THE CAPTURE OF THE WEAKER OF TWO CO-CHANNEL FREQUENCY-MODULATED SIGNALS

by

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ABSTRACT

An experimental study is made of a trapping technique for receiving the weaker of two co-channel FM signals. This method utilizes the basic static trap concept in conjunction with weaker-signal suppression devices (such as narrow-band limiters, feedforward circuits, etc.) in such a manner that predetection of the stronger signal is not required. The experimental receiver is able to capture the weaker signal with less than 20% distortion when the input ratio of weaker-to-stronger signal amplitudes is in the range $0.05 \leq \underline{a} \leq 0.9$.

The dependence of the performance of the system upon the degree of weaker-signal suppression is studied in order to ascertain the minimum degree of weaker-signal suppression required. The effect of various modulation conditions of interference and desired signal upon the system performance is explored in detail.

Among others, problems relating to the improvement of the fixed-trap circuit are also explored. It is found through an analysis of the instantaneous frequency response of filters that the delay accumulated by the stronger signal message in the suppressor and converter channel filters must be compensated for best fixed-trap system performance in suppressing the effects of the stronger unwanted signal.

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Chapter I

INTRODUCTION

Communication systems employing frequency-modulated signals are generally considered to be low-distortion systems. This property is manifested by the intrinsic ability of FM receivers to cpature only the desired signal provided that the amplitude of this signal exceeds the amplitude of any undesirable interfering signals by a sufficient margin. Often, the relative amplitude of the undesired interfering signal is low enough so that its disturbances do not disrupt the operation of the basic circuits and the frequency disturbances created by it are either low-level or are supersonic and hence filterable by the low frequency circuits in the receiver output. This is the main reason why FM systems perform well against interferences from signals that are weaker than the desired one.

Now suppose the amplitude of the undesired interfering signal exceeded this margin of safety or in fact even was well above the amplitude of the desired signal. In the first instance, the receiver output would deliver a badly distorted, almost unrecognizable desired signal message. In the second instance, where the amplitude of the interference exceeded the amplitude of the desired signal, the receiver would capture the interference and irretrievably suppress the desired signal. Such situations can and do occur quite frequently in FM communication systems. A typical example would be a commercial broadcast situation where a receiver is located in a poor environment, such as a fringe area, for reception of a particular distant station. Other stations or poor propagation characteristics may prevent reception of the desired signal. Another example is a situation in which a particular FM channel is intentionally jammed by a strong signal that could be either CW or frequency modulated.

Thus, although advantageous in some cases, the stronger signal capture effect is also a handicap and imposes a severe limitation on the conventional FM receiver which until quite recently could not be offset by simple modifications in receiver design.

The present thesis is concerned with an experimental study of the possibilities and limitations of a scheme which, by virtue of a few simple modifications in the IF stages immediately preceding the demodulator, is capable of capturing the weaker signal when a stronger CW or frequency modulated signal occupies the same pass band or "channel".

Historical Background

In retrospect, it is surprising to find that prior to 1955 so few attempts have been made to devise FM receivers capable of capturing the weaker of two co-channel interfering signals. To the author's knowledge the earliest reported attempt to capture the weaker signal was made by Wilmotte in 1954. (10) He suggested the following scheme for capturing the weaker signal. The stronger signal, at frequency p, is captured by a conventional FM receiver and a voltage is thus obtained which is directly proportional to p. Then if the weaker signal is at frequency $p + r$, the frequency difference r can be isolated from the amplitude variations of the resultant signal, and detected to yield a voltage proportional to r. The weaker signal is then obtained simply by adding the voltages from the two detectors giving a voltage which is proportional to $p + r$.

There are two difficulties with this scheme. In the first place, a single tone at the frequency r becomes extremely difficult to isolate and convert into a voltage that is proportional to r, as r decreases from supersonic values to values that are well within the range of the message spectrum. Second, an extremely complex mechanism would be required to keep track of whether the weaker signal lies above or below the stronger signal on the frequency scale, and automatically apply the necessary polarity to the corrective voltage difference. Evidently the difference frequency alternates between positive and negative values, depending on whether the weaker signal has a higher or lower frequency than the stronger, and unless the receiver can tell what the sign of r is it will not be able to decide whether to add or subtract the two detected voltages.

