A MULTIPLEX ADAPTER DESIGN USING A BALANCED SQUARE LAW PRODUCT DETECTOR

NORMAN W. PETERSEN

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Lieutenant, United States Navy

Submitted in partial fulfillment of
the requirements for the degree of

MASTER OF SCIENCE
IN
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ABSTRACT

During the past year, since the approval of a FM stereophonic broadcasting system by the Federal Communication Commission on April 20th, 1961, the audio industry has discussed two general methods of separating the two stereophonic signals—by matrixing techniques and by time-multiplexing techniques. It is the purpose of this paper to present a third method—a product detection technique. By using a uniquely designed "Balanced Square Law Product Detector" which produces no linear amplification, it is possible to extract the difference channel without the use of any filters. A system which can extract this difference channel without the use of filters is greatly desired in that using any type of filter will greatly lower the separation between the two stereophonic channels. As a further modification of the product detector technique, an examination of synchronous detection using the product detector technique is made, thereby eliminating the need for the 19,000 cycle pilot subcarrier which at times has a tendency to jitter or phase vary. This jitter may be due to audio frequency components of approximately 19,000 cycles or to the higher order power spectral density terms which result from frequency modulation.

The writer gratefully acknowledges the guidance and encouragement given him by Professor P. E. Cooper. He furthermore wishes to express his appreciation to his wife, Ann Petersen, for the many hours she spent in construction of the prototype model and its various modifications.

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CHAPTER I

INTRODUCTION

On April 20th, 1961 the Federal Communication Commission /1/, after three years of indecision, issued Docket No. 13506, establishing its policy for permitting Frequency Modulated Stereophonic Broadcasting. The decision, although actively sought and much in the public limelight, seemed to have caught the entire audio industry by surprise.

Prior to making its decision on July 20th, the FCC had released a Notice of Inquiry on July 8, 1958 /2/ for the purpose of exploring possible additional uses of FM multiplexing. Thereafter, the Electronic Industries Association organized the National Stereophonic Radio Committee (NSRC) for the purpose of developing and recommending national standards for FM stereophonic radio /3/. As a result of its studies, the NSRC submitted for consideration by the FCC, seven FM stereophonic broadcasting systems. These seven systems plus one additional system proposed by Philco Corporation were the eight systems (See Appendix A) that the FCC evaluated in arriving at their April 20th decision.

Looking first at the basic requirements of a stereophonic broadcast, we can enumerate the following criteria:

a). The system must be capable of producing a suitable frequency response up to 15,000 cycles.

b). The system should provide good separation over the major portion of the frequency spectrum.

c). The system must be compatible with tuners presently on the market, and in addition should provide a monophonic signal when received in tuners not utilizing the multiplex adapter.

Using these three criteria, the FCC was able to eliminate five of the eight systems. Of the remaining three systems, shortly prior to the deadline, one of the systems was modified to the extent that, except for minor parameter differences, it was identical to one of the remaining two systems. It therefore became a choice between the Crosby system and the modified Zenith-General Electric system. The final decision was the Zenith-General Electric system since it compared favorable on a theoretical as well as a practical basis to the Crosby system but at the same time had two advantages the Crosby system did not have:

1). The Zenith-General Electric system appeared to have a much lower cost.

2). It does not displace the Subsidiary Communications Authorization (SCA). This will be discussed later.

Evidentially the entire audio industry "guessed" that the Crosby system would be the accepted system (Heathkit even had a Crosby type multiplex adapter on the market for several years prior to April, 1961). Therefore, immediately after the announcement by the FCC that the Zenith-General Electric system was the accepted system, there was a great scramble by all the audio firms to get a multiplex adapter on the market. As a result, there are almost as many different types of multiplex adapters as there are firms manufacturing them.

CHAPTER II

STEREOPHONIC FM BROADCASTING

When the FCC issued Docket No. 13506 it merely established what was to be transmitted, not how it was to be achieved. In summary, the FCC stated:

a). The modulating signal for the main channel shall consist of the sum of the left and the right signal.

b). A pilot subcarrier at 19,000 cycles, plus or minus two cycles, shall be transmitted that shall frequency modulate the main carrier between the limits of eight and ten percent.

c). The stereophonic subcarrier shall be the second harmonic of the pilot subcarrier and shall cross the time axis with a positive slope simultaneously with each crossing of the time axis by the pilot subcarrier.

d). Amplitude modulation of the stereophonic subcarrier shall be used.

e). The stereophonic subcarrier shall be suppressed to a level less than one percent modulation of the main carrier.

f). The stereophonic subcarrier shall be capable of accepting audio frequencies from 50 to 15,000 cycles.

g). The modulating signal for the stereophonic subcarrier shall be equal to the difference of the left and the right signals.

h). The pre-emphasis characteristics of the stereophonic subchannel shall be identical with those of the main channel with respect to phase and amplitude at all frequencies.

i). The sum of the side bands resulting from amplitude modulation

of the stereophonic subcarrier shall not cause a peak deviation of the main carrier in excess of 45 percent of total modulation (excluding SCA subcarrier) when only a left (or right) signal exists; simultaneously in the main channel, the deviation when only a left (or right) signal exists shall not exceed 45 percent of the total modulation (excluding SCA subcarriers).

j). Total modulation of the main carrier including pilot subcarrier and SCA subcarriers shall meet the requirements of Section 3.268 with maximum modulation of the main carrier by all SCA subcarriers limited to ten percent.

k). At the instant when only a positive left signal is applied, the main channel modulation shall cause an upward deviation of the main carrier frequency; and the stereophonic subcarrier and its sidebands signal shall cross the time axis simultaneously and in the same direction.

l). The ratio of peak main channel deviation to peak stereophonic subchannel deviation when only a steady state left (or right) signal exists shall be within plus or minus 3.5 percent of unity for all levels of this signal and all frequencies from 50 to 15,000 cycles.

m). The phase difference between the zero points of the main channel signal and the stereophonic subcarrier side bands envelope, when only a steady state left (or right) signal exists, shall not exceed plus or minus three degrees for audio modulation frequencies from 50 to 15,000 cycles.

NOTE: If the stereophonic separation between left and right stereophonic channels is better than 29.7 decibels at audio mod-

ulating frequencies between 50 and 15,000 cycles, it will be assumed that the above two paragraphs have been complied with.

n). Cross-talk into the main channel caused by a signal in the stereophonic subchannel shall be attenuated at least 40 decibels below 90 percent modulation.

o). Cross-talk into the stereophonic subchannel by a signal in the main channel shall be attenuated at least 40 decibels below 90 percent modulation.

p). For required transmitter performance, all of the requirements of Section 3.254 shall apply with the exception that the maximum modulation to be employed is 90 percent (excluding pilot subcarrier) rather than 100 percent.

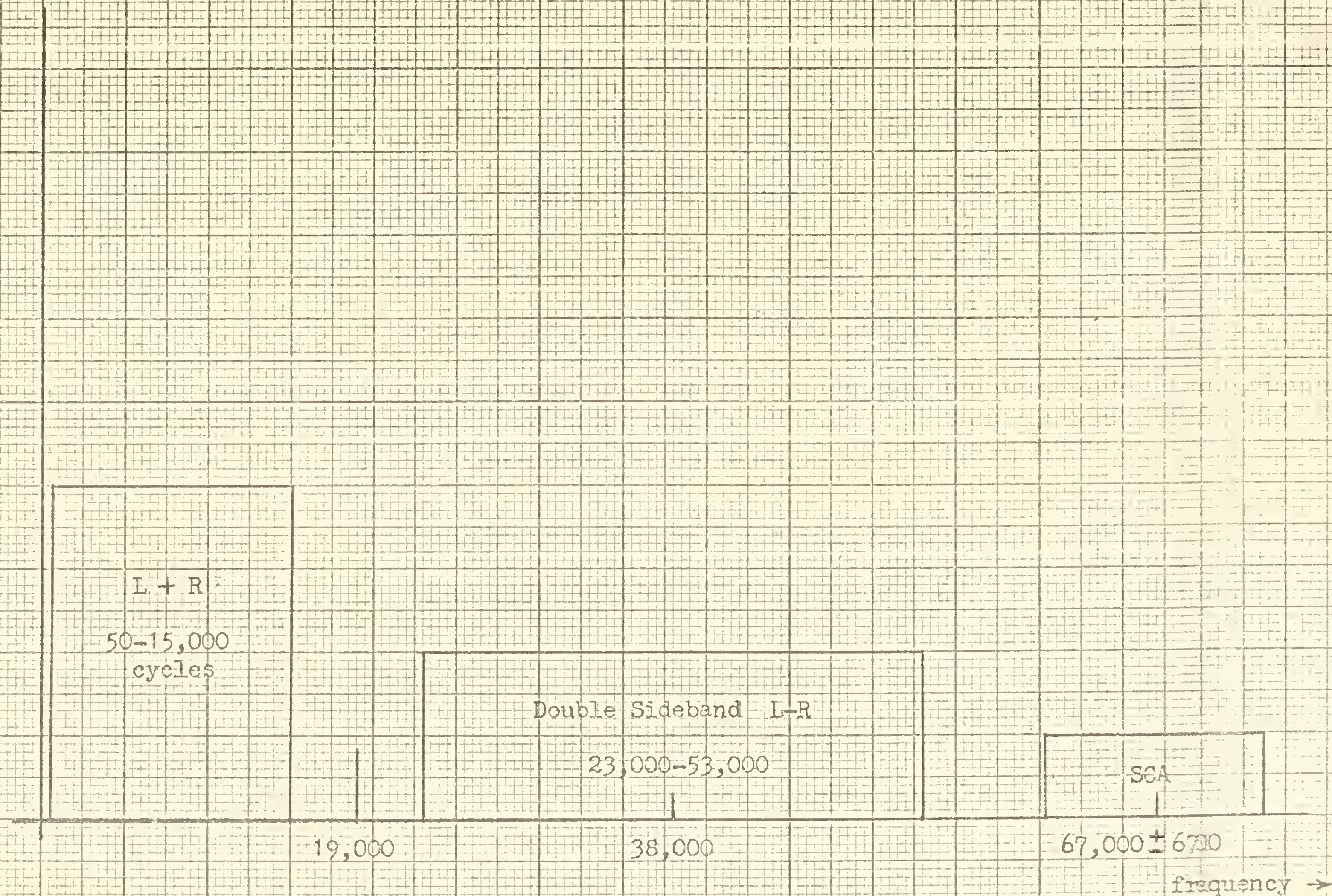
In addition to the above requirements, all the previous standards for SCA were to continue in effect. (In brief, SCA is a commercial signal transmitted on a multiplex basis along with the normal FM that has its subcarrier at 67,000 cycles, uses narrow band FM, and is limited to 67,000 cycles plus or minus 6,700 cycles. No more will be said about it other than the fact that it exists.)

The frequency spectrum of the full stereophonic FM broadcast signal would therefore appear as shown in Figure 1.

Mathematically, the composite signal (excluding the SCA) may be expressed as /4/:

$$(1) \quad E(t) = \left[L(t) + R(t)\right] + \left[L(t) - R(t)\right] \sin \omega_s t + A_{19} \sin 19{,}000 \times 2\pi t$$

where L(t) + R(t) represents the composite left and right sum channel signal comprising a frequency spectrum from 50 to 15,000 cycles (it is this signal that ordinary tuners not utilizing multiplex adapters



Frequency Spectrum of FM Stereophonic Signal

FIGURE 1

will use): $L(t) - R(t) \sin\omega_c t$ represents the difference channel comprising a frequency spectrum from 23,000 to 53,000 cycles (this signal may be thought of as normal amplitude-modulation with a suppressed carrier): and $A_{19}\sin 19{,}000 \times 2\pi t$ represents the 19,000 cycle pilot subcarrier. Writing equation (1) as a function of sine and cosine terms we have:

$$(2) \quad E(t) = E_{L+R}\cos(\omega_{L+R}t) + E'_{L-R}\cos(\omega_{L-R}t)\sin(\omega_s t) + A_{19}\sin 19{,}000 \times 2\pi t$$

or, expanding:

$$(3) \quad E(t) = E_{L+R}\cos(\omega_{L+R}t) + E_{L-R}\left[\sin(\omega_s t + \omega_{L-R}t) + \sin(\omega_s t - \omega_{L-R}t)\right]$$
$$+ A_{19}\sin 19{,}000 \times 2\pi t$$

understanding that in equations (2) and (3) we are using the simple expressions $\cos(\omega_{L+R}t)$ and $\sin(\omega_c t \pm \omega_{L-R}t)$ to represent a multitude of frequencies. From henceforth it shall be assumed that this equation represents the true transmitted signal (of course frequency-modulated about some frequency in the 88-108 Mc frequency range), thereby neglecting any possible sources of signal degradation produced at the transmitter.

CHAPTER III

STEREOPHONIC FM RECEPTION

Equation (3) represents the signals appearing at the multiplex output of an FM tuner after the station's signal has been converted to an intermediate frequency, amplified, limited, and detected.

NOTE: A multiplex output signal is not the same as an FM tuner output signal. Since all normal FM transmitted signals are pre-emphasised prior to transmission, the normal FM tuner output will provide de-emphasis to compensate for this and thereby achieve the original signal. For stereophonic reception this de-emphasis cannot be accomplished until after the demodulation of the "difference channel," therefore, the multiplex output of a tuner will provide the signal prior to de-emphasis.

By examination of equation (3), it can be seen that at least three basically different methods of separating the stereophonic signal are possible:

1). Use time-multiplexing techniques such as: the composite signal is applied to an AM detector (along with a 38,000 cycle reinserted carrier) to produce the channel L output, also, apply the composite signal to a second identical detector to produce the channel R output by inserting the 38,000 cycle reinserted carrier 180 degrees out of phase with the first.

2). Use a 23,000 to 53,000 cycle bandpass filter to separate the "difference channel," then a normal audio detector followed by the use of a sum-and-difference matrix network to obtain the L and R audio outputs.

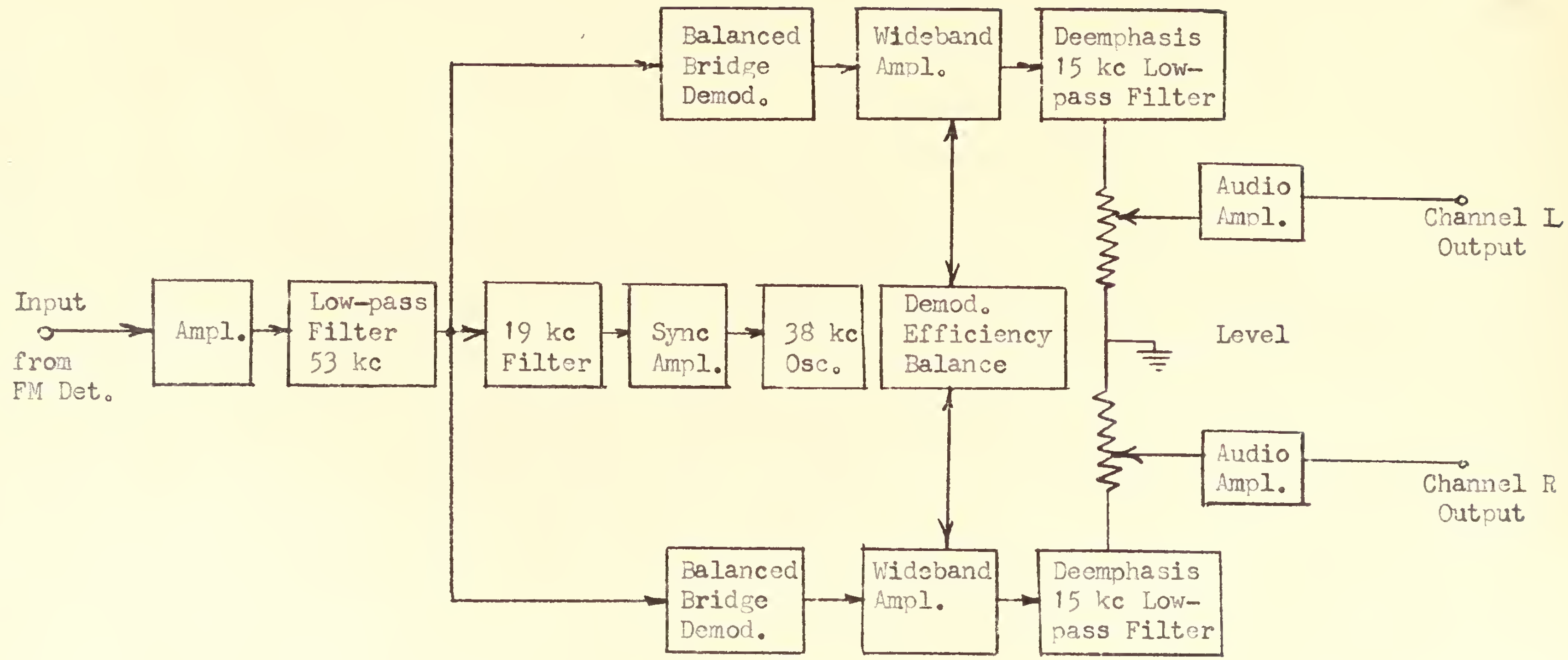
3). The system proposed by the author—use a product detector to demodulate the "difference channel," utilizing the de-emphasis requirements to act as the low pass filter for the product detector, and then use a sum-and-difference circuit to obtain the L and R audio outputs.

