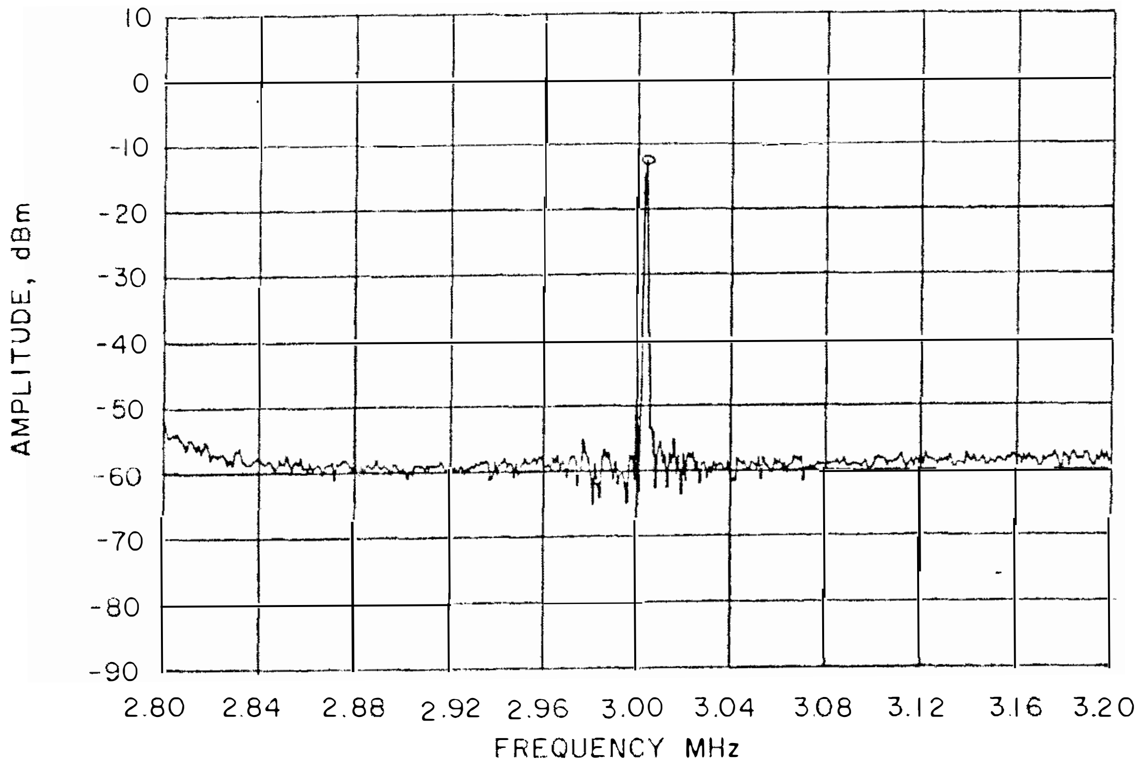


Figure B-4. Third harmonic of square wave.