The only other suggestions that the author is aware of were made by Baghdady (1) who proposed two methods, both of which perform a suppression of the amplitude of the stronger signal below the amplitude of the weaker signal in the IF stages of the receiver. The role of weaker and stronger is reversed by this stronger-signal suppression enabling a conventional demodulator to capture the augmented weaker signal. The two methods conceived by Baghdady - feedforward across a limiter and dynamic trapping - represent the simplest and most effective demonstrated techniques capable of extracting, with negligible distortion, the message from a much weaker co-channel FM signal.

Feedforward Across the Limiter

Fig. 1.1 (Fig. 1 of Ref. No. 1) illustrates the feedforward technique. This scheme was experimentally demonstrated by R. H. Small in 1958 under the supervision of Baghdady. (2)

The crux of operation for the feedforward system is the interference suppression properties of narrow band limiters. Baghdady has shown (6) that if an amplitude limiter is followed by a filter whose pass band extends over a well-defined bandwidth, the output spectrum will reveal a diminution in the influence of the weaker signal upon the overall resultant signal. The amount of weaker-signal suppression depends upon the amplitude ratio of the signals at the input to the limiter, the instantaneous frequency difference between the two co-channel signals, and the location of the instantaneous frequency of the stronger signal in the pass band of the IF filter preceding the limiter. The maximum weaker signal relative amplitude suppression than can be achieved by virtue of narrow band limiting is 6 db.

In the feedforward system two independent, unilateral channels are provided. The first channel is a linear, variable gain amplifier whose output is directly proportional to the input; hence, the relative strength of the weaker to stronger signal in the output of this amplifier is the same as that at the input. The second channel is a narrow band limiter that decreases the strength of the