First discussed shall be the time-multiplexing techniques using the H. H. Scott multiplex adapter to demonstrate the basic concept. Second discussed shall be the method using filtering followed by audio detection as in the original General Electric system, going into some detail in problems associated with "filtering." Lastly discussed shall be the third basic type—product detection, going into elaborate detail as to technique, theory, design, problems, etc.

FM STEREOPHONIC RECEPTION UTILIZING TIME-MULTIPLEXING TECHNIQUES

Figure 2 is a block diagram of the H. H. Scott type 355 Multiplex Stereophonic Adapter /5/. The signal from the FM detector (multiplex output) is first amplified in a high input-impedance stage so that the low output voltage (0.3 volts rms at 75,000 cycle deviation) is increased to the proper level for detection. The low output impedance of this stage also allows a 53,000 cycle low pass filter to be driven. This filter removes frequencies above 53,000 cycles (SCA) which might be present between 60,000 to 75,000 cycles.

After the 53,000 cycle filter, the 19,000 cycle pilot subcarrier is selected out by a 19,000 cycle filter. This filter is made as narrow as possible to prevent the adapter's 38,000 cycle oscillator from being synchronized by either the sum or difference channel modulation components or by noise when listening to a weak station. The 19,000



Block Diagram of the H. H. Scott Multiplex Adapter

FIGURE 2

cycle signal is then amplified and injected into a 38,000 cycle oscillator circuit, phase locking this oscillator to the 19,000 cycle pilot subcarrier.

The outputs of the 53,000 cycle low-pass filter and 38,000 cycle oscillator drive the two balanced-bridge stereophonic demodulators which produce basically the L and R channels. The two wide-band amplifiers following the demodulators have a common efficiency-balance circuit which compensates for the demodulator's detection efficiency of sum and difference signals.

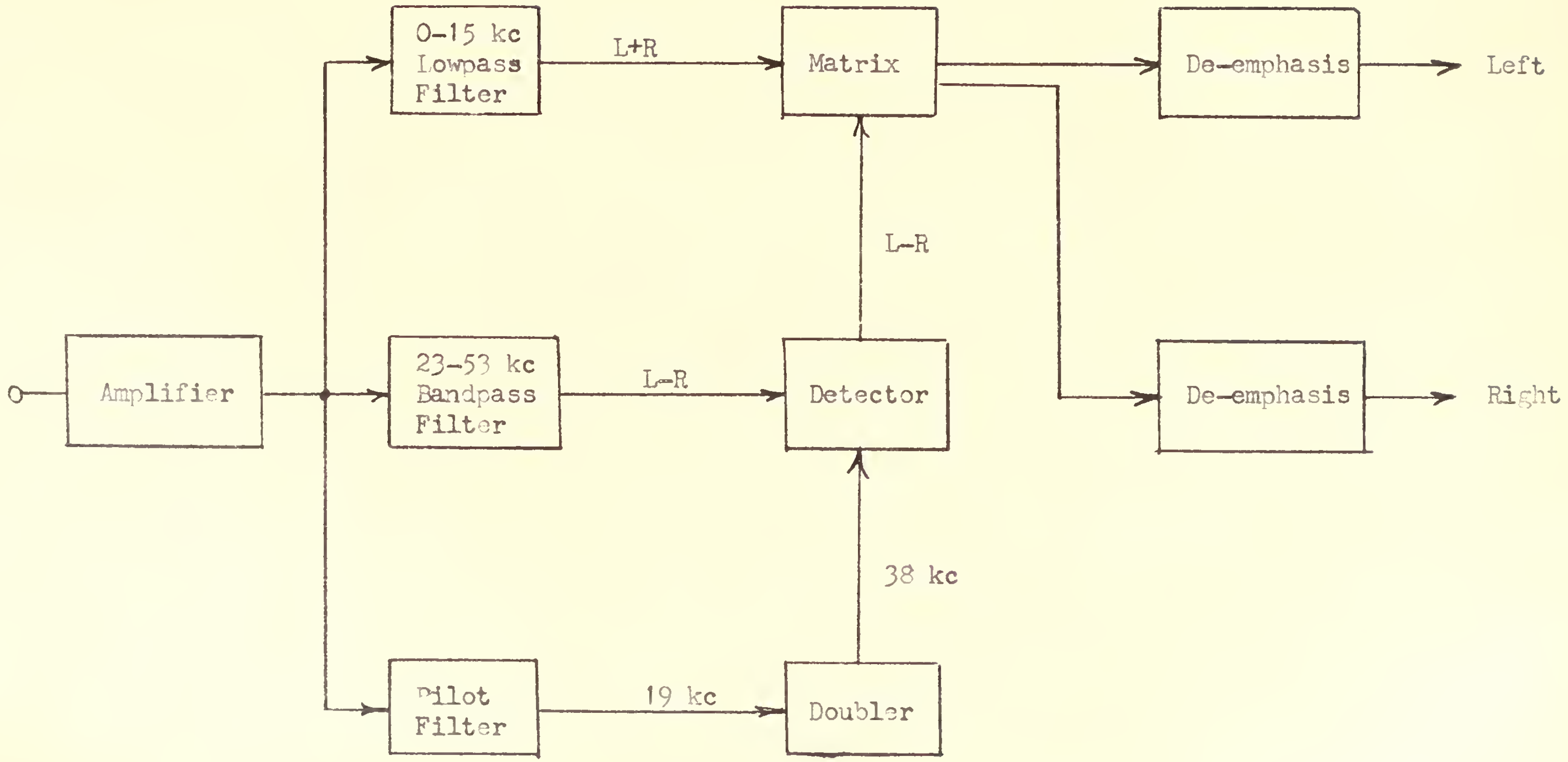
The outputs of the wide-band amplifiers drive 75-microsecond de-emphasis networks. The final audio amplifiers are used to bring the separate channels up to the proper signal level for feeding into the normal multiplex input of an audio amplifier for example, or a pre-amplifier.

FM STEREOPHONIC RECEPTION UTILIZING FILTERING AND MATRIXING TECHNIQUES

Figure 3 is a block diagram of the original General Electric multiplex adapter as considered by the FCC.[1] Figure 4 is a schematic diagram of an adapter utilizing this concept. No. values are shown except for necessary filter components.

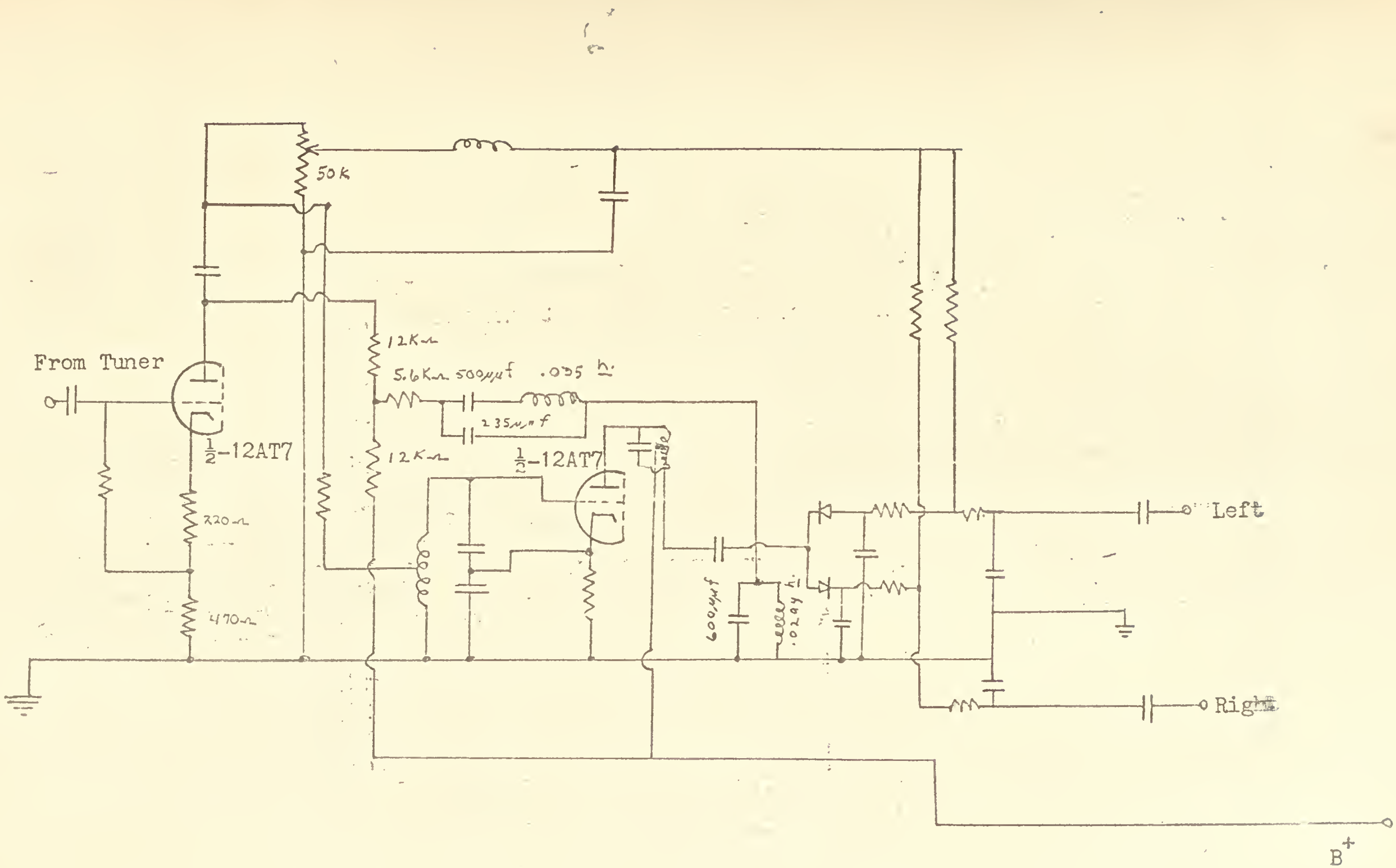
The signal from the FM detector is first amplified using a simple triode amplifier that is flat to 53,000 cycles. Next, the 19,000 cycle pilot subcarrier is filtered out by use of a 19,000 cycle frequency trap and then frequency doubled by use of a triode amplifier with a resonant plate load tuned to 38,000 cycles.

[1]Enclosure to General Electric letter to Lt. Norman W. Petersen dated August 16, 1961



Block Diagram of General Electric Multiplex Adapter

FIGURE 3



Schematic Diagram of General Electric Multuplex Adapter

FIGURE 4

Looking at the low-pass filter, the entire sum channel will pass with ideally no attenuation to a resistance matrix. Next, looking at the band-pass filter, only those frequencies from 23,000 to 53,000 cycles will pass to the diode detector, where, with the insertion of the frequency doubled 38,000 cycle signal, the difference channel will be demodulated (both plus L - R and minus L - R). These two signals are both feed to the resistance matrix where they combine to form the desired outputs:

(4) $(L+R) + (L-R) = 2L$

(5) $(L+R) - (L-R) = 2R$

Each channel then goes to a simple RC network (in actuality the RC network is a part of the resistance matrix) for the proper 75-microsecond de-emphasis. The resistance matrix, RC network is designed so as to compensate for any phase differences between the sum and difference channels created by the low-pass and band-pass filters so that the matrixing will be done with the signals in proper phase.

It is believed that this basic design multiplex adapter is the circuit the FCC had in mind when they approved the Zenith-General Electric system in that one of their reasons given was /1/:

> ...We do note, however, that the adapter for System 4-4A (i.e. the Zenith-General Electric system) recommended by the preponent of system 4A would be a relative small device which could be manufactured for a parts cost of less than eight dollars.

Let us analyze this circuit a little more closely though. General Electric /6/ has shown that:

(6) $$db_{separation} = 20 \log \frac{G_{L+R} + G_{L-R}}{G_{L+R} - G_{L-R}}$$

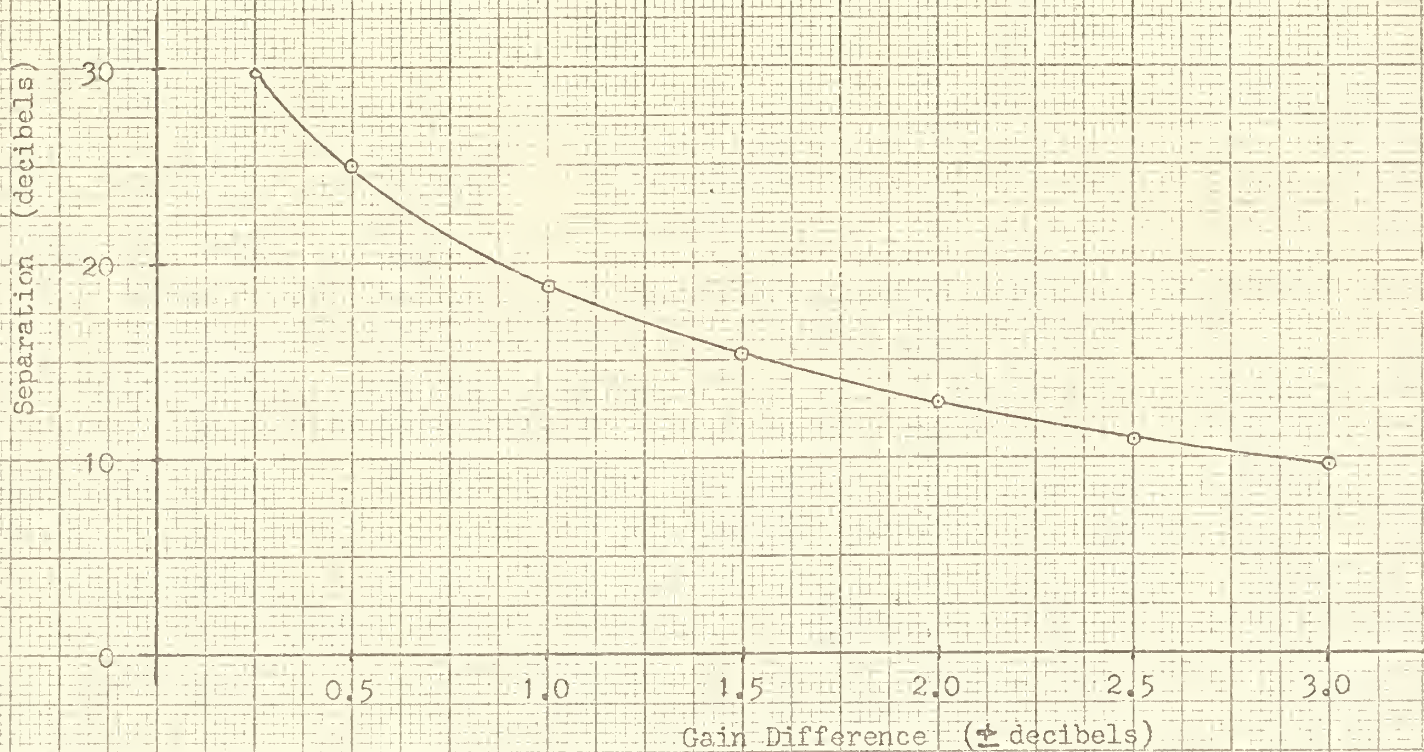
where G_{L-R} is gain variation in the difference channel, and G_{L+R} is gain variation in the sum channel. This formula is assuming that the phase difference between the sum channel audio and the difference channel side bands signal zero crossing equals zero. The FCC specifications state:

> The ratio of peak main channel deviation to peak stereophonic subchannel deviation when only a steady state left (or right) signal exists shall be within plus or minus 3.5 percent of unity for all levels of this signal and all frequencies from 50 to 15,000 cycles.

Maximum gain variation would therefore be when one subchannel had deviated plus 3.5 percent, while the other subchannel had deviated minus 3.5 percent, for a total of seven percent, or approximately 0.5 decibels. By using this maximum allowable deviation in equation (6) we have:

$$(7)\quad db_{separation} = 20 \log \frac{(1 + 1 \times 0.035) + (1 - 1 \times 0.035)}{(1 + 1 \times 0.035) - (1 - 1 \times 0.035)}$$

or, approximately 30 decibels. Figure 5 contains a plot of this equation for various gain differences. It can be seen that once this maximum deviation is exceeded, there is a rapid deterioration of separation. It therefore becomes mandatory that the gain in the sum channel and the difference channel be as flat as possible over the entire frequency spectrum in order to maintain a desirable separation figure. For example, assume that the transmitting station is producing a signal that is just within the maximum deviation range. Also, let us assume that the multiplex adapter is capable of demodulating the stereophonic signal within the same deviation range. The maximum plus gain variation for the transmission-demodulation path would therefore be:



Gain Variation Between Sum Channel and Difference Channel versus Separation

FIGURE 5

(8) $G^{+} = \left[(1 + 1 \times 0.035) + (1 + 1 \times 0.035)(0.035)\right]$

$= 1.071125$

For the maximum minus gain variation for the transmission-demodulating path, we have:

(9) $G^{-} = \left[(1 - 1 \times 0.035) - (1 - 1 \times 0.035)(0.035)\right]$

$= 0.931225$

Applying these two values to equation (6) we have:

(10) $db_{separation} = 20 \log \frac{1.071125 + 0.931225}{1.071125 - 0.931225} = 23.6 \text{ db}$

It can therefore be seen that the maximum separation that can be expected for a stereophonic signal that is demodulated by a multiplex adapter that is operating within the same limits that the FCC has imposed for transmitting is 23.6 decibels (assuming no phase variation). Before we compare this result to the maximum theoretical separation obtainable with the filtering-matrixing technique, let us also look at the phase variation specifications since phase variation and attenuation go hand in hand when discussing filters.