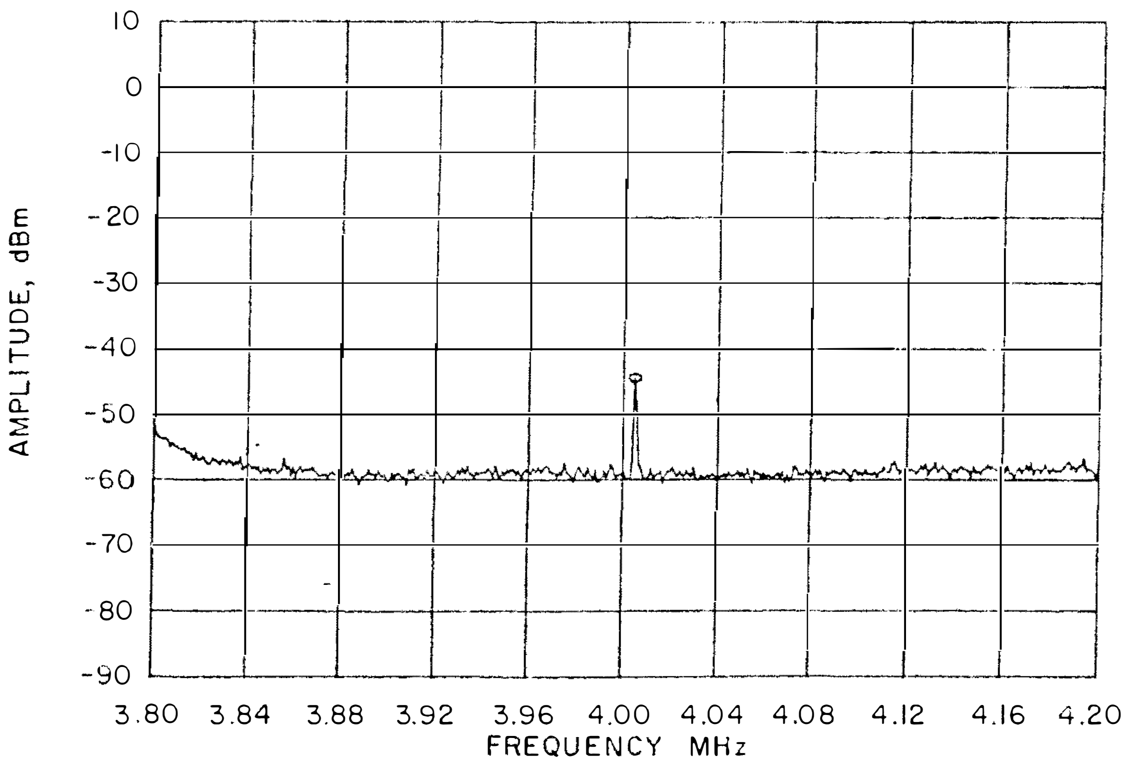


Figure B-5. Fourth harmonic of square wave.

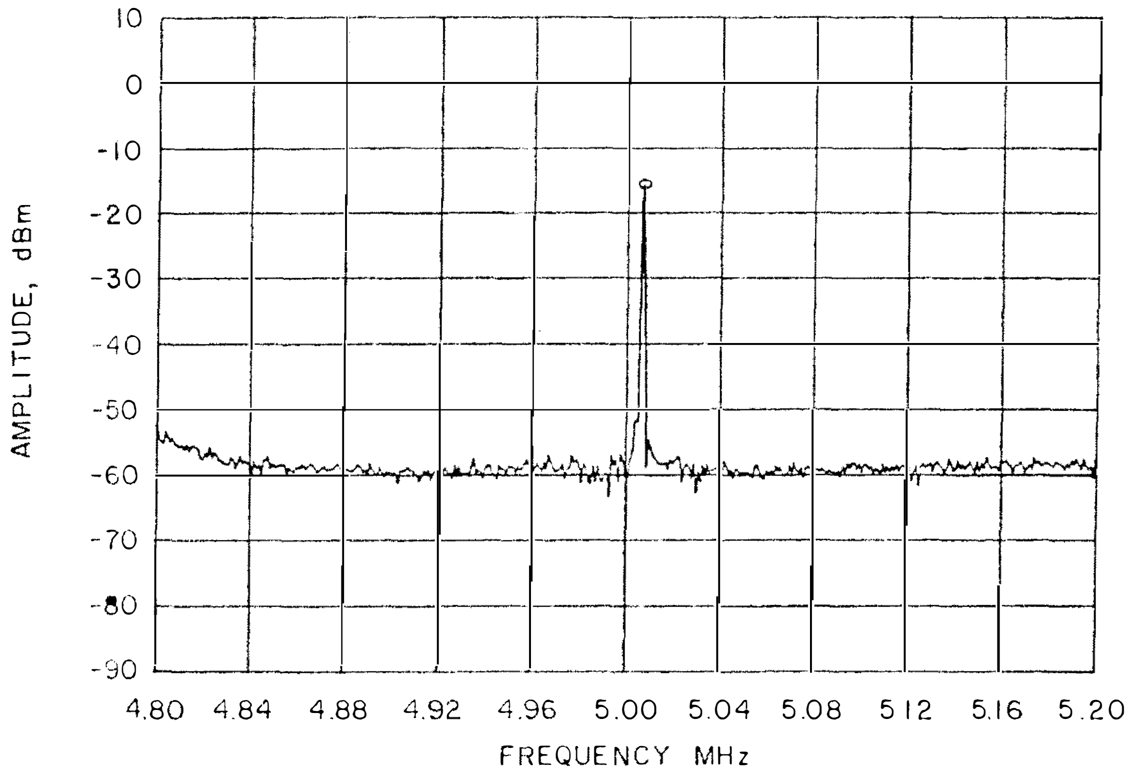


Figure B-6. Fifth harmonic of square wave.