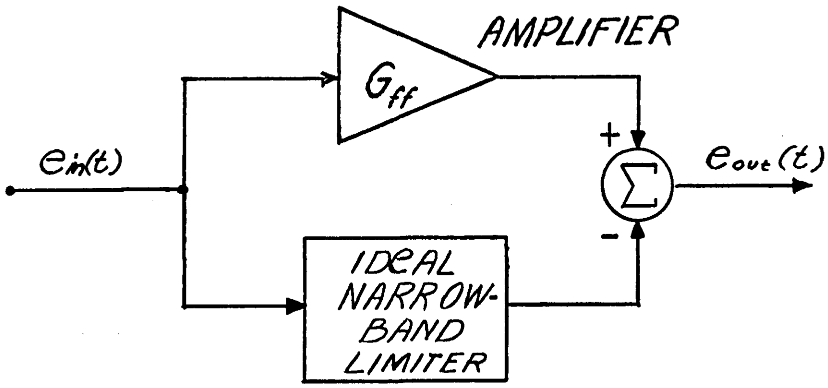


Fig. 1.1 Basic Feedforward Scheme

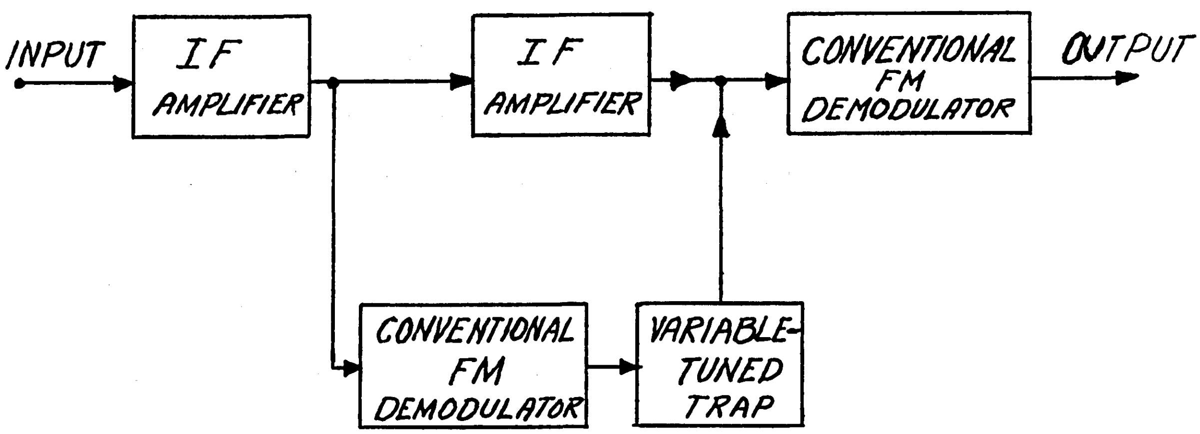


Fig. 1.2 Variable-Tuned Trap System

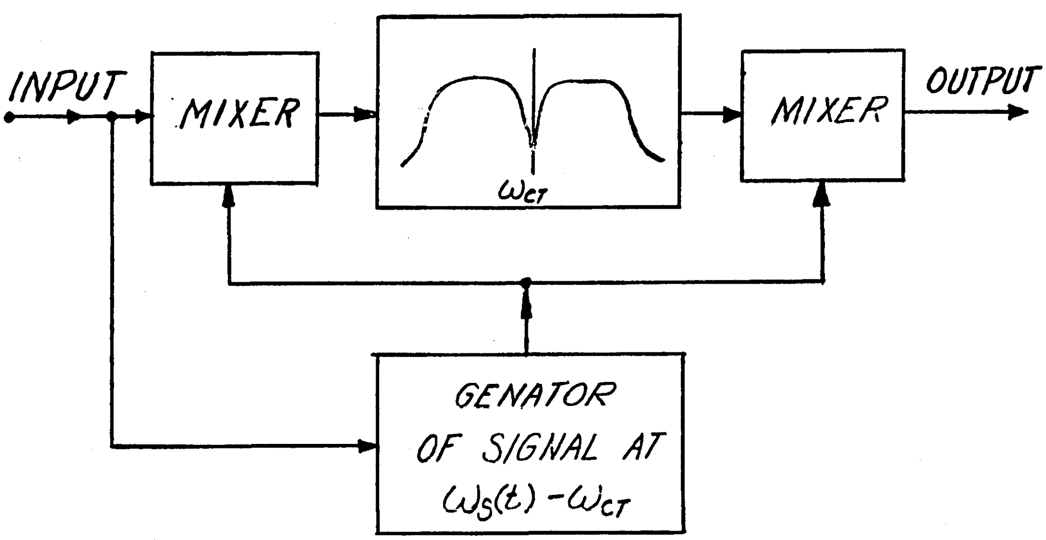


Fig. 1.3 General Form of Fixed-Trap System

weaker signal relative to the stronger signal. The two channels are provided with a phase difference of 180° and the outputs are added (which actually amounts to a subtraction because of the phase differential of 180°). If now the gain of the linear amplifier is adjusted so that the amplitudes of the weaker signal in each channel are equal, the subtraction of the outputs of the two channels will result in a relatively augmented stronger signal. The improvement in the ratio of weaker-to-stronger signal amplitude is greater than the achieved by the narrow band limiter alone.

If instead, the gain of the linear amplifier is adjusted so that the amplitudes of the stronger signal in each channel are equal, the subtraction results in the difference between the amplitudes of the weaker signals in each channel appearing in the output. In this manner, the weaker signal can be made to predominate in the output of the feedforward and a conventional FM demodulator can further suppress the effects of the originally stronger signal and deliver the message of the weaker signal.

Thus we see that by simply turning a knob on this feedforward device we can either improve the stronger-signal capture characteristics of a demodulator or enable this same demodulator to capture the weaker of the two co-channel FM signal. Some results of the weaker-signal capture ability of a system employing feedforward are given in Chapter VII.