The FCC specifications state:

> The phase difference between the zero points of the main channel signal and the stereophonic subcarrier side bands envelope, when only a steady state left (or right) signal exists, shall not exceed plus or minus three degrees for audio modulating frequencies from 50 to 15,000 cycles.

Assuming that the gain difference between the sum and difference channels is zero, General Electric /6/ has shown that the decibel separation may be expressed as a function of the phase difference crossings for the sum and difference channels thusly:

(11) $db_{separation} = 20 \log \frac{\sin \phi}{1 - \cos \phi}$

where ϕ is the phase shift between the sum channel zero crossing and the difference channel zero crossing (of the side band envelope), assuming gain differences between the sum and difference channels equals zero.

Maximum separation will of course occur when ϕ equals zero degrees, at which time there would theoretically be an infinite separation. Assuming a signal is generated with the maximum allowable phase difference (ϕ equals three degrees), we have from equation (11):

(12) $db_{separation} = 20 \log \frac{\sin 3°}{1 - \cos 3°}$

$= 31.6 \text{ db}$

Figure 6 is a plot of this function for various phase differences.

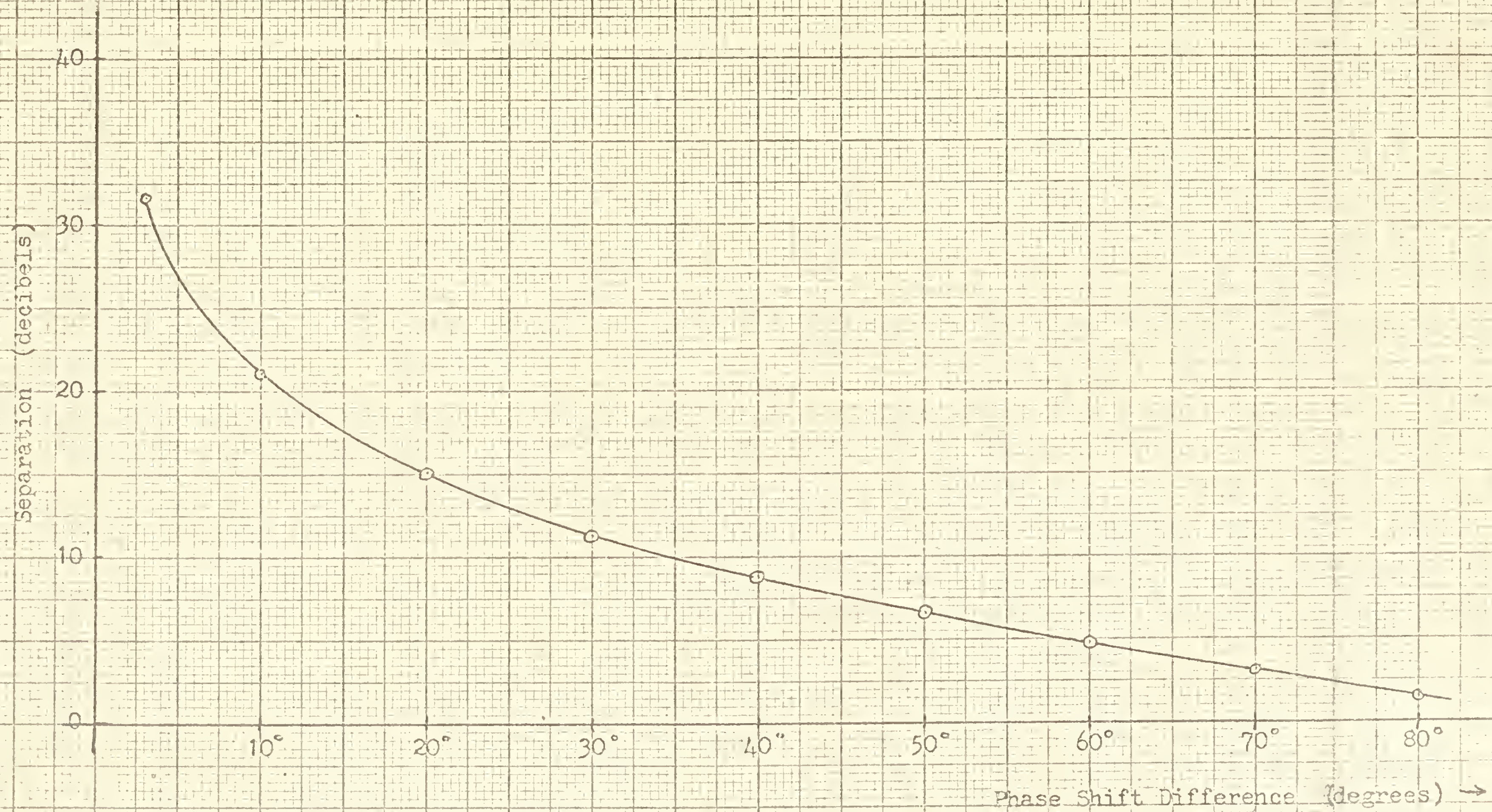
Again, let us assume that the multiplex adapter is capable of demodulating the stereophonic signal within the same allowable maximum phase difference. The maximum phase difference would therefore be six degrees, or ϕ equals six degrees. By substituting this value into equation (11) we have:

(13) $db_{separation} = 20 \log \frac{\sin 6°}{1 - \cos 6°}$

$= 25.6 \text{ db}$

It can therefore be seen that the maximum separation that can be expected for a stereophonic signal that is demodulated by a multiplex adapter that is operating within the same limits that the FCC has imposed for transmitting is somewhat less than 23.6 decibels.

We would therefore like to have the multiplex adapter designed



Phase Shift Difference Between Sum Channel and Difference Channel versus Separation

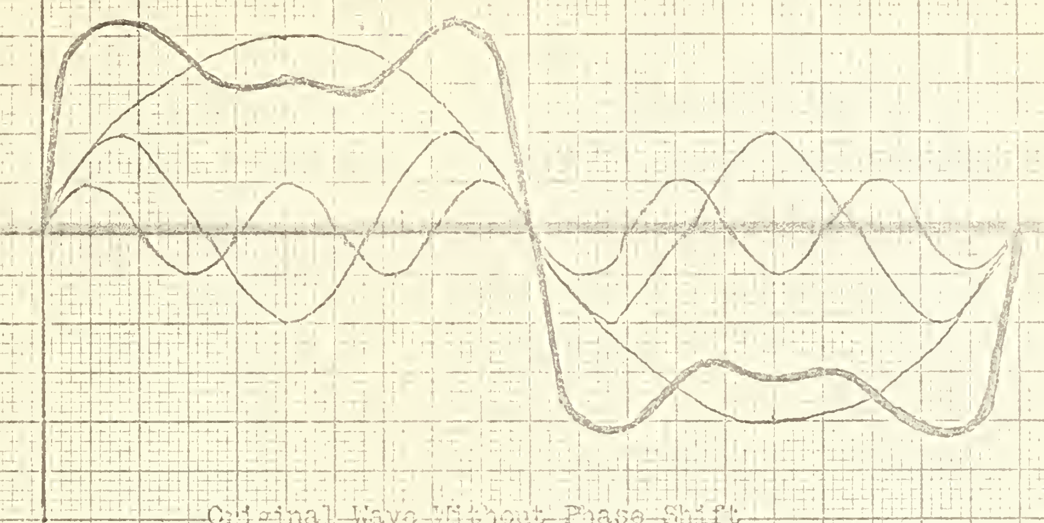
FIGURE 6

so as to maintain the same separation at the multiplex adapter that is maintained at the transmitter—gain variation to less than 3.5 percent, and phase variation to less than three degrees. For the moment, let us assume that the multiplex adapter is ideal in that all amplifiers and stages are absolutely "flat" across the frequency spectrum and do not introduce any phase variation. Then all we need concern ourselves with is the two required filters:

1). Low-pass; Cut-off frequency equal to 15,000 cycles, infinite attenuation at 23,000 cycles.

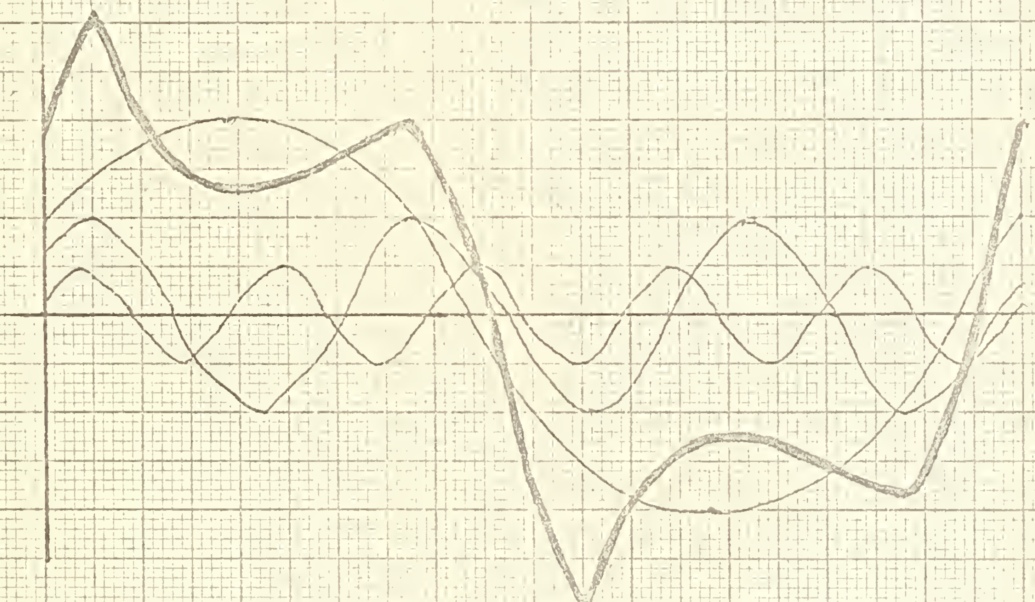
2). Band-pass; Cut-off frequencies at 23,000 cycles and 53,000 cycles, infinite attenuation at 15,000 cycles and also at 67,000 cycles.

It is furthermore required that the filters be flat within 0.5 decibels over the entire pass band, and at the same time be phase linear within three degrees. Before an attempt to design these filters is made, let us define what is meant by "phase linear." In usual plots of phase response of filters, the phase response versus frequency is plotted on a logarithmic scale. Plotted this way, no filter will look linear—all will follow a "tangent law" curve (or a sum of "tangent law" curves depending on the number of arms). But in actuality what is desired is a filter that, if any phase variation is present, all frequencies are shifted directly proportional to their frequency. For example, suppose we had a signal consisting of a fundamental frequency plus a third and fifth harmonic. (See figure 7.) The composite wave would almost appear as a square wave. Now, suppose that in passing through the filter each frequency were to be retarded by 22.5 degrees due to the phase response of the filter. The resultant wave form would

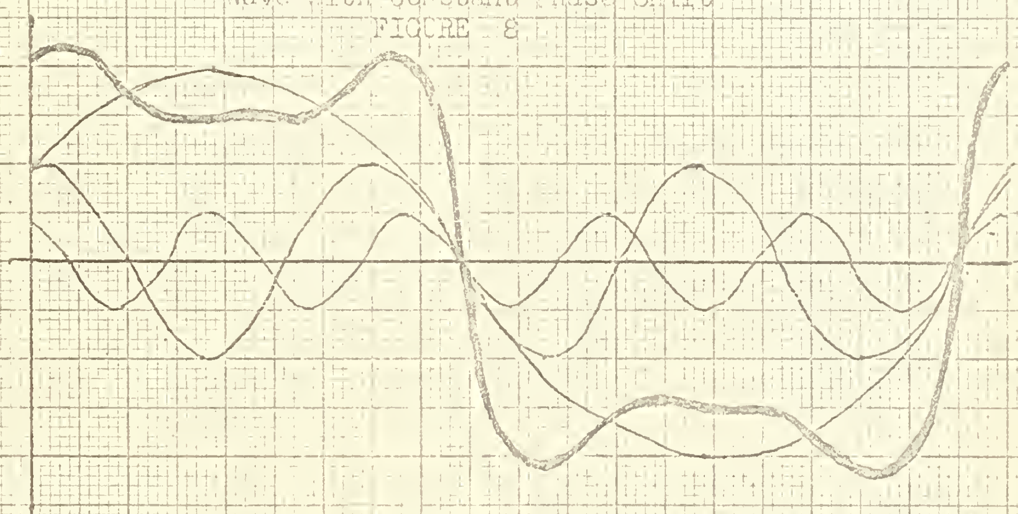


Original Wave Without Phase Shift

FIGURE 7



Wave With Constant Phase Shift
FIGURE 8



Wave With Linear Phase Shift

FIGURE 9

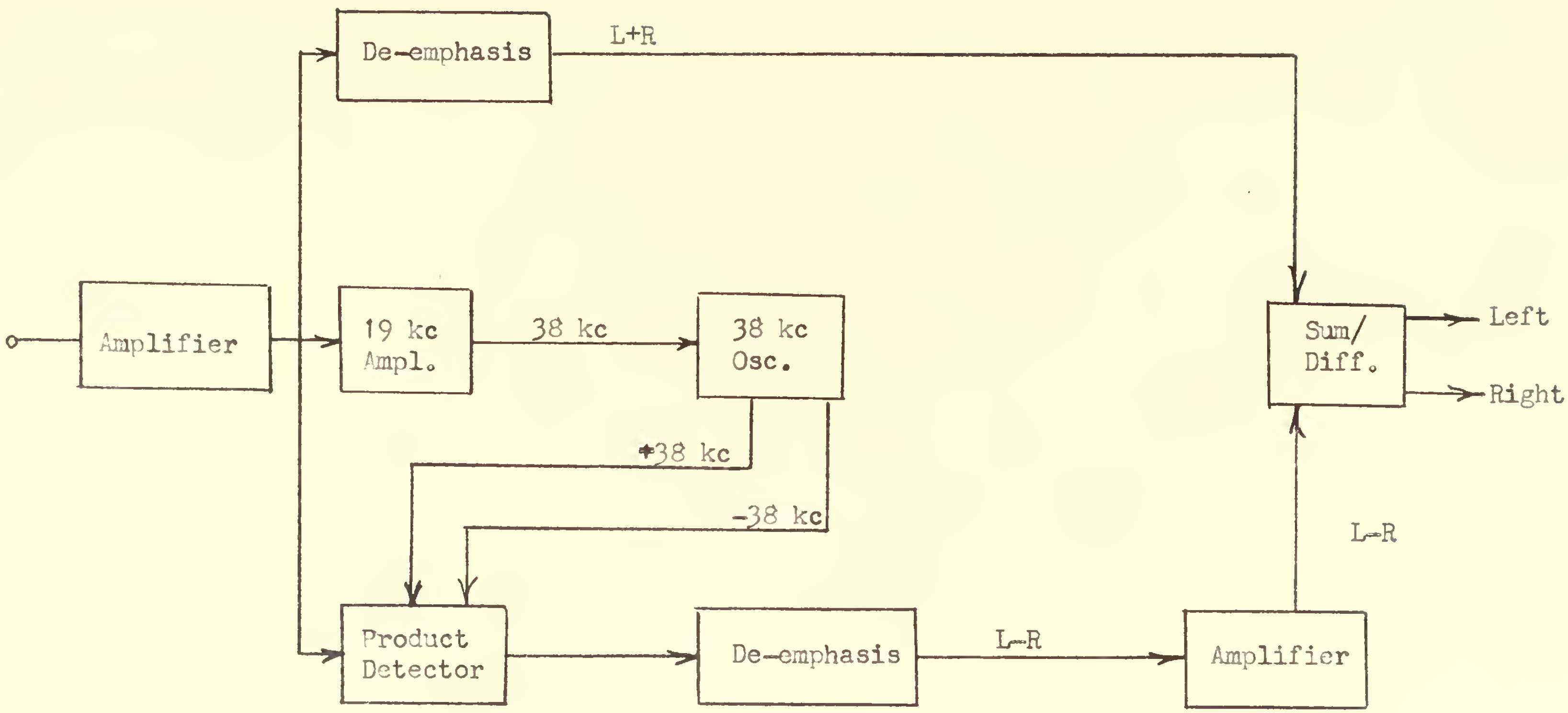
be that of figure 8. It is quite different from the original wave. Next, suppose that in passing through the filter, each frequency were to be retarded proportional to its frequency such that the fundamental is retarded by 22.5 degrees, the third harmonic by 67.5 degrees, and the fifth harmonic by 112.5 degrees. The resultant wave form would be that of figure 9. This wave form is identical to the original wave except that it is delayed in time by an amount equal to 22.5 divided by 360 times the frequency. A filter that has this type of phase response is defined to be a "phase linear" filter. To determine its linearity we can plot phase response versus frequency on a linear scale. A perfect phase linear filter would plot as a straight line. It is therefore quite easy to determine whether or not a filter is phase linear within three degrees by plotting the phase response as aforementioned, draw a straight line between the extremities of interest, and then examine the deviation of the curve from the straight line.

Reference 7 states that design of a low pass filter with the aforementioned characteristics would not be possible using common filter design techniques. Let us therefore consider what would be the effect of not filtering the sum channel. At the input to the matrix we would have the full frequency spectrum, but due to the de-emphasis we would have all frequencies above 15,000 cycles greatly attenuated. (See Appendix D for the frequency response of a 75-microsecond de-emphasis.) The only difficulty encountered might be some beat frequencies produced by the 67,000 cycle SCA or the 19,000 cycle pilot subcarrier. These components can be removed by simple LC combinations tuned to eliminate them. Then, after matrixing, the de-emphasis is used to attenuate all undesired higher frequency components.