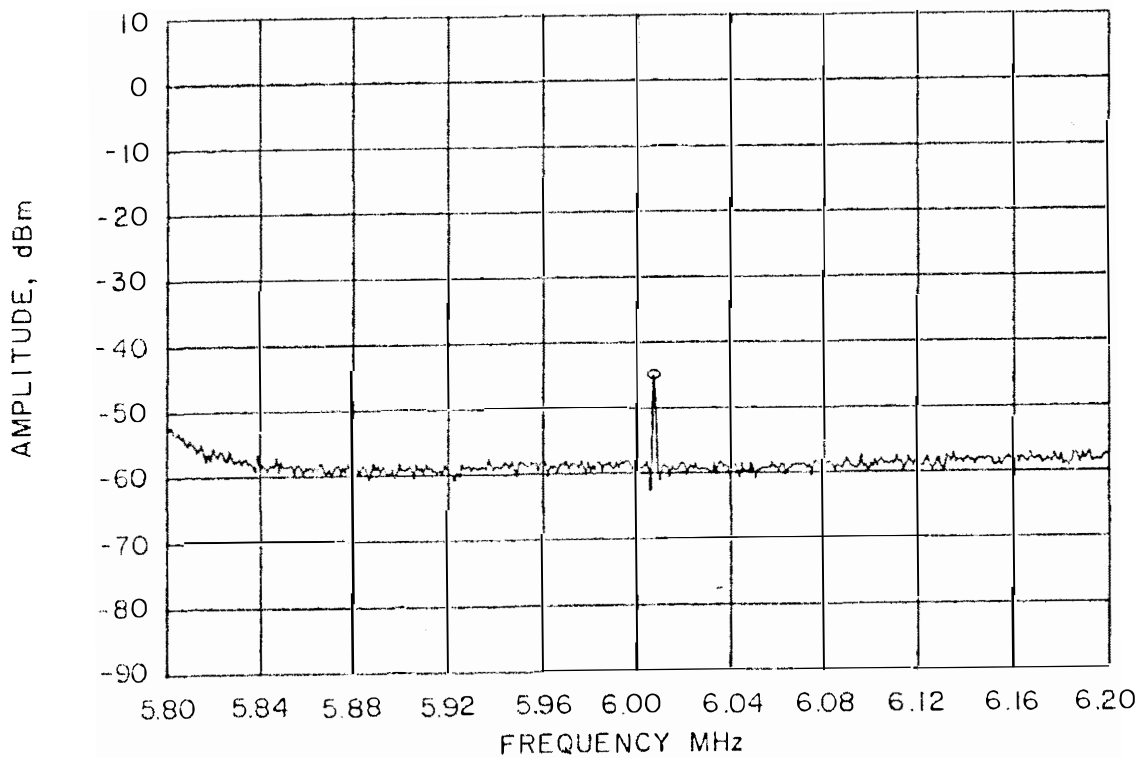


Figure B-7. Sixth harmonic of square wave.