Dynamic Trapping

In dynamic-trapping systems, a trap of narrow bandwidth is used to attenuate or cancel the undesired stronger signal. There are two trapping techniques illustrated in Figs. 1.2 and 1.3 (Figs. 12 and 13 of Ref. No. 1) that will effect this signal cancellation. The trapping system illustrated in Fig. 1.2 was successfully demonstrated by G. J. Rubissow in 1958 (3). This system utilizes a variable-tuned trap and requires knowledge of the instantaneous frequency behavior of the stronger signal. Thus, the frequency modulation of the stronger signal is first extracted by a stronger-signal-capture receiver. This modulation is then used to change the value of an electronically simulated reactance that forms a variable tuning element in the trap circuit. In this way, the attenuation band of the trap can be made to follow the instantaneous-frequency position of the stronger signal. The resulting attenuation should decrease the amplitude of the undesired signal by an amount that is sufficient to enable the initially weaker (desired) signal to emerge as the predominant one during most of its modulation cycle. The low-pass filter that is associated with the second FM demodulator will then deliver a voltage that varies directly with the instantaneous frequency of the desired signal, except when both signals fall simultaneously within the heavy attenuation band of the trap. A summary of the weaker signal capture of this system is given in Chapter VII.

This thesis is concerned with the realization of one form of the second dynamic-trapping technique of Fig. 1.3, (block diagram shown in Fig. 2.4) which does not require prior demodulation of the stronger signal.

GENERAL DISCUSSIONS OF STATIC TRAP TECHNIQUE

The technique used in this thesis to capture the weaker of two co-channel FM signals is based on an alternate form of the dynamic trapping technique. As originally conceived by Baghdady, (1) the variable tuned trap system (henceforth denoted as the dynamic trap system) and one possible form of the static-trap system require predetection of the instantaneous frequency behavior of the stronger signal. The manner in which this predetected stronger signal modulation is put to use is where the static-trap system differs from the dynamic-trap system. In the static-trap system, the predetected modulation is used to "freeze" the instantaneous frequency of the stronger signal so that it always equals the resonant frequency of a fixed tuned trap. A thesis presented in 1958 by F. I. Sheftman (4) utilized the static-trap technique for capturing the weaker of two co-channel FM signals.

Whereas the dynamic trap system cannot be implemented without predetection of the stronger signal modulation, the static-trap system can. (1) This thesis is devoted to the exploration of the static-trap system realization that does not require predetection of the stronger signal.

Before embarking on a discussion of the system actually used, a general discussion of the fixed trap system is necessary in order to provide a clear understanding of the approximations that will be made to avoid predetection of the stronger signal.

Fig. 1.3 (Fig. 13 of Ref. No. 1) illustrates the fixed trap weaker signal receiver as orignially conceived by Baghdady. The discussion of the system will be restricted to the case of independent FM signals whose carrier frequency separation is zero but whose RF amplitudes differ in magnitude. The stronger of the two incoming FM signals is assumed to be the interference.

Representing these signals in the familiar manner we have,

Strong Signal: $e_S(t) = E_S \operatorname{Cos}[\omega_1 t + \theta_S(t)]$ 2.1

Weak Signal: $e_W(t) = \underline{a} E_S \operatorname{Cos}[\omega_1 t + \theta_W(t)]$ 2.2

$\theta(t)$ represents the message information and is any arbitrary function of time-periodic, aperiodic or random.

A form of static-trap realization that requires predetection of the stronger signal is shown in Fig. 2.2 (Fig. 14 of Ref. No. 1). Let ρ_1 denote the capture ratio of the first FM demodulator. The output of this demodulator will be the message of the stronger signal $\theta_S(t)$ provided that the relative amplitude ratio $\underline{a}$ is less than ρ_1. If $\underline{a}$ lies in the range $\rho_1 \leq \underline{a} \leq 1$, the output of the first demodulator will still be $\theta_S(t)$ in addition to considerable distortion. For the moment assume that $\underline{a} < \rho_1$. The message $\theta_S(t)$ is now used to frequency modulate a variable frequency local oscillator whose carrier frequency we will denote as $(\omega_1 - \omega_0)$. The signal at this point is given by

$$e_c(t) = \operatorname{Cos}[(\omega_1 - \omega_0)t + \theta_S(t)] \qquad 2.3$$

Essentially what we have accomplished by this demodulation-modulation process is a linear translation of only the stronger