For the band pass filter we are again faced with similar problems. If design of a low pass filter was impossible, then design of a band pass filter using similar criteria would be even more difficult. Since there is no way to avoid this problem with the General Electric circuit, compromises with either phase linearity, gain variation, or both must be made, and a corresponding loss of separation expected. Appendix B is a mathematical analysis of the General Electric band pass filter. Due to this one filter, the maximum separation possible over the full frequency spectrum would therefore be 13 decibels, considerably less than the desired amount. Looking carefully at the characteristics of the band pass filter though, it appears that the problem was realized by General Electric and that they actually designed the filter to be "flat" up to 10,000 cycles. Using this criterion, the separation over the range of 50 to 10,000 cycles is approximately 20 decibels. This again is still short of the desired amount, and of course any other non-linear variations of phase or gain would further decrease the separation.

FM STEREOPHONIC RECEPTION UTILIZING A PRODUCT DETECTOR

As shown in the previous section, in order to maintain maximum stereophonic separation between the Left and Right channel, the use of filters must be avoided. A block diagram of a multiplex adapter not using any filters is shown in figure 10. The signal from the FM detector (multiplex output) is first amplified so that the low output voltage is increased to the proper level for detection. From this amplifier, the signal goes to a special de-emphasis network. This de-emphasis network, designed by Norman H. Crowhurst /7/, is actually



Multiplex Adapter Using a Product Detector

FIGURE 10

a special twin-T filter. It is special in the sense that it is not symetric. By adjusting the various arms and the total series source impedance, its response (both phase and attenuation) from 50 to 15,000 cycles is made identical to a 75-microsecond RC de-emphasis, yet at the same time it provides a wide deep null over the 38,000 cycle region, giving a measured attenuation of 45 decibels at 38,000 cycles. See Appendix D for the response characteristics of this filter. The output of this filter will therefore clearly be only the sum channel.

Next, going back to the amplifier, we select out the 19,000 cycle pilot subcarrier using a 19,000 cycle filter. This pilot subcarrier is then amplified, frequency doubled by a full wave rectifier, and injected into a 38,000 cycle oscillator circuit, phase locking the oscillator to the 19,000 cycle subcarrier. The 38,000 cycle oscillator is so designed to provide both plus and minus 38,000 cycles. These two outputs plus the full frequency spectrum from the first amplifier then go to the product detector. The product detector used is quite different from the normal type of product detector. Under normal operation the output of a product detector is the product of each signal multiplied by the oscillator frequency injected (i.e. this will result in the frequency spectrum being transposed by an amount equal to the oscillator frequency), PLUS a linear amplification of all frequencies at the input. Normally, this linear amplification does not matter since the frequencies at the input are greatly separated from the audio output and are therefore filtered out by the audio low pass filter in the product detector output. However, for our application, the input will contain an undesired frequency band (i.e. the sum channel of 50 to 15,000 cycles) that will interfere with the audio difference

channel output since we have previously stated that we desire not to use any types of filters to remove the sum channel. It therefore becomes necessary to develop a new type of product detector that will generate products but at the same time will not provide any linear amplification. Such a product detector has been developed by the author and will be more fully discussed in Chapter IV.

Assuming we have the proper output out of the product detector, the signal goes to a twin-T de-emphasis network (identical to the previously mentioned asymmetric twin-T de-emphasis network) where the higher frequency components are all attenuated. The output of the network will therefore be the difference channel. This signal is then amplified to bring it up to the same level as the sum channel and then both it and the sum channel go to two paraphase amplifiers to produce the sum of the sum and difference channel (i.e. therefore the left channel) and the difference of the sum and the difference channel (i.e. therefore the Right channel).

CHAPTER IV

DESIGN OF A FM STEREO MULTIPLEX ADAPTER UTILIZING A PRODUCT DETECTOR

Chapter III discussed a block diagram of a multiplex adapter using a product detector. We shall next discuss the actual circuit design of such an adapter (See figure 11 for the schematic diagram.)

AUDIO AMPLIFIER

Since the signal input from the multiplex output of the tuner must follow three separate paths in the adapter, two of the paths (the 19,000 cycle pilot subcarrier and the signal input to the product detector) requiring considerable gain, it was decided to use a 12AT7 because of its high gain characteristics, and by using a fair amount of negative feedback produce a "flat" frequency response. For the a.c. equivalent circuit of such an amplifier we have:

$$\mu (e_g - i_p R_k) = i_p (R_p + R_L + R_K) \qquad (14)$$

or,

$$i_p = \frac{\mu e_g}{R_p + R_L + (\mu+1) R_K} \qquad (15)$$

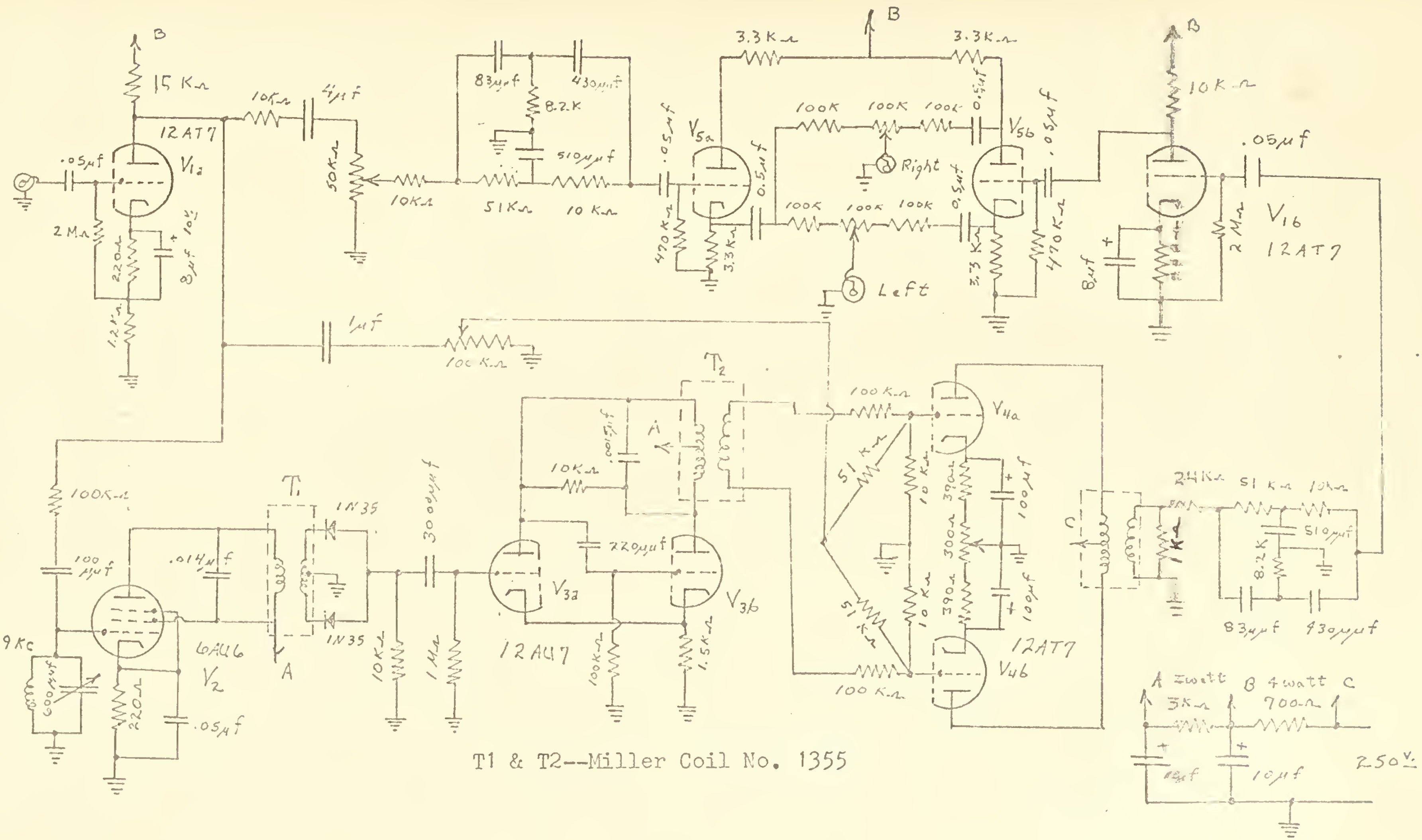
But,

$$\text{Gain} = \frac{i_p R_L}{e_g} \qquad (16)$$

$$= \frac{\mu R_L}{R_p + R_L + (\mu+1) R_K}$$

Using the values shown in figure 11, we have:

$$\text{Gain} = \frac{60 \times 15{,}000}{13{,}000 + 15{,}000 + (60+1)\, 1500} \qquad (17)$$



T1 & T2--Miller Coil No. 1355

Schematic Diagram of Multiplex Adapter Using a Product Detector

FIGURE 11

$= 7.5$

which should provide sufficient signal to drive the 19,000 cycle amplifier and the product detector.

Looking at the sum channel path, the signal goes to a twin-T de-emphasis network where all higher frequencies are greatly attenuated. (This was discussed in Chapter III.) From there it goes to a paraphase amplifier which will be discussed after the product detector.

19,000 CYCLE AMPLIFIER

From the first amplifier we also select off the 19,000 cycle pilot subcarrier using a toroid tank circuit sharply tuned to 19,000 cycles and a 100 micro-microfarad capacitor for light coupling. By placing a 100,000 ohm resistor in series in this line we can utilize the variable capacitor in the trap to change the relative phase of the 19,000 cycle pilot subcarrier and thereby adjust for any changes in phase throughout the signal path. The amplifier is simply a 6AU6 pentode with a tuned plate load in order to get as much gain as possible and at the same time reject all frequencies other than the 19,000 cycle pilot subcarrier.

The output of this tank circuit is coupled to a center-tapped coil with a diode in each arm. When the two diode outputs are tied together we have essentially a 38,000 cycle signal that is used to phase-lock the 38,000 cycle oscillator.

38,000 CYCLE OSCILLATOR

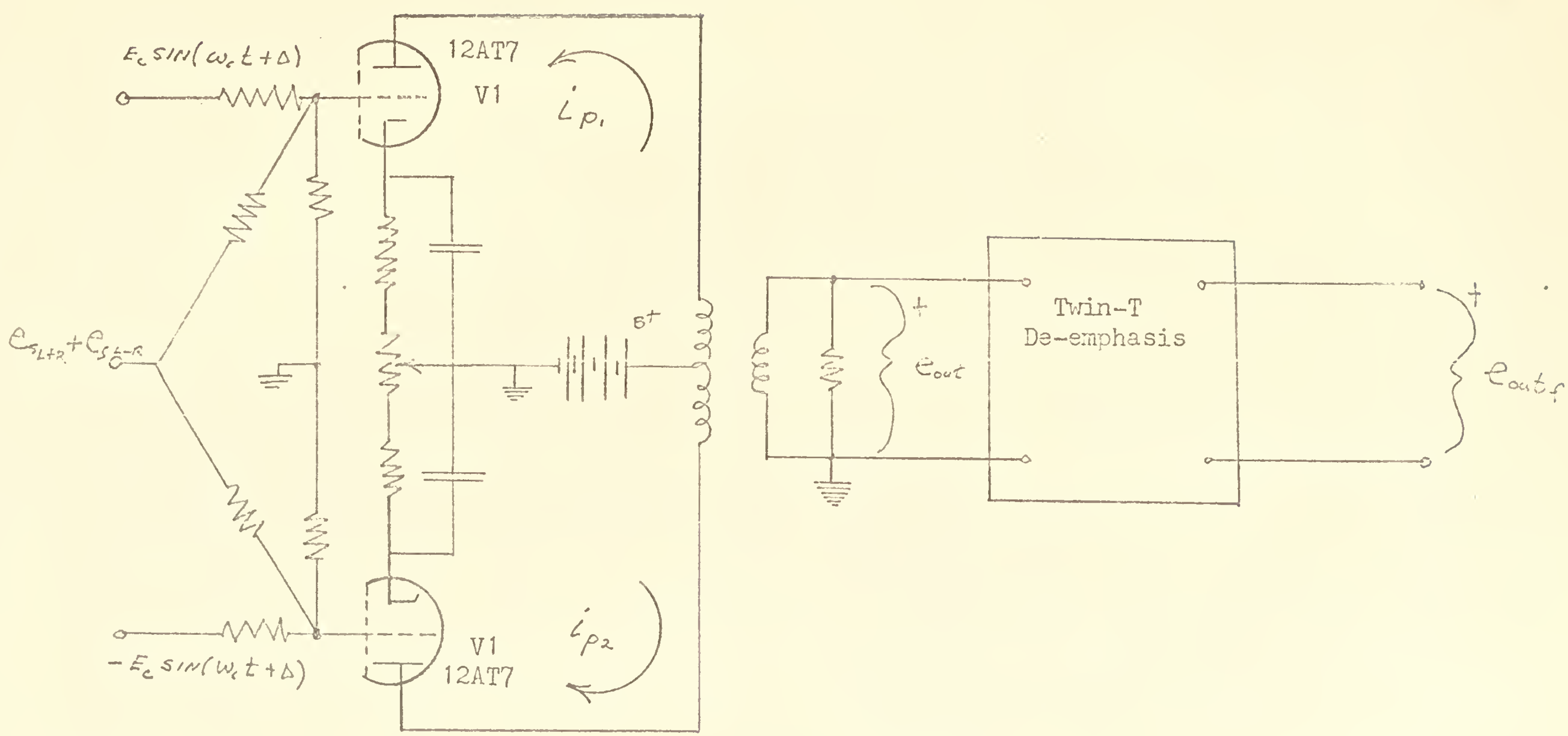
The signal output from the two diodes of the 19,000 cycle amplifier

is injected into the grid of a cathode coupled balanced oscillator thus locking both the frequency and the phase of the oscillator to the 19,000 cycle pilot subcarrier. The output of this oscillator has a floating ground so as to be able to produce both plus and minus 38,000 cycles.

BALANCED SQUARE LAW PRODUCT DETECTOR

We next inject the plus and minus 38,000 cycle phase-locked signal and the full frequency spectrum into the product detector. As previously mentioned, we require a product detector with the unique ability to form products yet not produce any pure linear amplification. Figure 12 is a detailed schematic diagram of just such a product detector. In essence, it is two square law devices operating push-pull such that all linear amplification is canceled in the transformer output (that is all in-phase linear amplification, not 180 degrees out-of-phase amplification). By feeding the 38,000 cycle carrier frequency 180 degrees out-of-phase at the two grids we can see that the squared terms will have <u>plus</u> twice the cross product in one tube while in the other tube we will have <u>minus</u> twice the cross product (of course all other terms in each tube will be identical). Due to the subtracting action of the transformer, the only audio output would therefore be four times the cross product. This will be shown in more detail in the following paragraphs.

For the transfer characteristic of a vacuum tube, the signal component of the plate current can be expressed in terms of a power series involving the input signal voltage. Thus:



Balanced Square Law Product Detector

FIGURE 12

(18) $$i_p = a_0 + a_1 e_s + a_2 e_s^2 + a_3 e_s^3 + \cdots$$

or, in terms of an a.c. voltage developed across the plate transformer:

(19) $$e_p = A_1 e_s + A_2 e_s^2 + A_3 e_s^3 + \cdots$$

The signals on the grid of tube V1 are:

(20) $$e_c = E_c \sin(\omega_c t + \Delta)$$

(21) $$e_s = e_{s_{L+R}} + e_{s_{L-R}}$$

where Δ is the variation in phase and/or frequency between the oscillator frequency and the true carrier frequency. Since:

(22) $$e_{s_{L+R}} = E_a \cos(\omega_{L+R} t)$$

and,

(23) $$e_{s_{L-R}} = E_s \left[\sin(\omega_c t + \omega_{L-R} t) + \sin(\omega_c t - \omega_{L-R} t)\right]$$

we may state that the total signal on the grid of tube V1 is:

(24) $$e_{g1} = E_c \sin(\omega_c t + \Delta) + E_a \cos(\omega_{L+R} t) + E_s\left[(\sin(\omega_c t + \omega_{L-R} t) + \sin(\omega_c t - \omega_{L-R} t)\right]$$

Similarly, the total signal on the grid of tube V2 is:

(25) $$e_{g2} = -E_c \sin(\omega_c t + \Delta) + E_a \cos(\omega_{L+R} t) + E_s\left[\sin(\omega_c t + \omega_{L-R} t) + \sin(\omega_c t - \omega_{L-R} t)\right]$$

The voltage developed at the plate transformer output is:

(26) $$e_{out} = e_{p1} - e_{p2}$$

Looking at the voltage developed across the plate transformer

output due to the first harmonic, we have:

$$(27)\quad e_{out} = A_{11}\left\{E_c \sin(\omega_c t+\Delta) + E_a \cos(\omega_{L+R} t) + E_s\left[\sin(\omega_c t+\omega_{L-R} t)+\sin(\omega_c t-\omega_{L-R} t)\right]\right\}$$
$$-A_{12}\left\{-E_c \sin(\omega_c t+\Delta) + E_a \cos(\omega_{L+R} t) + E_s\left[\sin(\omega_c t+\omega_{L-R} t)+\sin(\omega_c t-\omega_{L-R} t)\right]\right\}$$

and assuming that tubes V1 and V2 are balanced such that A_{11} and A_{12} are equal, equation (27) simplifies to:

$$(28)\quad e_{out} = 2A_1 E_c \sin(\omega_c t+\Delta)$$

Next, looking at the voltage developed due to the second harmonic, we have:

$$(29)\quad e_{s1}^2 = E_s^2\left[\sin(\omega_c t+\omega_{L-R} t)+\sin(\omega_c t-\omega_{L-R} t)\right]^2 + E_c^2 \sin^2(\omega_c t+\Delta)$$
$$+2E_c E_s\left[\sin(\omega_c t+\omega_{L-R} t)\sin(\omega_c t+\Delta) + \sin(\omega_c t-\omega_{L-R} t)\sin(\omega_c t+\Delta)\right]$$
$$+2E_a E_s\left[\cos(\omega_{L+R} t)\sin(\omega_c t+\omega_{L-R} t) + \cos(\omega_{L+R} t)\sin(\omega_c t-\omega_{L-R} t)\right]$$
$$+2E_a E_c\left[\cos(\omega_{L+R} t)\sin(\omega_c t+\Delta)\right] + E_a^2\cos^2(\omega_{L+R} t)$$

and similarly:

$$(30)\quad e_{s2}^2 = E_s^2\left[\sin(\omega_c t+\omega_{L-R} t)+\sin(\omega_c t-\omega_{L-R} t)\right]^2 + E_c^2 \sin^2(\omega_c t+\Delta)$$
$$-2E_c E_s\left[\sin(\omega_c t+\omega_{L-R} t)\sin(\omega_c t+\Delta) + \sin(\omega_c t-\omega_{L-R} t)\sin(\omega_c t+\Delta)\right]$$
$$+2E_a E_s\left[\cos(\omega_{L+R} t)\sin(\omega_c t+\omega_{L-R} t) + \cos(\omega_{L+R} t)\sin(\omega_c t-\omega_{L-R} t)\right]$$
$$-2E_a E_c\left[\cos(\omega_{L+R} t)\sin(\omega_c t+\Delta)\right] + E_a^2\cos^2(\omega_{L+R} t)$$

Subtracting, and solving for e_{out} we have (assuming A_{21} equals A_{22}):

$$(31)\quad e_{out} = A_2\Big\{4E_cE_s\big[\sin(\omega_c t + \omega_{L-R}t)\sin(\omega_c t + \Delta) + \sin(\omega_c t - \omega_{L-R}t)\sin(\omega_c t + \Delta)\big]$$
$$+ 4E_aE_c\big[\cos(\omega_{L+R}t)\sin(\omega_c t + \Delta)\big]\Big\}$$

Expanding this by using simple trignometric identities, we have:

$$(32)\quad e_{out} = A_2\Big\{2E_cE_s\big[\cos(\omega_{L-R}t - \Delta) - \cos(2\omega_c t + \omega_{L-R}t + \Delta) + \cos(\omega_{L-R}t + \Delta)$$
$$- \cos(2\omega_c t - \omega_{L-R}t + \Delta)\big] + 2E_aE_c\big[\sin(\omega_c t + \omega_{L+R}t + \Delta) + \sin(\omega_c t + \Delta - \omega_{L+R}t)\big]\Big\}$$

or, the total voltage out at the transformer output due to the first and second harmonic is therefore:

$$(33)\quad e_{out} = 2A_1E_c\sin(\omega_c t + \Delta) + A_2\Big\{4E_cE_s\big[\cos(\omega_{L-R}t)\cos\Delta\big]$$
$$- 2E_cE_s\big[\cos(2\omega_c t + \omega_{L-R}t + \Delta) + \cos(2\omega_c t - \omega_{L-R}t + \Delta)\big]$$
$$+ 2E_aE_c\big[\sin(\omega_c t + \omega_{L+R}t + \Delta) + \sin(\omega_c t + \Delta - \omega_{L+R}t)\big]\Big\}$$

This entire signal is then applied to the input of the twin-T de-emphasis network (see Appendix D for characteristics of network). As previously stated, the network is designed such that it will not pass any frequencies above 15,000 cycles while at the same time providing the proper de-emphasis (75-microseconds) and also a wide "notch" at 38,000 cycles.

Analyzing the frequency spectrum at the input to the de-emphasis network we can see that each component can possess the following bands of frequencies (assuming that the 19,000 cycle pilot subcarrier is the highest frequency of $e_{s_{L+R}}$):

(34)	ω_c	38,000 cycles
(35)	ω_{L-R}	50-15,000

(36) $2\omega_c + \omega_{L-R}$ 76,000-91,000 cycles

(37) $2\omega_c - \omega_{L-R}$ 61,000-76,000

(38) $\omega_c + \omega_{L+R}$ 38,000-57,000

(39) $\omega_c - \omega_{L+R}$ 19,000-38,000

It can therefore be seen that the only output of the de-emphasis network will be the 50 to 15,000 cycles components. Therefore:

(40) $$e_{out_f} = 4 A_2 E_c E_s \cos(\omega_{L-R} t) \cos \Delta$$

Of course, when the oscillator is operating at the correct frequency and in the proper phase, Δ will be zero and the output will therefore be:

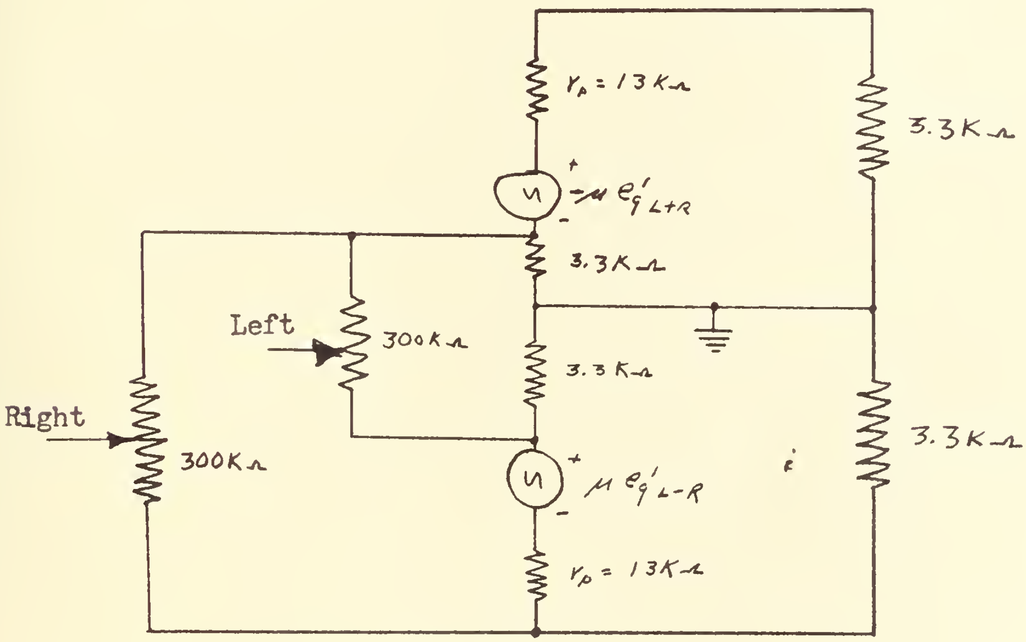
(41) $$e_{out_f} = 4 A_2 E_c E_s \cos(\omega_{L-R} t)$$

Or, as can be seen, the output of the de-emphasis will be simply the demodulated difference channel. (For further analysis of the third harmonic see Appendix C.)

SUM/DIFFERENCE AMPLIFIER

We now have the two signals seperated and demodulated such that we may represent one as L + R and the other as L − R. Mathematically, all we need do is add the L − R signal to the L + R and produce twice the Left signal, and subtract the L − R signal from the L + R signal to produce twice the Right signal. Using two paraphase amplifiers as shown in figure 11 we can do this. Looking at figure 13, an a.c. equivalent circuit for the two amplifiers, we can see by applying the

super-position theorem, that the signal voltage produced across each of the 300,000 ohm potentiometer will be essentially one-half the signal applied to the grid, or, $\frac{1}{2}(L + R)$ for the $L + R$ amplifier, while for the $L - R$ channel (again using super-position), one 300,000 ohm potentiometer would essentially have minus $\frac{1}{2}(L - R)$ while the other would have plus $\frac{1}{2}(L - R)$. The output of one potentiometer would therefore be the Left channel, while the other would be the Right channel.



Equivalent Circuit of the Sum/Difference Amplifier Circuit

FIGURE 13

CHAPTER V

ACTUAL OPERATION OF PRODUCT DETECTOR MULTIPLEX ADAPTER

Considerable difficulty was encountered in construction and operation of the prototype multiplex adapter. In order to check the operation of the adapter it is necessary to have a multiplex signal. Although there are multiplex test generators in existance for this purpose (for example, H. H. Scott Type 830 Multiplex Stereo Generator, approximately $800.00), procurement of one locally was not possible. It was therefore decided to attempt reception of one of the two FM stereophonic transmitting stations in Northern California. The closest station, KSJO (92.3 Mc) is located approximately seventy airmiles away across a mountain range. The other station, KPEN (101.2 Mc), is located approximately 100 airmiles away. Although the FCC /1/ has stated that satisfactory reception beyond 23 miles for the average tuner or 61 miles for an outstanding tuner cannot be expected, it was decided to attempt to utilize this source since no other signal source appeared available

To further complicate the problem, no adequate antenna system was available, nor was a suitable tuner available. The antenna used was a single folded dipole with approximately 100 feet of cable leading to the tuner. The tuner was a 1945 Zenith model which most certainly did not have sufficient band width. Never-the-less, an attempt was made over a period of several months to utilize the existing equipment.

Reception of KPEN proved to be impossible, but KSJO could be located and was of sufficient strength to phase-lock the 38,000 cycle oscillator. However, the tuner was so noisy that no analysis of

operation could be made—the noise level of the difference channel was higher than the signal level. The only measurement that could be made was that the lock-in range of the 38,000 cycle oscillator was from 35,256 to 41,010 cycles—more than adequate.

In a further attempt to obtain proper operation it was decided to move all the experimental work out to La Mesa Village where the La Mesa Television Company kindly agreed to provide a tap off their cable-vision service. Furthermore, at this time a Heathkit model PT-1 wideband stereo tuner (with a sensitivity of two microvolts for 20 decibels of quieting) was made available. With these two improvements to the system, it was only a matter of minutes before the adapter was operating properly. The noise level was found to vary from inaudible to barely perceptible.

Several attempts were made to measure the separation between the two channels, but due to the short duration of any test signals sent by the transmitting station (approximately five seconds) it was very difficult to properly balance the sum channel, the difference channel, the Left channel, and the Right channel. Readings were taken for three different attempts to balance all channels, the maximum separation reading being 19.2 decibels. Although this appears quite close to the maximum separation that should be expected (as shown in Chapter III), it is not a fair comparison. The figure of 23 decibels of separation as developed in Chapter III was for the entire audio range, 50 to 15,000 cycles, whereas the aforementioned reading of 19.2 decibels was only over a limited audio range.

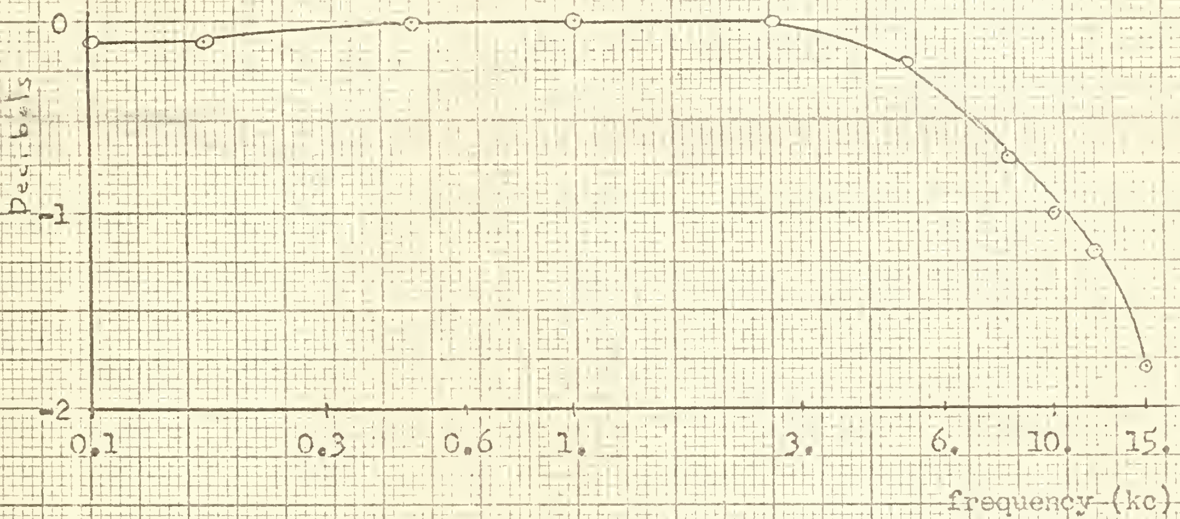
The most critical component of the multiplex adapter is the push-pull transformer of the product detector. Any non-linear response

in it will produce a considerable loss in separation as previously shown. The transformer used was a 60 cycle power transformer, the secondary provided with a center-tap. By hooking-up the transformer in reverse it was found that it could be used. To check to see if the transformer was producing any appreciable gain variation or phase shift over the audio range, a plot of the characteristics of the transformer was made. Figure 14 and figure 15 are plots of these characteristics.

As can be seen, the gain variation was satisfactory over the range of 50 to 8,000 cycles, but the phase variation was completely unsatisfactory. It could only meet the required phase variation of plus or minus three degrees over the range of 50 to 2,200 cycles. The total phase variation of 40.2 degrees, or, plus or minus 20.2 degrees would mean a maximum separation of 15 decibels.

By further experimenting with the transformer it was found that the phase variation could be reduced to plus or minus three degrees (with a corresponding improvement in gain variation) by increasing the 1000 ohm load resistor to 10,000 ohms. This was tried, but it was found that it produced the undesired effect of moving the operating point of the Balanced Square Law Product Detector to an area where it did not operate well as a square law device. The net result was a loss in separation.

Based upon the separation achieved for various arrangements of balance and transformer loading, it appears that the determining factor for the amount of separation obtainable is the push-pull transformer for the product detector. In most cases the separation obtained was very closely related to the phase and gain variation of the transformer.



Voltage Gain versus Frequency for Push-Pull Transformer

FIGURE 14

Phase Response versus Frequency for Push-Pull Transformer

FIGURE 15

It therefore appears that with a push-pull transformer that is properly balanced, has the correct turns ratio to provide the proper A.C. load to the tubes, and has a flat gain and phase response from 50 to 15,000 cycles, it would be possible to achieve at least 23 or more decibels separation over the entire audio range. Finding such a transformer would simply be a matter of checking any of numerous commercially available hi-fi transformers that are currently available.

CHAPTER VI

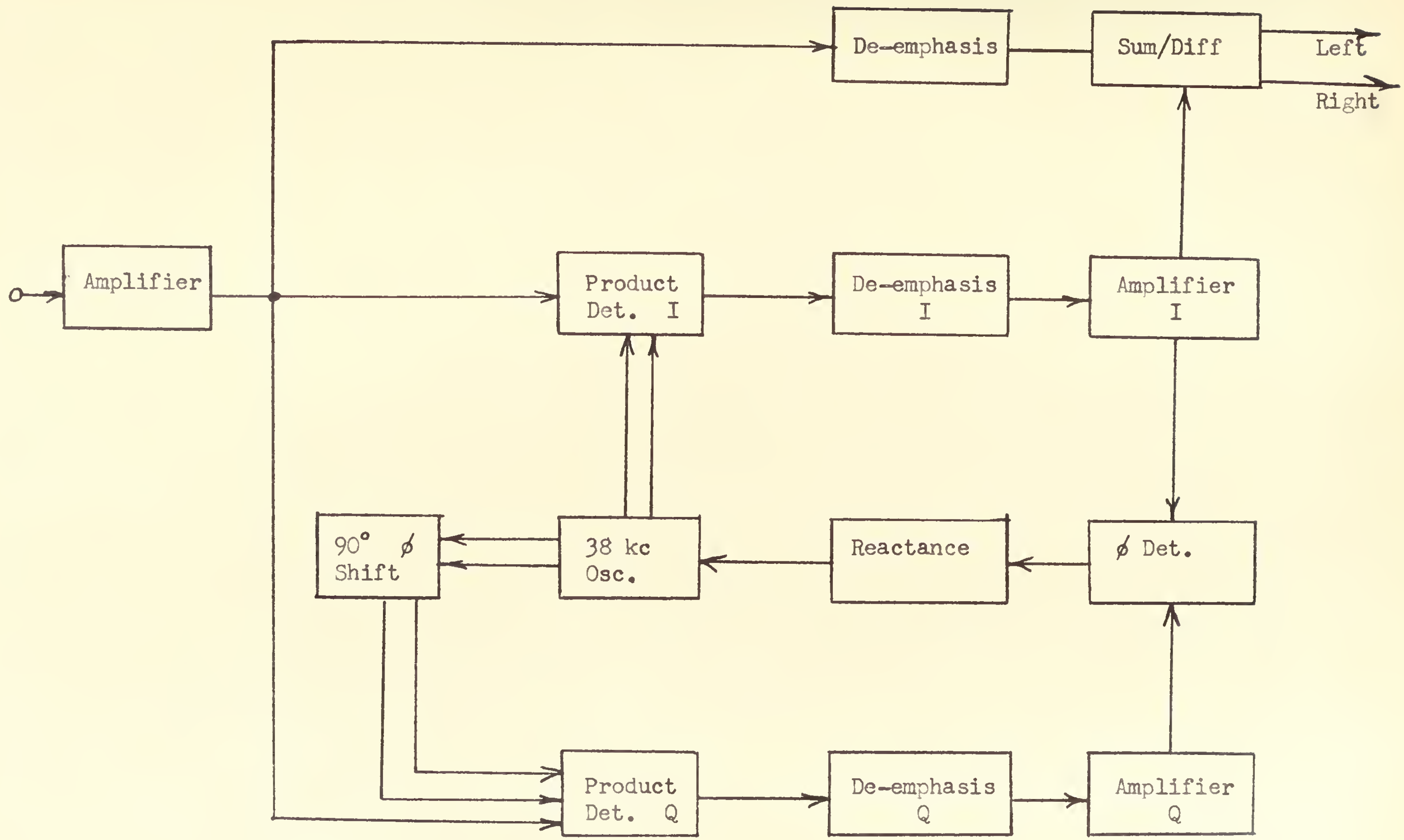
DESIGN OF A MULTIPLEX ADAPTER USING SYNCHRONOUS DETECTION

In checking the operation of the previously discussed multiplex adapter it was found that considerable amplitude variation (and at times even phase variation) existed in the 19,000 cycle pilot subcarrier. This amplitude variation appeared to be a function of the modulation. Upon checking several other commercially made multiplex adapters it was found that this was an ailment common to them all. Although not verified, this "jitter" is probably due to audio frequency components of approximately 19,000 cycles in either channel and to the higher order power spectral density terms which result from frequency modulation. Such variation could be a source of noise generation, gain variation (since the gain is dependent upon the 38,000 cycle amplitude which in turn is dependent upon the phase-lock signal amplitude), or phase variation.