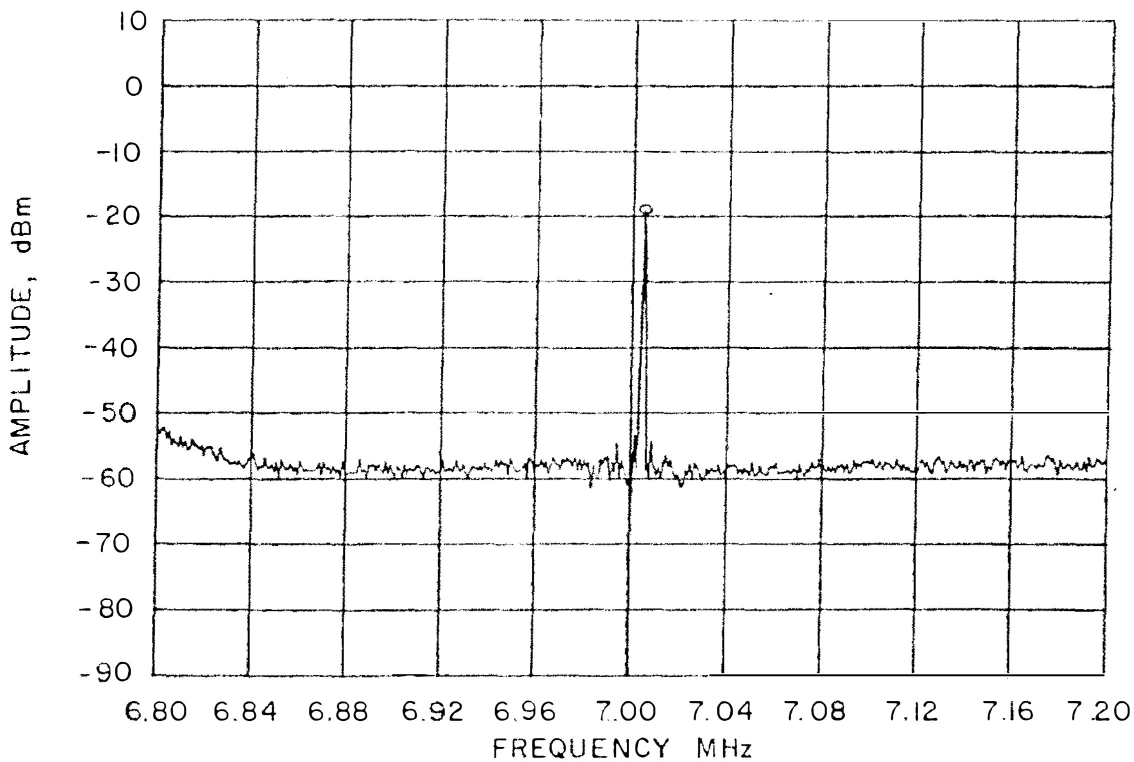


Figure B-8. Seventh harmonic of square wave.