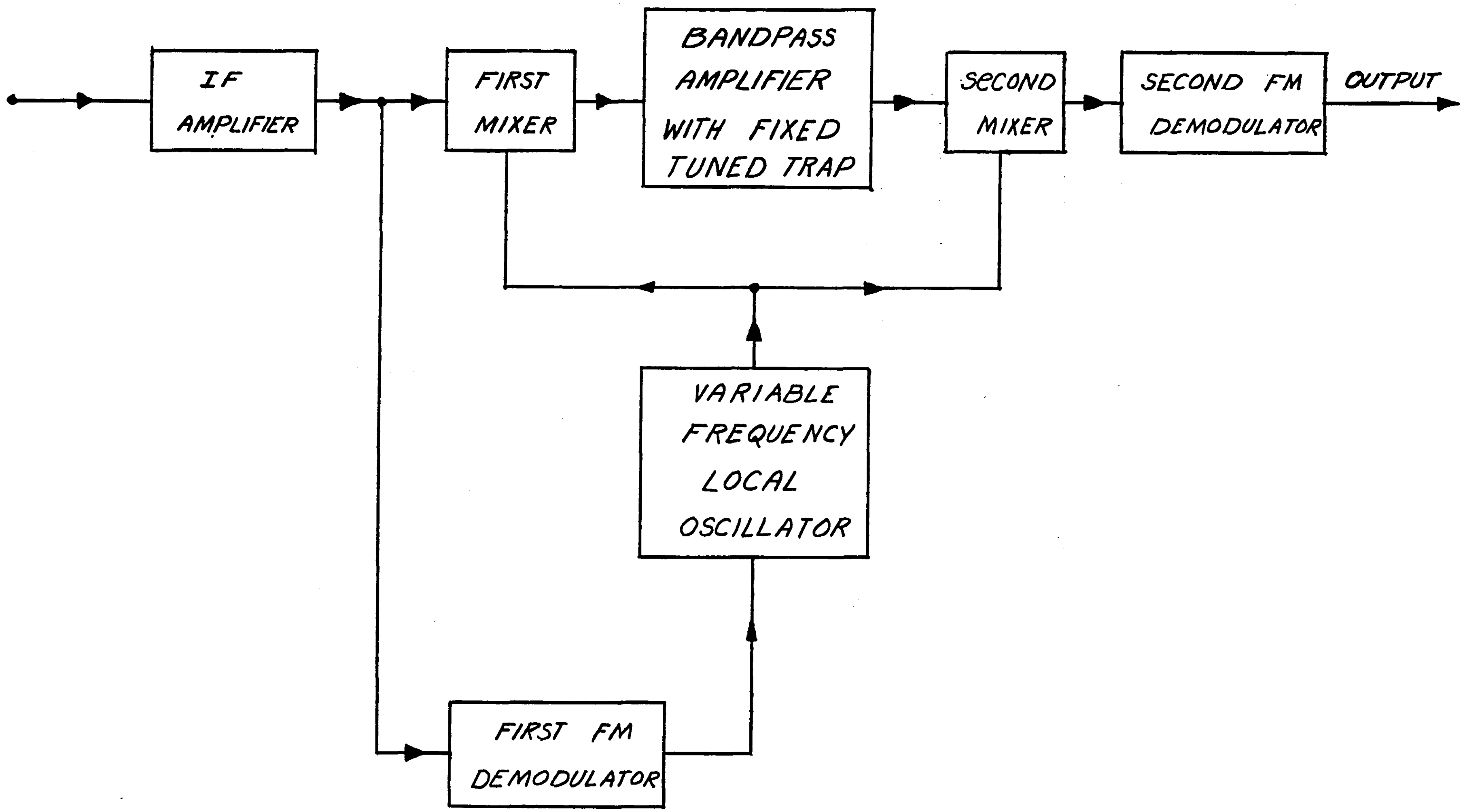


Fig 2.2 A fixed trap system that requires prior demodulation of the stronger signal.

signal to a new location (centered about $\omega_1 - \omega_0$) in the frequency spectrum. It is important to note two salient features of this operation. The first function of this operation was to completely suppress the effect of the desired weaker signal (provided $\underline{a} < \rho_1$). Secondly, a signal was derived whose instantaneous frequency differs from the instantaneous frequency of the incoming undesired stronger signal by a <u>constant</u> <u>difference</u> dependent <u>only</u> upon the center frequency of the local oscillator. (Note that if $\rho_1 \leq \underline{a} \leq 1$ we do not have this condition since $\theta_S(t)$ in Eq. 2.3 is distorted by capture noise and we have not preserved the stronger signal modulation.)