In order to eliminate this possible source of noise and/or distortion, it was decided to attempt to modify the previously discussed circuit to enable product detection WITHOUT the 19,000 cycle pilot subcarrier. Since the difference channel is actually Double Sideband Surpressed Carrier, a system that has previously been used for radio communication is synchronous detection. Figure 16 is a block diagram of such a system. Going back to equation (40) of Chapter IV, we will recall that:

$$(40) \quad e_{out} = 4 A_2 E_c E_s \cos(\omega_{L-R} t) \cos \Delta$$

which was the output of the de-emphasis network, where Δ was



Multiplex Adapter Using Synchronous Detection

FIGURE 16

the variation in phase and/or frequency between the oscillator frequency and the true carrier frequency. For the moment let us consider it to be just the phase variation (assuming the frequency is correct). We can therefore see that if the phase is correct (i.e. Δ equals zero degrees) the output will be:

$$(42) \quad e_{out} = 4 A_2 E_c E_s \cos(\omega_{L-R} t)$$

but, if it were to be 90 degrees out of phase we would therefore have:

$$(43) \quad e_{out} = 4 A_2 E_c E_s \cos(\omega_{L-R} t) \cos 90^\circ$$

$$= 0$$

Therefore, by using a second product detector, de-emphasis network, and amplifier circuit, henceforth referred to as the Q channel, to which the oscillator frequency is injected 90 degrees out of phase with the existing product detector, de-emphasis network, and amplifier circuit, henceforth referred to as the I channel, with all other parameters remaining the same, we would have no output when the correct 38,000 cycle carrier frequency were to be injected into the I channel. In the event the carrier frequency is not at the proper phase, then the difference between the I channel output and the Q channel output will be proportional to the phase deviation Δ. As shown in the block diagram, we inject both the I channel signal and the Q channel signal into a phase detector to detect this difference.

Weaver /8/, and Wood and Whyland /9/, have shown that a gated rectifier phase detector performs in a manner similar to a product

multiplier which has the additional property of removing the audio frequency components and only passing the direct current component. The output of the phase detector then becomes:

$$(44)\quad e_{\phi} = e_{g_1} \times e_{g_2} = 16 A_2^2 E_c^2 E_s^2 \cos^2(\omega_{L-R}\, t) \sin\Delta \cos\Delta$$

$$= A'\left\{\sin 2\Delta + \tfrac{1}{2}\left[\sin(2\omega_{L-R} t + 2\Delta) - \sin(2\omega_{L-R} t - 2\Delta)\right]\right\}$$

But due to the inherent elimination of the audio frequency components, this becomes:

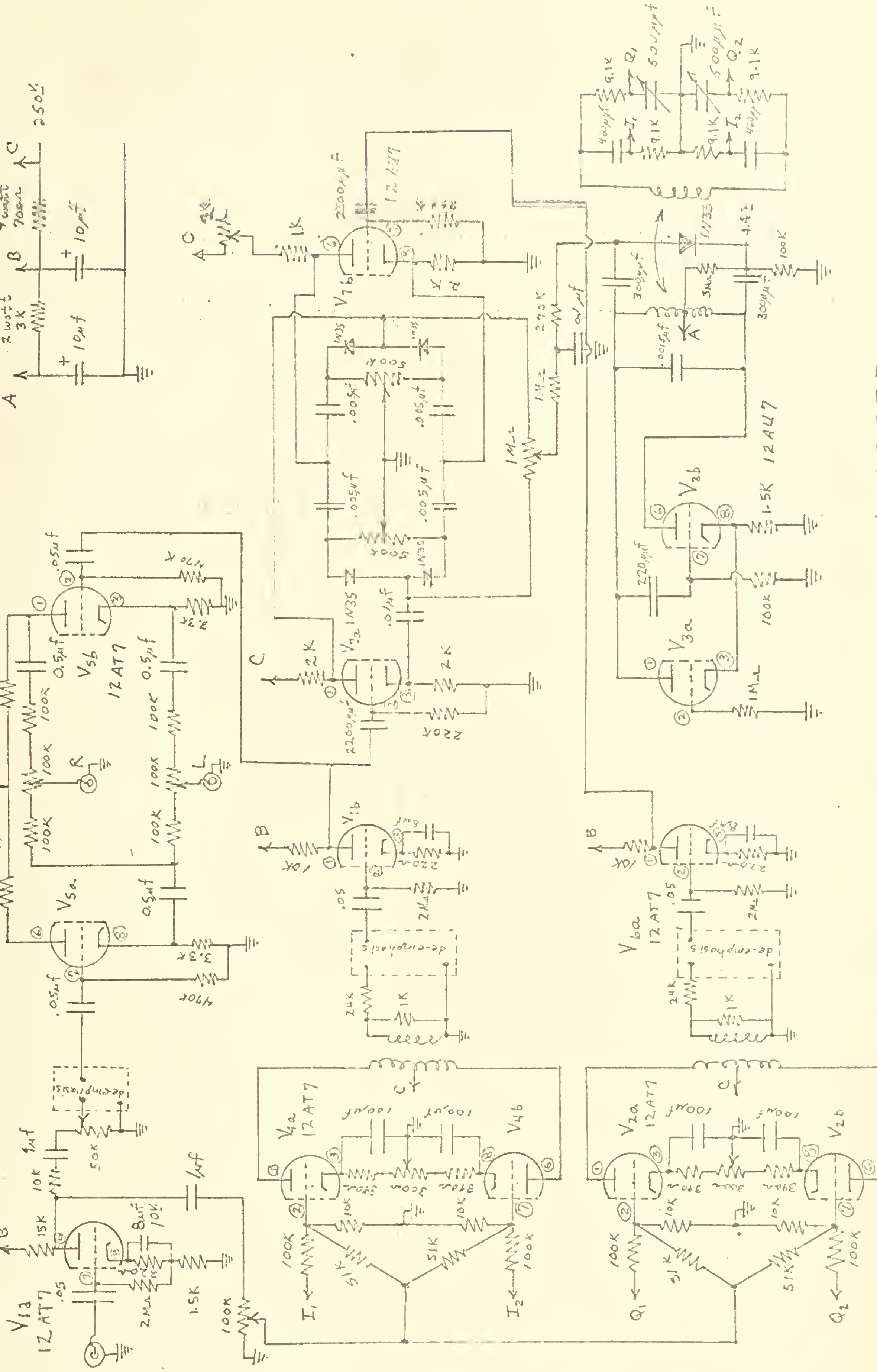
$$(45)\quad e_{p} = A' \sin 2\Delta$$

This equation represents the phase control voltage obtained from the phase detector. A qualitative analysis of the operation of this circuit has been shown by Nupp /10/.

This direct current voltage is then used to vary the reactance of the tank circuit of the oscillator and thereby adjust the frequency or phase of the oscillator to bring it back into synchronization.

Figure 17 is the schematic diagram of the multiplex adapter as modified for synchronous detection. It is noted that the 19,000 cycle pilot subcarrier amplifier and associated rectifier have been replaced by an additional Balanced Square Law Product Detector, de-emphasis network, difference channel amplifier, and two 45 degrees phase shifting networks.

Operation of this circuit is as just described except that instead of using one 90 degrees phase shifting network for the Q channel, we shift the Q channel 38,000 cycle oscillator frequency by 45 degrees and also the I channel 38,000 cycle oscillator frequency by the same



SYNCHRONOUS DETECTOR ADAPTER

FIGURE 17

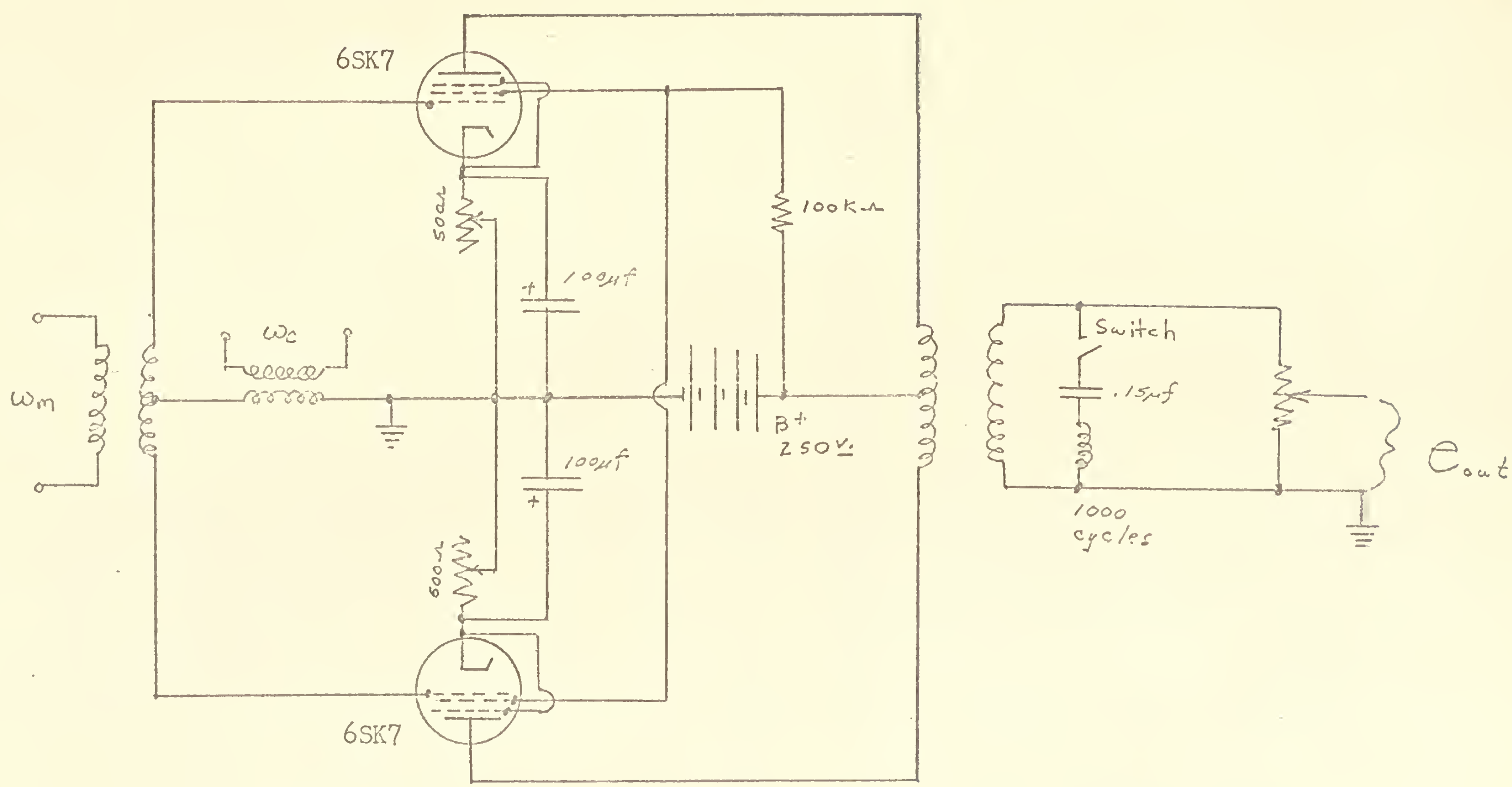
amount only in the opposite direction to make a net difference in phase between the two channels of 90 degrees. Of course, the net effect will still be the same.

In order to keep the number of changes in the circuit to a minimum, it was decided to utilize the existing oscillator. This meant trying to vary the reactance of a floating oscillator tank circuit (floating at plus 180 volts). To do this a 1N35 type diode was used as a varicap and reversed biased at minus four volts. As the direct current voltage from the phase detector varies, this will in turn vary the reverse bias on the diode which will therefore change its effective capacitance.

OPERATION OF SYNCHRONOUS DETECTOR

Almost all the problems encountered by Feit /11/ in developing his hybrid synchronous detector prototype were encountered by the author. As Feit stated, all components must be within one percent for near perfect balancing. Then physical balancing of the circuits becomes the remaining problem. Webb /12/, in discussing the design of a synchronous detector, suggests a few techniques to assist in alignment and balance of the circuits. However, utilizing all the available techniques (including a phase-meter to insure that the two channels were in phase quadrature), the adapter would not phase lock nor frequency lock-in the oscillator when receiving the FM stereophonic signals.

To check the operation of the synchronous detector multiplex adapter, a balanced modulator was built that was capable of producing a Double Sideband Surpressed Carrier signal centered at 38,000 cycles. (See



Balanced Modulator

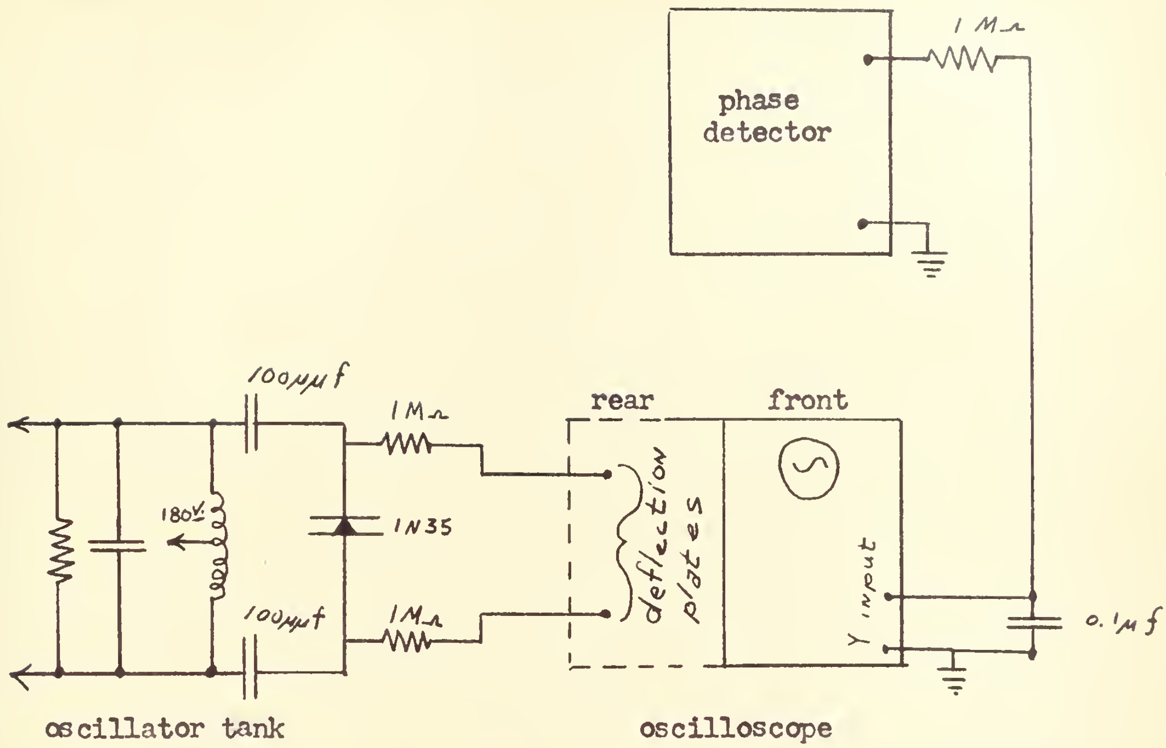
FIGURE 18

figure 18 for a schematic diagram of the balanced modulator.) Such a signal would be identical to the difference channel of a multiplex signal where the Left (or Right) channel output was equal to zero.

Using this procedure it became much easier to balance the channels, insure 90 degrees phase difference, and adjust the oscillator frequency. By careful tuning of the oscillator it was found that the synchronous detection loop would lock-in for any and all modulating frequencies from 50 to 15,000 cycles when the oscillator was adjusted within two cycles. Once locked-in, it would stay locked-in for extended periods unless the 38,000 cycle signal generator for the balanced modulator drifted by more than two cycles. At no time did it fall out of synchronization due to drift of the 38,000 cycle oscillator in the multiplex adapter.