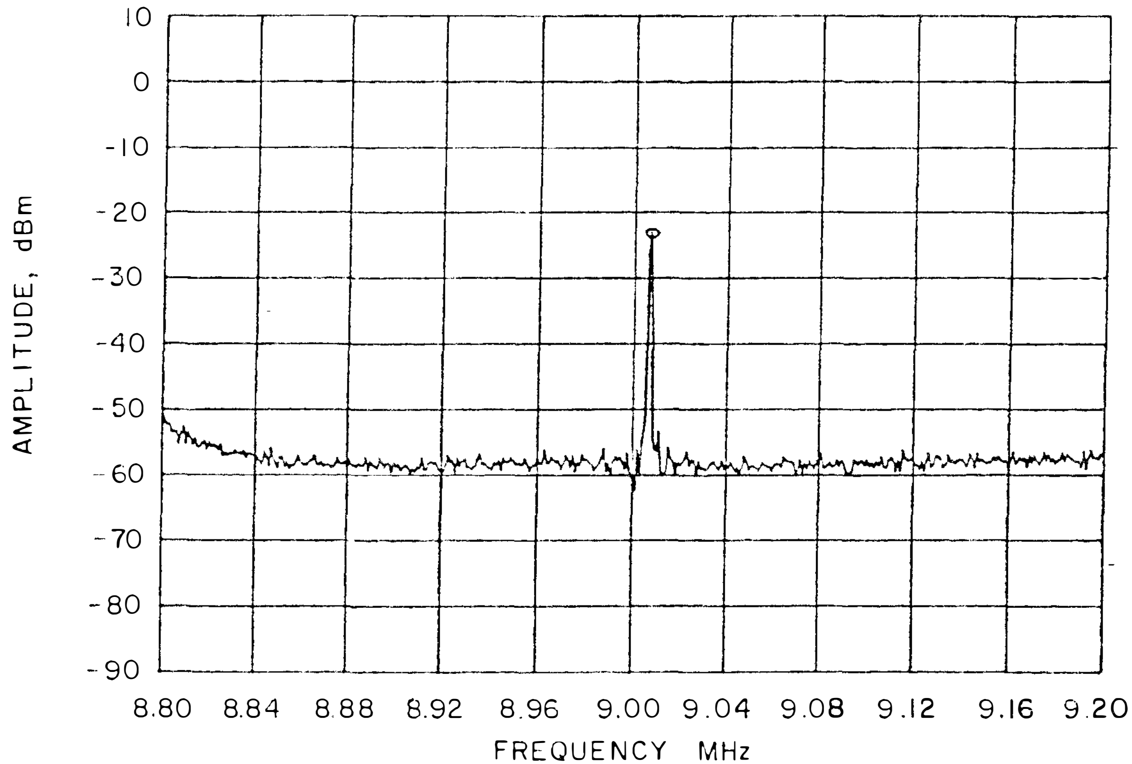
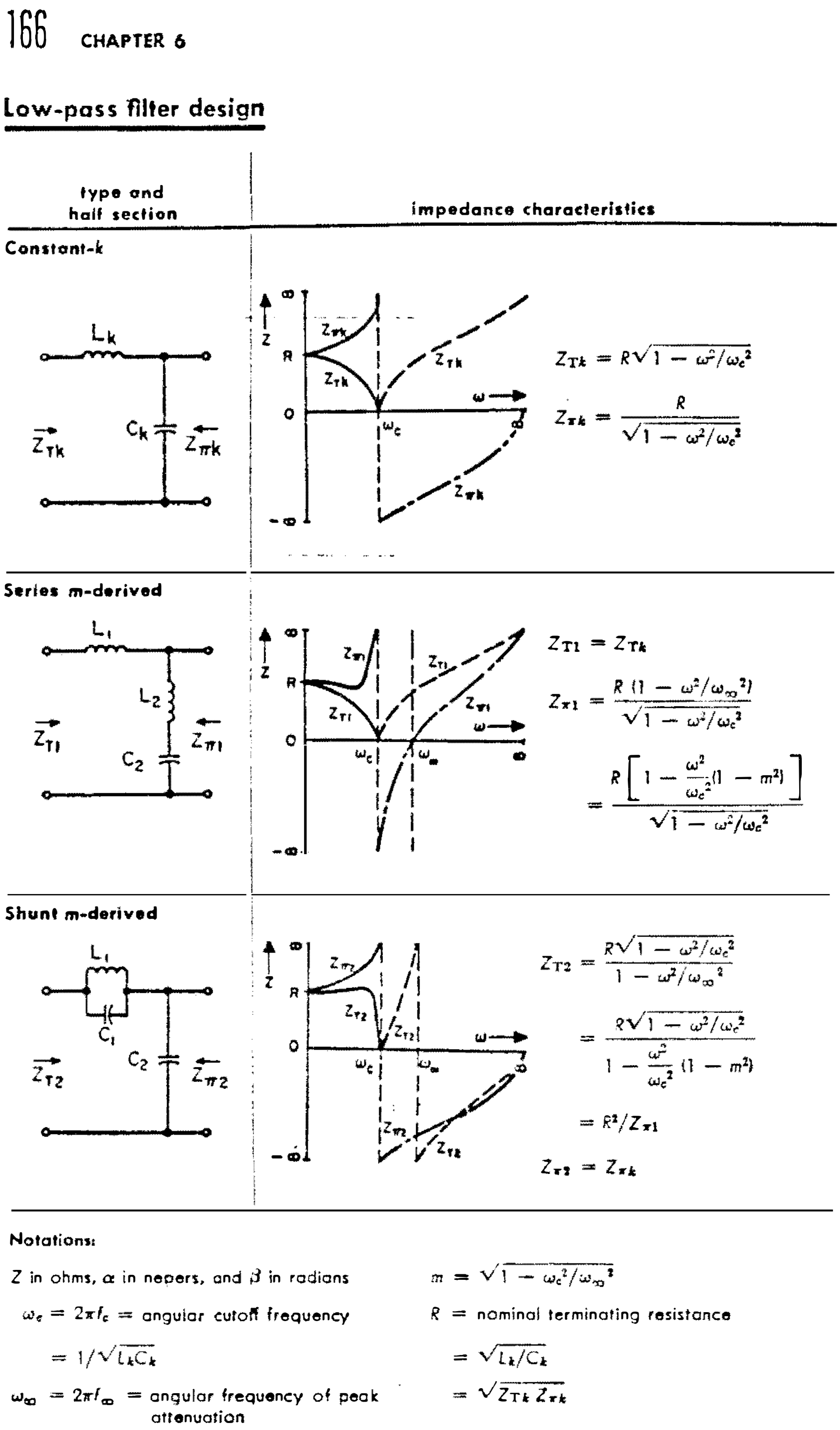


Figure B-9. Ninth harmonic of square wave.