This regenerated translated stronger signal is mixed with the incoming weaker and stronger signals producing in the output of the first mixer (after filtering all harmonics and sidebands) the constant frequency sinusoid whose frequency coincides with the center frequency of the succeeding filters and a second signal that is frequency modulated about this new center frequency by the algebraic sum of the messages contained in both input signals. The center frequency of this signal is the constant frequency difference referred to in the previous paragraph. The output of the mixer is expressed by

$$e_{mix.\#1}(t) = \cos \omega_0 t + \underline{a} \cos[\omega_0 t + \theta_w(t) - \theta_s(t)] \qquad 2.4$$

This signal is fed into a bandpass amplifier with a fixed narrow tuned trap located in the center of the pass band at ω_0. For the time being, the effect of the trap on the frequency modulated component will not be considered. The effect of the trap will be focused on the carrier frequency component $\cos\omega_0 t$. Representing the trap attenuation factor by δ, we have in the output of this filter

$$e_{TRAP}(t) = \delta\cos\omega_0 t + \underline{a}\cos[\omega_0 t + \theta_w(t) - \theta_s(t)] \qquad 2.5$$

This signal is remixed with the output of the variable frequency local oscillator restoring the original frequency modulations of both signals.

$$e_{mix\#2}(t) = \delta\cos[\omega_1 t + \theta_s(t)] + \underline{a}\cos[\omega_1(t) + \theta_w(t)] \qquad 2.6$$

The output of the second mixer feeds the second demodulator but now the desired message will be carried by the predominant signal at the input to this demodulator. The ratio of desired to undesired signal at the input to the demodulator is now given by $\frac{\underline{a}}{\delta}$. If $\underline{a} > \delta$, the desired message will be faithfully captured by this second demodulator provided $\frac{\delta}{\underline{a}}$ does not exceed the capture ratio, ρ_2, of this second demodulator and provided that the desired signal strength exceeds the limiter threshold of this demodulator. Thus, the fixed trap receiver will deliver the desired message of the weaker signal with a minimum of distortion for all input values of the amplitude ratio $\underline{a}$ in the range $\frac{\delta}{\rho_2} \leq \underline{a} \leq \rho_1$.

The advantage of the static-trap system over the dynamic trap system is that the value of δ, the trap attenuation factor, need be controlled at one frequency only instead of over a wide range of frequencies as for the variable trap. Design of a fixed trap band-pass filter is much simpler. However, in the fixed trap system, that portion of the circuit more responsible for good performance is the first demodulator and variable frequency local oscillator. It is not an easy task to faithfully produce the translated stronger signal as required. Any errors in the reproduced signal will greatly influence the overall response. The reason for this lies in the fact that if we do have an error in reproducing this signal, the error will be manifested by a frequency modulated signal (whose deviation depends upon the magnitude of this error) appearing at the center frequency of the trap in place of the required CW signal. Thus, the signal we wish to attenuate will drift in and out of the trap and the response of the system will be degraded or the system may fail completely. However, this is mostly a practical problem and careful design could alleviate most of the trouble.

In the general discussion of the fixed trap system, trap distortion of any sort was ignored. Distortion in the desired message due to the trap can arise from two different considerations. The first concerns itself with the ability of the trap to follow the instantaneous frequency variations of the FM signal through quasi-stationary states. (Note that the trap will be swept in accordane with a rate given by the algebraic sum of the desired and undesired messages ($\theta_w(t)-\theta_s(t)$). If the trap filter is too sluggish with respect to this FM excitation, FM transient distortion will be introduced into the modulation of the FM signal. This is a limitation of both dynamic trapping techniques and transient distortion of this form will have to be tolerated. When the emphasis of the response is on intelligibility rather than fidelity, the transient distortion is tolerable. In the event this transient distortion is to be minimized, widening of the trap bandwidth is necessitated. However, widening of the trap BW may at first seem detrimental to the capture of the desired signal. Fortunately, as Rubissow has shown through his investigation of the dynamic trap, there generally is no need in having an exceptionally narrow trap filter. He has shown that the trap bandwidth mostly influences the abruptness with which the capture transition from one signal to another is achieved when the value of the amplitude ratio lies in the neighborhood of $\underline{a}=\delta$. The abruptness of this capture transition region is also affected by the capture ratio of the second demodulator.