By making some D.C. voltage measurements (using a D.C. transistor power supply at the output of the phase detector) versus frequency change (using a frequency meter that read to the nearest cycle), it was found that it took plus or minus one volt D.C. to decrease or increase the oscillator frequency by one cycle, respectively; two volts to change it by two cycles; three volts to change it by three cycles; etc. Since the D.C. voltage output of the phase detector was of the order of 0.1 volts, it was obvious that this was not sufficient D.C. voltage to make a sufficient change in the capacitance of the varicap.

To verify that lack of sufficient voltage was the reason the synchronous detector would not phase lock-in for an FM multiplex signal, it was decided to use the D.C. amplifiers of a Dumont 304-H oscilloscope to provide sufficient D.C. voltage so as to provide sufficient change in capacitance of the varicap to lock-in the oscillator. See



Scheme for Using D.C. Amplifiers of Dumont 304-H

FIGURE 19

figure 19 above for the installation.

Using this method it was found that the synchronous detection loop would lock-in the 38,000 cycle oscillator, both in frequency and phase. Several values of capacitance were tried to achieve the best charge and discharge constant. However, a value could not be found that would provide adequate lock-in for all signal levels. It was found that once the signal would rise to a comfortable level, the oscillator would lock-in, however, after a pause in program material of over a second, the oscillator would fall out and it would take a quite audible signal to bring it back in. This was particularily annoying in listening to certain types of classical music in which

there is an extended build-up of level since the synchronous detector would not lock in for several seconds. However, once it was locked-in, the fidelity appeared to be of a higher quality than did the fidelity of the other system.

CHAPTER VII

CONCLUSIONS

The results of the adapter test which were discussed in Chapter V substantiate the theoretical development given in Chapter IV. Many of the problems encountered in obtaining a useable multiplex signal were also discussed. The proposed multiplex adapter which is described in this paper was based upon demodulation of the stereophonic "difference" channel signal without the use of any filters since, as shown, utilization of filters would result in an appreciable loss of separation.

Although separation achieved was not as great as expected, it could be attributed to the non-linearities of the push-pull transformer. Should a model using the features incorporated in this adapter be made commercially, it would certainly be possible to obtain a transformer that would have adequate phase and gain characteristics from 50 to 15,000 cycles to keep the phase variation within three degrees and the gain variation within 0.25 decibels, or, as a further possibility, resistive loading might be used for the plate loads of the Balanced Square Law Product Detector and then the two output signals applied to the two inputs of a differential amplifier. This would require one additional tube (a double triode, 12AU7, for example) but may result in a lower overall cost of the unit.

This adapter unit did have the problem common to all adapters—"jitter" of the 19,000 cycle pilot subcarrier. An effort was made to eliminate this problem using synchronous detection techniques in lieu of phase locking with the 19,000 cycle pilot subcarrier. Although

phase locking was achieved using this method, it was not satisfactory in that it would not lock-in for signals below a certain level—this level being well within the hearing range. Possibly, further experimentation in this area, particularily with a different type of oscillator and with a reactance tube, might result in a satisfactory system. Utilizing special arrangements of capacitors and resistors it would be possible to design the multiplex system with a very rapid charge time (say 0.25 seconds for a signal level that is just at the threshold of normal hearing) and a very, very slow discharge time. This should be quite possible since there is only one frequency that the system would always be operating at—38,000 cycles.

Should such a design prove workable, it is expected that it would be superior to any multiplex adapter presently of the market. Cost-wise, it should be within the price range of the better priced multiplex adapters presently on the market. This would be in the range of $80.00 to $120.00

The operation of the Balanced Square Law Product Detector is of particular attention worth noting. Although hampered by the lack of an adequate push-pull transformer, it did operate exceptionally well. It is quite possible that other applications of this device might be feasible—particularly in the low frequency range.

It is suggested that any future work in this area include the following problems:

1). Operation of the multiplex adapter with an "adequate" push-pull transformer.

2). Operation of the multiplex adapter with a differential amp-

lifier in lieu of the push-pull transformer.

3). Operation of the synchronous detection multiplex adapter with a more practical oscillator and reactance tube.

4). Further applications of the Balanced Square Law Product Detector.

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APPENDIX A

LISTING OF EIGHT DIFFERENT SYSTEMS PROPOSED TO FCC

Crosby-Teletronics Corporation

The Crosby System produces low distortion for frequencies below 7,500 cycles except when the transmitted Left and Right signals are equal and of opposite polarity. Then, a much higher distortion is produced. To lessen this effect, a filter which is "flat" from 25,000 to 75,000 cycles and attenuates all frequencies below 25,000 cycles is required. Such a filter has not yet been designed and built, nor is it expected that it could be built at a moderate cost. The System has a greater loss in signal to noise ratio for monophonic reception and a lesser loss for stereo. It cannot utilize SCA multiplexing transmission.

Calbest Electronics

In the Calbest System the subcarrier is limited to 7,000 cycles and hence there is no separation above this frequency. It also produces high stereo subchannel noise characteristics and excessive cross-talk.

Multiplex Development Corporation

The Multiplex Development System has an 8,000 cycle limitation of subcarrier and hence no stereo separation above this frequency. It produces high stereo subchannel noise characteristics and excessive cross-talk. It can provide a channel separation without a de-matrix network at the receiver.

Electric and Musical Industries Ltd.

Theoretically, the Electric and Musical Industries System is superior to all other systems in most respects. However, in actual operation its capability for producing a subjective stereophonic effect is handicapped by orchestral dynamics in that separation of Left and Right channels is not accurately preserved for reproduction in the respective loud speakers. On sustained tones, for example, the output for the stereophonic receiver becomes monophonic, while for any material consisting of a number of sound sources, such as an orchestra would produce, a rapid undesirable shifting of gain between the two output channels will be produced.

Zenith Radio Corporation

The Zenith System produces low values of distortion (i.e. harmonic distortion, cross-talk, and intermodulation products due to system non-linearity either in transmission or reception) under all test conditions for frequencies below 7,500 cycles. It has a smaller loss in signal to noise ratio for monophonic reception but a greater loss for stereophonic reception. It is compatible with most present day tuners or receivers if the IF stages are re-aligned or modified for a broader, less "peaked" frequency response. This of course will reduce selectivity, sensitivity, and signal to noise performance of the receiver. It is also campatible with the previously authorized SCA transmission.

General Electric Company

The General Electric System, except for a few parameter differences ,

is identical to the Zenith System. It therefore has the same characteristices as the Zenith System.

General Electric's Alternate Proposal

There is no data available on this System since it was withdrawn by General Electric prior to test by the FCC.

Philco Corporation

There is no data available on this System since it was withdrawn by Philco Corporation prior to test by FCC.

APPENDIX B

ANALYSIS OF THE GENERAL ELECTRIC BANDPASS FILTER

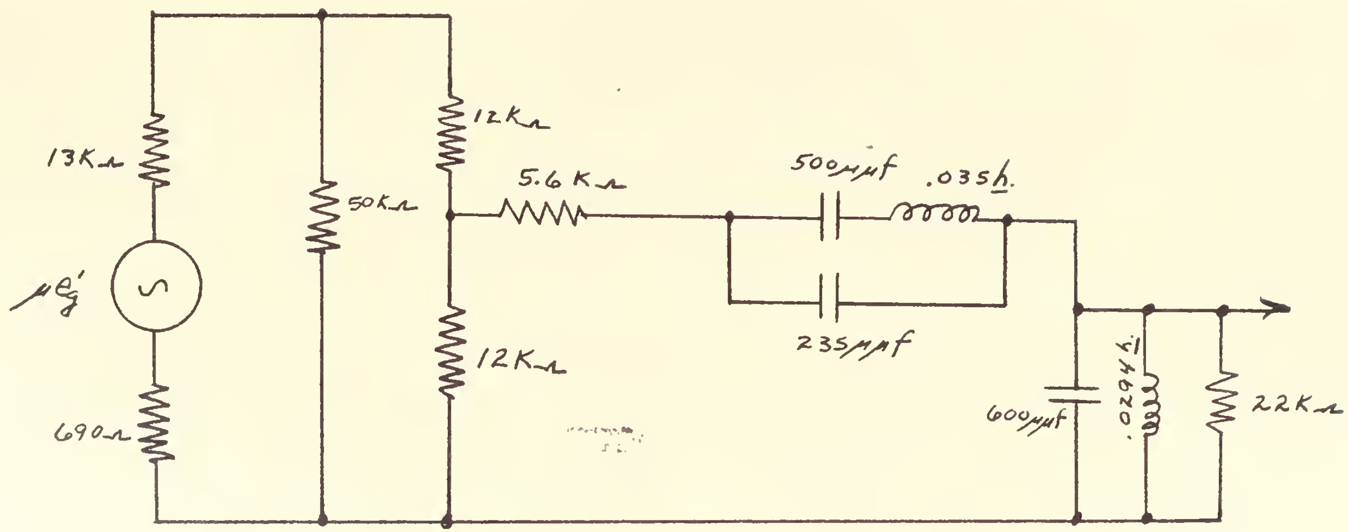
Figure B-1 is an equivalent circuit of the bandpass filter in the General Electric adapter shown in figure 4 of Chapter III. A mathematical analysis of this circuit has been made using "loop currents" for numerous frequencies between 10,000 to 100,000 cycles. Figure B-2 is a plot of the attenuation over this range. As can be seen, over the range of 23,000 to 53,000 cycles there is a variation of plus or minus 1.92 decibels. Assuming for the moment that the filter is phase linear, it can be seen from figure 5 that the separation over the full frequency spectrum will be about 13 decibels.

Figure B-3 is a plot of the phase response from 10,000 to 100,000 cycles. Figure B-4 is a plot of the same data only it is plotted on linear paper and is plotted versus the demodulated frequency. On this plot a phase linear filter should appear as a straight line. As can be see from the plot, each sideband is "phase linear" within three degrees for the entire 50 to 15,000 cycle frequency spectrum. However, there is a difference in time delay between the upper and lower sideband. It is believed that this difference in time delay is not compensated for anywhere in the circuit so that there will be some small degradation of separation due to it. However, this difference of time delay could be very easily compensated for in the matrix.

It can therefore be stated that over the full frequency spectrum of 50 to 15,000 cycles there is less than 13 decibels separation. However, using the range of 50 to 10,000 cycles there is approximately 20 decibels separation, while for the range of 50 to 8,000 cycles there

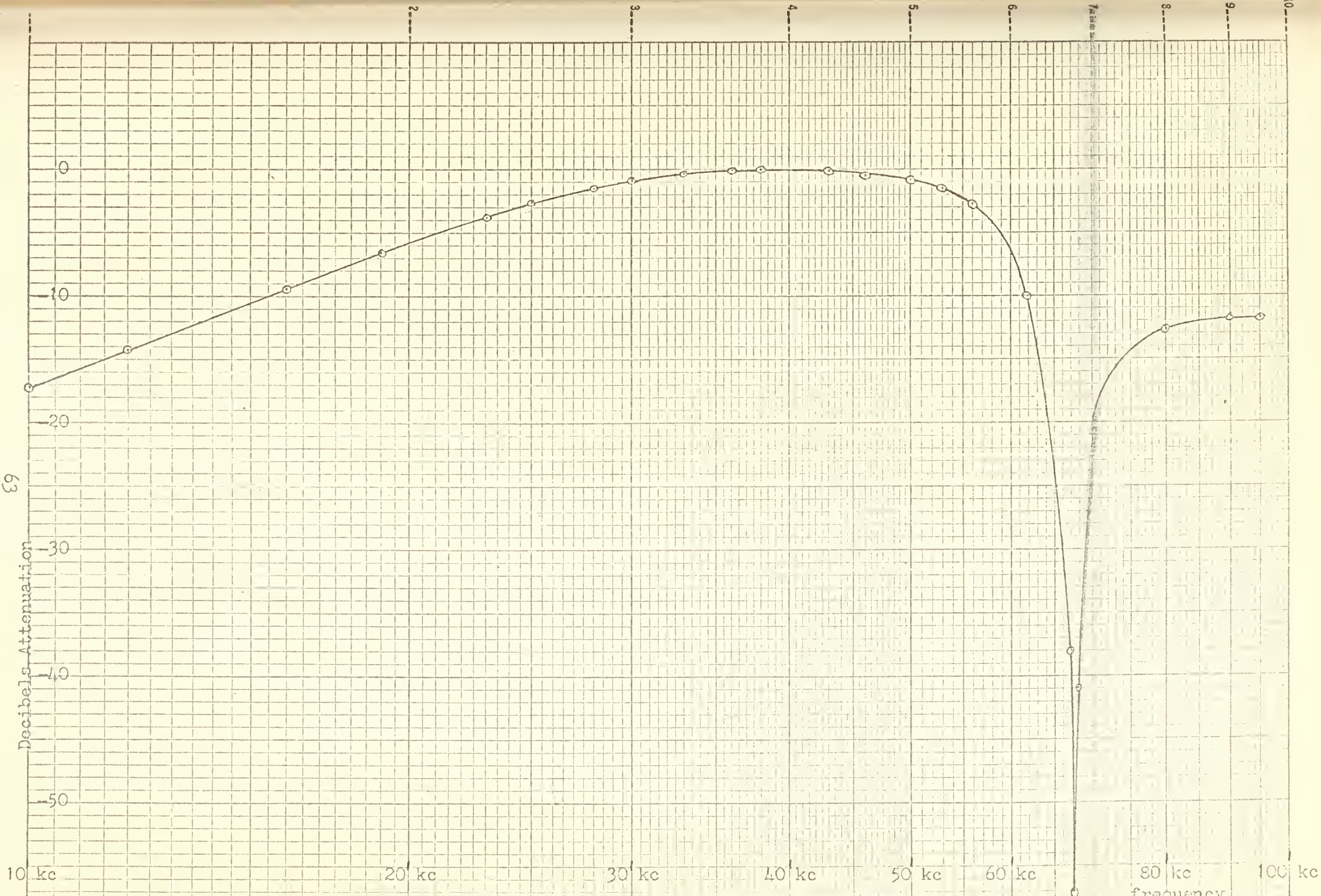
is over 25 decibels separation.

This is actually fairly "good" since the majority of the audio power is in the range of 50 to 8,000 cycles where there is very "good" separation. It would only be the small higher frequency components that would lose separation.