APPENDIX C

DESIGN DATA SHEETS FOR COMPOSITE FILTERS



166 CHAPTER 6

Low-pass filter design

type and half section	impedance characteristics
Constant-k	$Z_{Tk} = R\sqrt{1 - \omega^2/\omega_c^2}$ $Z_{\pi k} = \dfrac{R}{\sqrt{1 - \omega^2/\omega_c^2}}$
Series m-derived	$Z_{T1} = Z_{Tk}$ $Z_{\pi 1} = \dfrac{R(1 - \omega^2/\omega_\infty^2)}{\sqrt{1 - \omega^2/\omega_c^2}} = \dfrac{R\left[1 - \dfrac{\omega^2}{\omega_c^2}(1 - m^2)\right]}{\sqrt{1 - \omega^2/\omega_c^2}}$
Shunt m-derived	$Z_{T2} = \dfrac{R\sqrt{1 - \omega^2/\omega_c^2}}{1 - \omega^2/\omega_\infty^2} = \dfrac{R\sqrt{1 - \omega^2/\omega_c^2}}{1 - \dfrac{\omega^2}{\omega_c^2}(1 - m^2)} = R^2/Z_{\pi 1}$ $Z_{\pi 2} = Z_{\pi k}$

Notations:

Z in ohms, α in nepers, and β in radians

$\omega_c = 2\pi f_c$ = angular cutoff frequency $= 1/\sqrt{L_k C_k}$

$\omega_\infty = 2\pi f_\infty$ = angular frequency of peak attenuation

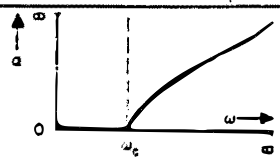
$m = \sqrt{1 - \omega_c^2/\omega_\infty^2}$

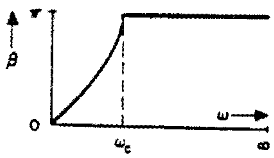
R = nominal terminating resistance $= \sqrt{L_k/C_k} = \sqrt{Z_{Tk} Z_{\pi k}}$

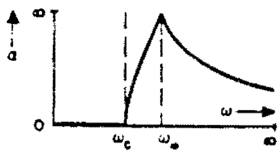
Figure C-1. Composite filter low-pass design equations.

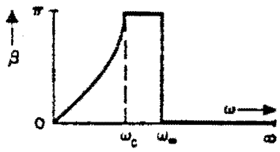
SOURCE: Reference Data for Radio Engineers, fourth edition, H. P. Westman, editor, International Telephone and Telegraph Corporation, pp. 166-167.

full-section attenuation α and phase β characteristics	design formulas: half-section series arm	design formulas: half-section shunt arm
When $0 \leqslant \omega \leqslant \omega_c$: $\alpha = 0$, $\beta = 2 \sin^{-1} \frac{\omega}{\omega_c}$. When $\omega_c < \omega < \infty$: $\beta = \pi$, $\alpha = 2 \cosh^{-1} \frac{\omega}{\omega_c}$	$L_k = \frac{R}{\omega_c}$	$C_k = \frac{1}{\omega_c R}$
(m-derived, series)	$L_1 = mL_k$	$L_2 = \frac{1 - m^2}{m} L_k$; $C_2 = mC_k$
(m-derived, shunt)	$L_1 = mL_k$; $C_1 = \frac{1 - m^2}{m} C_k$	$C_2 = mC_k$









When $\omega_c < \omega < \omega_\infty$, $\beta = \pi$ and

$$\alpha = \cosh^{-1}\left[2\,\frac{1/\omega_\infty^2 - 1/\omega_c^2}{1/\omega_\infty^2 - 1/\omega^2} - 1\right] = \cosh^{-1}\left[2\,\frac{m^2}{\omega_c^2/\omega^2 - (1 - m^2)} - 1\right]$$