In order to understand the effect of the trap attenuation characteristic on the capture transition, it is necessary to examine the instantaneous signal behavior of the two signals in the pass band of the bandpass trap filter. We will assume that our system has been ideal up to this point in that we were able to produce faithfully, without any error, the constant difference frequency signal at the input to the trap. As shown previously, in addition to the constant frequency signal we have the frequency modulated signal (Eq. 2.4). In order to facilitate the discussion we will assume that the modulation is slow enough and the frequency deviation small enough with respect to the center frequency so that at any instant of time the FM signal may be approximated by

$$\cos(\omega_0 + r)t \tag{2.7}$$

where r at this instant of time is the instantaneous deviation from the center frequency ω_0. Consider now the trap attenuation characteristic shown in Fig. 2.3

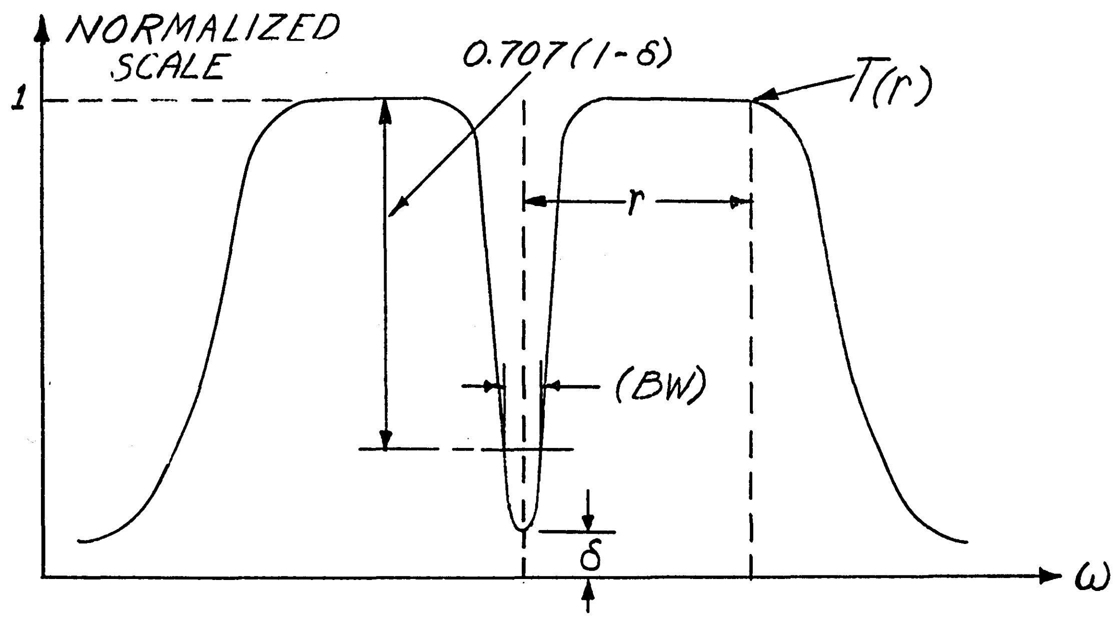


Fig. 2.3 Definition of Trap Attenuation Characteristic

Let the input signals to the trap be denoted by

$$
\begin{gathered}
E_s \cos \omega_0 t \\
\text{and} \\
a E_s \cos(\omega_0 + r)t \\
a \leq 1 \\
r \ll \omega_0
\end{gathered}
\qquad 2.8
$$

Let T (r) denote the normalized transfer function of the trap filter; within the pass band of this filter, $T(r)_{max}=$ unity. The minimum value of T(r) in the pass band occurs when r = 0, i.e., $T(0) = \delta$. The signals at the output of the trap will be sinusoids given by

$$
\begin{gathered}
T(0) E_s \cos \omega_0 t = \delta E_s \cos \omega_0 t \\
+ \quad a\, T(r) E_s \cos(\omega_0 + r)t
\end{gathered}
\qquad 2.9
$$