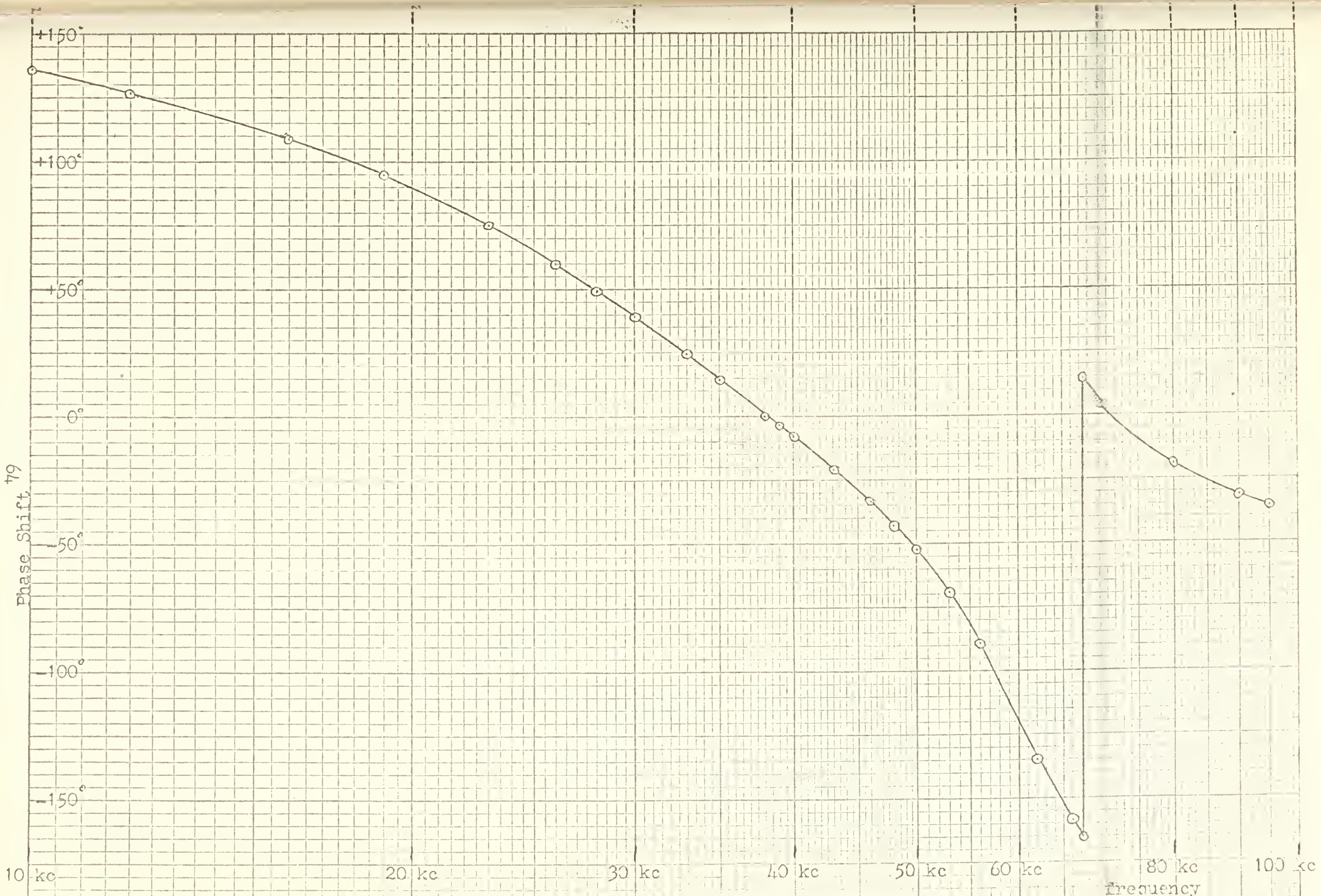
GE Bandpass Filter Equivalent Circuit

FIGURE B-1



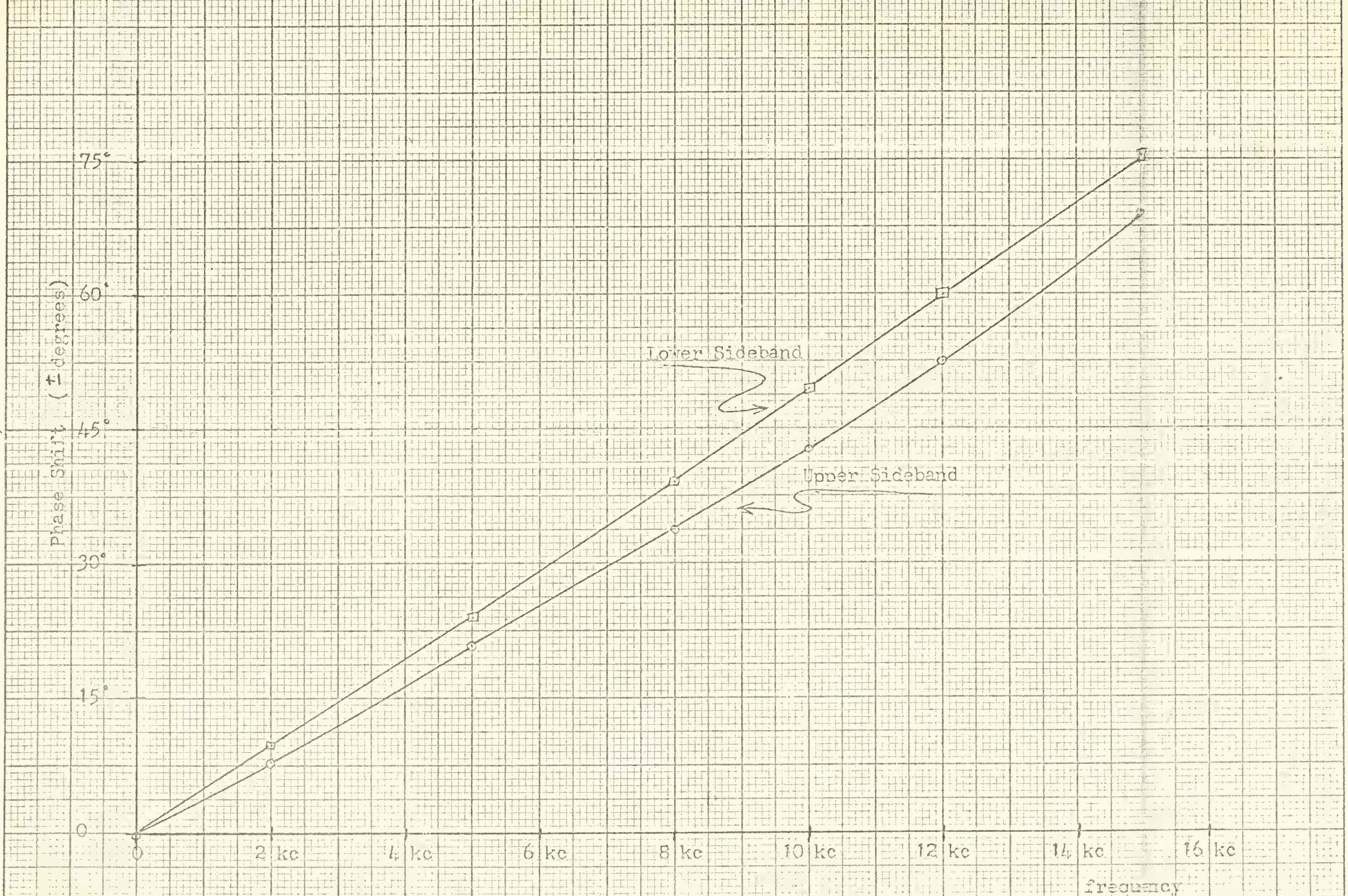
Attenuation Due to Bandpass Filter in GE Adapter

FIGURE B-2



Phase Shift Due to Bandpass Filter in GE Adapter

FIGURE B-3



Phase Shift Due to Bandpass Filter in GE Adapter

FIGURE B-4

APPENDIX C

ANALYSIS OF CUBIC TERM IN BALANCED SQUARE LAW PRODUCT DETECTOR

As shown in Chapter IV, equations (18) through (41), if the linear and squared terms in a Balanced Square Law Product Detector are considered, the output after filtering (assuming from henceforth that $\triangle$ equals zero) is:

(C-1) $$e_{out} = 4A_2 E_c E_s \cos(\omega_{L-R} t)$$

However, higher order terms do exist. Looking at the next higher order term from equation (19), we have for the total signal on the grid of tube V1 (see figure 12);

(C-2) $$e_{g1} = E_c \sin(\omega_c t) + E_a \cos(\omega_{L+R} t) + E_s\left[\sin(\omega_c t + \omega_{L-R} t) + \sin(\omega_c t - \omega_{L-R} t)\right]$$

and on the grid of tube V2:

(C-3) $$e_{g2} = -E_c \sin(\omega_c t) + E_a \cos(\omega_{L+R} t) + E_s\left[\sin(\omega_c t + \omega_{L-R} t) + \sin(\omega_c t - \omega_{L-R} t)\right]$$

The cubic output will be:

(C-4) $$e_{out_3} = A_3\, e_{g1}^3 - A_3\, e_{g2}^3$$

assuming that the gain of both tubes are identical.

The voltage developed at tube V1 due to the third harmonic will therefore be:

$$(C\text{-}5)\quad e_{g_1}^{3} = E_s^3\left[\sin(\omega_c t + \omega_{L-R} t) + \sin(\omega_c t - \omega_{L-R} t)\right]^3$$

$$+ 3E_s^2 E_a \cos(\omega_{L+R} t)\left[\sin(\omega_c t + \omega_{L-R} t) + \sin(\omega_c t - \omega_{L-R} t)\right]^2$$

$$+ 3E_s^2 E_c \sin(\omega_c t)\left[\sin(\omega_c t + \omega_{L-R} t) + \sin(\omega_c t - \omega_{L-R} t)\right]^2$$

$$+ 3E_s E_a^2 \cos^2(\omega_{L+R} t)\left[\sin(\omega_c t + \omega_{L-R} t) + \sin(\omega_c t - \omega_{L-R} t)\right]$$

$$+ 6E_s E_a E_c \sin(\omega_c t)\cos(\omega_{L+R} t)\left[\sin(\omega_c t + \omega_{L-R} t) + \sin(\omega_c t - \omega_{L-R} t)\right]$$

$$+ 3E_s E_c^2 \sin^2(\omega_c t)\left[\sin(\omega_c t + \omega_{L-R} t) + \sin(\omega_c t - \omega_{L-R} t)\right]$$

$$+ E_c^3 \sin^3(\omega_c t) + 3E_c^2 E_a \sin^2(\omega_c t)\cos(\omega_{L+R} t)$$

$$+ 3E_c E_a^2 \sin(\omega_c t)\cos^2(\omega_{L+R} t) + E_a^3 \cos^3(\omega_{L+R} t)$$

The voltage developed at tube V2 will be identical except that the sign of each coefficient containing an odd power of E_c will be negative.

Subtracting, and solving for e_{out_3} we have:

$$(C\text{-}6)\quad e_{out_3} = A_3\Big\{6E_s^2 E_c \sin(\omega_c t)\left[\sin(\omega_c t + \omega_{L-R} t) + \sin(\omega_c t - \omega_{L-R} t)\right]^2$$

$$+ 12E_s E_a E_c \sin(\omega_c t)\cos(\omega_{L+R} t)\left[\sin(\omega_c t + \omega_{L-R} t) + \sin(\omega_c t - \omega_{L-R} t)\right]$$

$$+ 2E_c^3 \sin^3(\omega_c t) + 6E_c E_a^2 \sin(\omega_c t)\cos^2(\omega_{L+R} t)\Big\}$$

Expanding this using simple trignometric idenities, we have:

$$(C\text{-}7)\quad e_{out_3} = A_3\Big\{\tfrac{3}{2}E_s^2 E_c\left[6\sin(\omega_c t) - \sin(3\omega_c t + 2\omega_{L-R} t)\right.$$

$$- \sin(3\omega_c t - 2\omega_{L-R} t) + 3\sin(\omega_c t + 2\omega_{L-R} t) - 2\sin(3\omega_c t)$$

$$\left. + 3\sin(\omega_c t - 2\omega_{L-R} t)\right] + 3E_s E_a E_c\left[2\cos(\omega_{L+R} t + \omega_{L-R} t) + \right.$$

$$+ 2\cos(\omega_{L+R}t - \omega_{L-R}t) - \cos(2\omega_c t + \omega_{L-R}t + \omega_{L+R}t)$$

$$- \cos(2\omega_c t + \omega_{L-R}t - \omega_{L+R}t) - \cos(2\omega_c t - \omega_{L-R}t + \omega_{L+R}t)$$

$$- \cos(2\omega_c t - \omega_{L-R}t - \omega_{L+R}t)\Big] + \frac{E_c^3}{2}\Big[3\sin(\omega_c t) - \sin(3\omega_c t)\Big]$$

$$+ \frac{3}{2}E_c E_a^2\Big[2\sin(\omega_c t) + \sin(\omega_c t + 2\omega_{L+R}t) + \sin(\omega_c t - 2\omega_{L+R}t)\Big]\Big\}$$

Analyzing this frequency spectrum, we can see that each component can possess the following bands of frequencies (assuming that the 19,000 cycle pilot subcarrier is the highest frequency of $e_{s_{L+R}}$):

(C-8)	$\omega_{L+R} - \omega_{L-R}$	0-19,000 cycles
(C-9)	$\omega_{L+R} + \omega_{L-R}$	0-34,000
(C-10)	$\omega_c - 2\omega_{L+R}$	8,000-38,000
(C-11)	$\omega_c - 2\omega_{L-R}$	8,000-38,000
(C-12)	ω_c	38,000
(C-13)	$\omega_c + 2\omega_{L-R}$	38,000-68,000
(C-14)	$2\omega_c - \omega_{L-R} - \omega_{L+R}$	32,000-76,000
(C-15)	$\omega_c + 2\omega_{L+R}$	38,000-76,000
(C-16)	$2\omega_c + \omega_{L-R} - \omega_{L+R}$	57,000-91,000
(C-17)	$2\omega_c - \omega_{L-R} + \omega_{L+R}$	61,000-95,000
(C-18)	$2\omega_c + \omega_{L-R} + \omega_{L+R}$	76,000-110,000
(C-19)	$3\omega_c - 2\omega_{L-R}$	84,000-114,000

(C-20) $3\omega_a$ 114,000 cycles

(C-21) $3\omega_c + 2\omega_{L-R}$ 114,000-144,000

By examining each frequency component in equation (C-7), it can be seen that the only output of the de-emphasis network due to the third harmonic (i.e. the 0 to 15,000 cycle frequency components) will therefore be:

$$\text{(C-22)}\quad e_{out_{3f}} = A_3 \Big\{ \tfrac{9}{2} E_s^2 E_c \sin(\omega_c t - 2\omega_{L-R} t) + 6 E_s E_a E_c \left[\cos(\omega_{L+R} t + \omega_{L-R} t) + \cos(\omega_{L+R} t - \omega_{L-R} t)\right] + \tfrac{3}{2} E_c E_a^2 \sin(\omega_c t - 2\omega_{L+R} t) \Big\}$$

Since these are all relatively large components (neglecting A_3 for the moment), it can be seen that to keep intermodulation distortion due to the third harmonic to a minimum the two tubes must be biased and loaded to minimize A_3 . To further minimize this distortion, E_c should be made much greater than E_s or E_a , say eight or ten times greater. For example, if E_s and E_a are each 0.2 volts, and E_c is 2.0 volts, we have for the cubic term:

$$\text{(C-23)}\quad e_{out_{3f}} = A_3 \left[0.36 \sin(\omega_c t - 2\omega_{L-R} t) + 0.48 \cos(\omega_{L+R} t + \omega_{L-R} t) + 0.48 \cos(\omega_{L+R} t - \omega_{L-R} t) + 0.12 \sin(\omega_c t - 2\omega_{L+R} t)\right]$$

while for the desired square term we have:

$$\text{(C-24)}\quad e_{out_{2f}} = 1.6 A_2 \cos(\omega_{L-R} t)$$

Now, assuming that the two tubes are operating such that A_2 is

twenty times A_3 (say A_2 is 1.0 and A_3 is 0.05) we therefore have for the cubic term:

$$(C\text{-}25)\quad e_{out_{3f}} = 0.018 \sin(\omega_c t - 2\omega_{L-R} t) + 0.024 \cos(\omega_{L+R} t + \omega_{L-R} t)$$
$$+ 0.024 \cos(\omega_{L+R} t - \omega_{L-R} t) + 0.006 \sin(\omega_c t - 2\omega_{L+R} t)$$

and for the square term:

$$(C\text{-}26)\quad e_{out_{2f}} = 3R \cos(\omega_{L-R} t)$$

Using these criteria, it can be seen that the cubic term will produce no significant distortion.

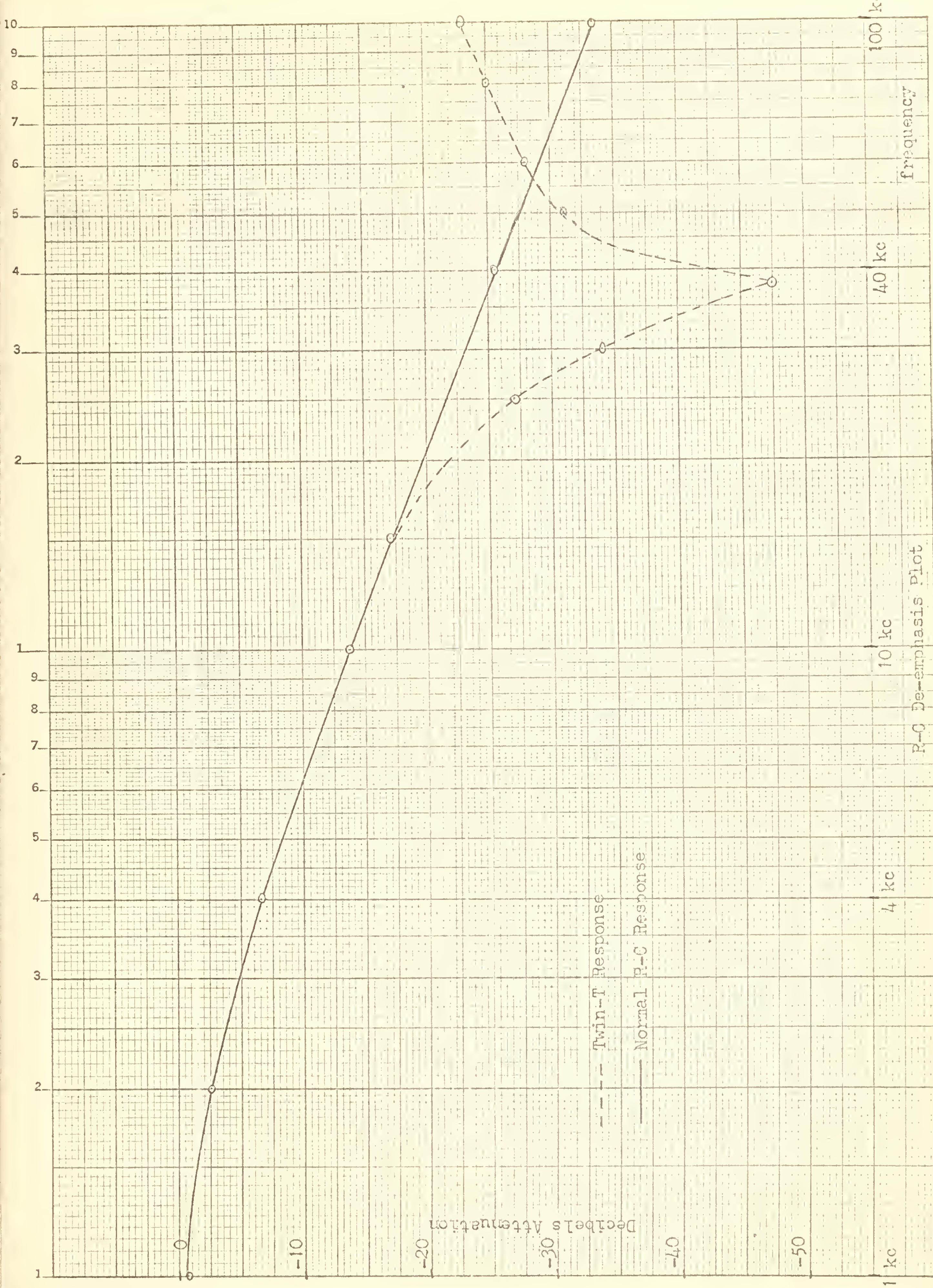


FIGURE D-1
R-C De-emphasis Plot