When $0 \leqslant \omega \leqslant \omega_c$, $\alpha = 0$ and

$$\beta = \cos^{-1}\left[1 - 2\,\frac{1/\omega_\infty^2 - 1/\omega_c^2}{1/\omega_\infty^2 - 1/\omega^2}\right] = \cos^{-1}\left[1 - 2\,\frac{m^2}{\omega_c^2/\omega^2 - (1 - m^2)}\right]$$

When $\omega_\infty < \omega < \infty$, $\beta = 0$ and

$$\alpha = \cosh^{-1}\left[1 - 2\,\frac{1/\omega_\infty^2 - 1/\omega_c^2}{1/\omega_\infty^2 - 1/\omega^2}\right] = \cosh^{-1}\left[1 - 2\,\frac{m^2}{\omega_c^2/\omega^2 - (1 - m^2)}\right]$$

For constant-k type
$R^2 = Z_{1k} Z_{2k} = k^2$

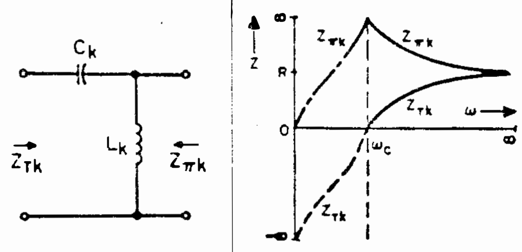
For m-derived type
Curves drawn for $m \approx 0.6$
$R^2 = Z_{T2} Z_{\pi 1}$
$= Z_{1(\text{series-}m)} Z_{2(\text{shunt-}m)}$
$= Z_{1(\text{shunt-}m)} Z_{2(\text{series-}m)}$

Figure C-1 (continued).

High-pass filter design

type and half section	impedance characteristics

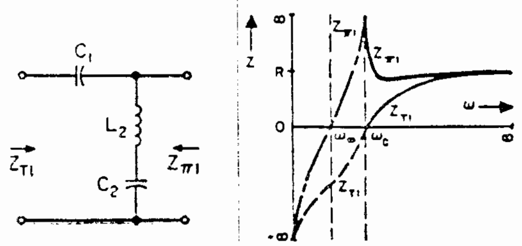
Constant-k



$$Z_{Tk} = R\sqrt{1 - \frac{\omega_c^2}{\omega^2}}$$

$$Z_{\pi k} = \frac{R}{\sqrt{1 - \frac{\omega_c^2}{\omega^2}}}$$

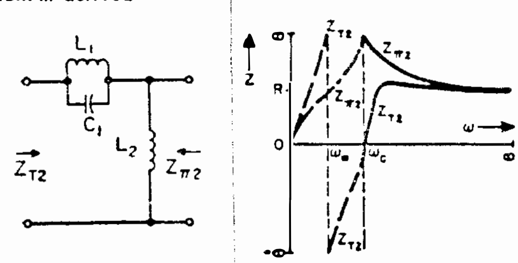
Series m-derived



$$Z_{T1} = Z_{Tk}$$

$$Z_{\pi 1} = \frac{R\left(1 - \frac{\omega_\infty^2}{\omega^2}\right)}{\sqrt{1 - \omega_c^2/\omega^2}} = \frac{R\left[1 - \frac{\omega_c^2}{\omega^2}(1 - m^2)\right]}{\sqrt{1 - \omega_c^2/\omega^2}}$$

Shunt m-derived



$$Z_{T2} = \frac{R\sqrt{1 - \omega_c^2/\omega^2}}{1 - \omega_\infty^2/\omega^2} = \frac{R\sqrt{1 - \omega_c^2/\omega^2}}{1 - \frac{\omega_c^2}{\omega^2}(1 - m^2)} = R^2/Z_{\pi 1}$$

$$Z_{\pi 2} = Z_{\pi k}$$

Notations:

Z in ohms, α in nepers, and β in radians

$\omega_c = 2\pi f_c$ = angular cutoff frequency $= 1/\sqrt{L_k C_k}$

$\omega_\infty = 2\pi f_\infty$ = angular frequency of peak attenuation

$m = \sqrt{1 - \omega_\infty^2/\omega_c^2}$

R = nominal terminating resistance $= \sqrt{L_k/C_k} = \sqrt{Z_{Tk} Z_{\pi k}}$

Figure C-2. High-pass design equations for composite filters.

SOURCE: Reference Data for Radio Engineers, fourth edition, H. P. Westman, editor, International Telephone and Telegraph Corporation, pp. 168-169.

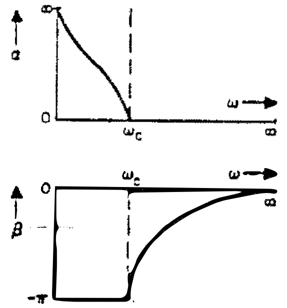
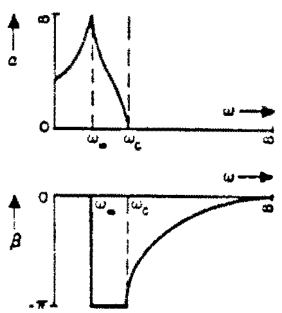
full-section attenuation α and phase β characteristics	design formulas: half-section series arm	design formulas: half-section shunt arm
 When $0 < \omega < \omega_c$ $\alpha = 2\cosh^{-1}\frac{\omega_c}{\omega}$, $\beta = -\pi$. When $\omega_c < \omega < \infty$ $\alpha = 0$, $\beta = -2\sin^{-1}\frac{\omega_c}{\omega}$	$C_k = \frac{1}{\omega_c R}$	$L_k = \frac{R}{\omega_c}$
	$C_1 = \frac{C_k}{m}$	$L_2 = \frac{L_k}{m}$ $C_2 = \frac{m}{1-m^2}C_k$
When $\omega_\infty < \omega < \omega_c$, $\beta = -\pi$ and $\alpha = \cosh^{-1}\left[2\frac{\omega_c^2-\omega_\infty^2}{\omega^2-\omega_\infty^2} - 1\right] = \cosh^{-1}\left[2\frac{m^2}{\frac{\omega^2}{\omega_c^2}-(1-m^2)} - 1\right]$ When $0 < \omega < \omega_\infty$, $\beta = 0$ and $\alpha = \cosh^{-1}\left[1 - 2\frac{\omega_\infty^2-\omega_c^2}{\omega_\infty^2-\omega^2}\right] = \cosh^{-1}\left[1 + 2\frac{m^2}{(1-m^2)-\frac{\omega^2}{\omega_c^2}}\right]$ When $\omega_c < \omega < \infty$, $\alpha = 0$ and $\beta = \cos^{-1}\left[1 - 2\frac{\omega_\infty^2-\omega_c^2}{\omega_\infty^2-\omega^2}\right] = \cos^{-1}\left[1 + 2\frac{m^2}{(1-m^2)-\frac{\omega^2}{\omega_c^2}}\right]$	$L_1 = \frac{m}{1-m^2}L_k$ $C_1 = \frac{C_k}{m}$ For constant-k type $R^2 = Z_{1k}Z_{2k} = k^2$ For m-derived type Curves drawn for $m \approx 0.6$ $R^2 = Z_{T2}Z_{\pi1} = Z_{1(\text{series-m})}Z_{2(\text{shunt-m})} = Z_{1(\text{shunt-m})}Z_{2(\text{series-m})}$	$L_2 = \frac{L_k}{m}$

Figure C-2 (continued).

VITA

The author was born in Dickson, Tennessee, on May 2, 1960. He attended high school at Father Ryan in Nashville, Tennessee. He obtained a Bachelor's degree in Electrical Engineering from The University of Tennessee, Knoxville in December of 1982.

The author was employed with Boeing Aerospace Company from January 1983 to January 1985. The position's responsibilities included design and development of electronic and microwave systems and components.

In January 1985, the author returned to The University of Tennessee, Knoxville to work on his Master's degree in Electrical Engineering. The major courses of study for the Master's degree were electronics and networks.

The author received his M. S. in Electrical Engineering in March 1987. He then went to work for a company called Seal-Tech in Birmingham, Alabama.