For values of a that are less than δ/β_2 , where β_2 represents the capture ratio of the FM demodulator that follows the trap system, the undesired signal (at ω_0 rad/sec) will continue to predominate by a sufficient margin to dictate the level of the average voltage at the output of the receiver, irrespective of the value of r. The undesired signal will than be faithfully captured over the whole modulation cycle (all values of r) and no perceptible trace of the desired message will be received.

As a is increased, so that it becomes greater than δ , we may still have situations arising where for some values of a, $a \cdot T(r_c) \leq \delta$. For values of a in this range the desired weaker signal will become the stronger signal at the output of the trap for all $r > r_c$ and the weaker of the two, $r < r_c$. We have a condition then where the desired signal is captured (assuming $\beta_2 \cong 1$) only during that fraction of the modulating cycle where $r > r_c$. Thus if the trap BW is small, then the fraction of the modulating cycle for which the desired signal is captured is increased for values of a in the range slightly greater than δ. As a is increased, the fraction of the modulating cycle for which the desired signal is captured increases since r_c approaches zero and $T(r_c)$ approaches δ.

The range of a values that form this transition region in which the receiver will shift capture from undesired to desired signal is given approximately by $\delta\beta_2 < a < \delta/\beta_2$. The response of the FM demodulator driven by the trap system when operating in this range of a is unable to capture satisfactorily either the desired or undesired signal without excessive distortion and cross talk from the other signal.

For values of a for which $\frac{\delta}{a}$ does not exceed the capture ratio, ρ_2, of the second demodulator (i.e., $a > \delta/\rho_2$), the desired signal will be captured with minimum distortion from the undesired signal. Only when the instantaneous frequency of the desired signal is very near or coincides with the center frequency of the trap, will distortion appear. This distortion will be in the form of a sharp burst of capture transition noise. The duration of these bursts are dependent upon the capture ratio of the receiver following the trap, and also, upon the bandwidth of the trap. Using a narrow BW trap reduces the fraction of the modulating cycle that the instantaneous frequency of the desired signal falls within the stop band of the trap. Both Sheftman and Rubissow have shown that the bursts of distortion are barely perceptible after low pass filtering in the output of the receiver. This distortion does affect the quality of the received weaker message somewhat, but it only has a small effect on the intelligibility of this message. Sheftman has shown that this form of distortion for a in this range does not exceed 3%.

Static-Trap System Without Strong Signal Predection

It will be recalled that the sole purpose of the first demodulator and variable frequency local oscillator was to derive an undistorted third FM signal whose instantaneous frequency differs from the instantaneous frequency of the undesired stronger signal by a constant that equals the center frequency of the trap.

This result may be approximated by a method which does not require demodulation-remodulation process. This cumbersome operation may be circumvented by utilizing a device capable of suppressing the weaker of two co-channel FM signals. For the moment, assume that we have an ideal weaker signal suppressor - a device that completely suppresses the weaker signal for any value of $\underline{a}$ ($\underline{a} < 1$). The output of this weak signal suppressor consisting of the stronger signal only, can be mixed with a local oscillator that generates a signal at the center frequency of the trap. The output of the suppressor-mixer operation will be the required translated stronger signal. Thus we see that we do not rely explicitly on the information contained in the stronger signal. Fig. 2.4 (Fig. 15. Ref. No. 1) depicts the block diagram of a system ulitizing a weak signal suppressor. This modified version of the fixed trap system shown here is the system used in this thesis to capture the weaker signal.

In practice, we can only expect an approximation to complete weaker-signal suppression by employing practical narrow-band limiters. These narrow-band limiters may be equipped with feed-forward or regenerative feedback in order to improve their weaker signal suppression properties. Thus, the appearance of the stronger signal in the output of the converter will not be entirely "cleansed" of the weaker signal as we assumed previously. One of the objectives of this thesis is to attempt to determine to what degree must the weaker signal be suppressed in order to derive an

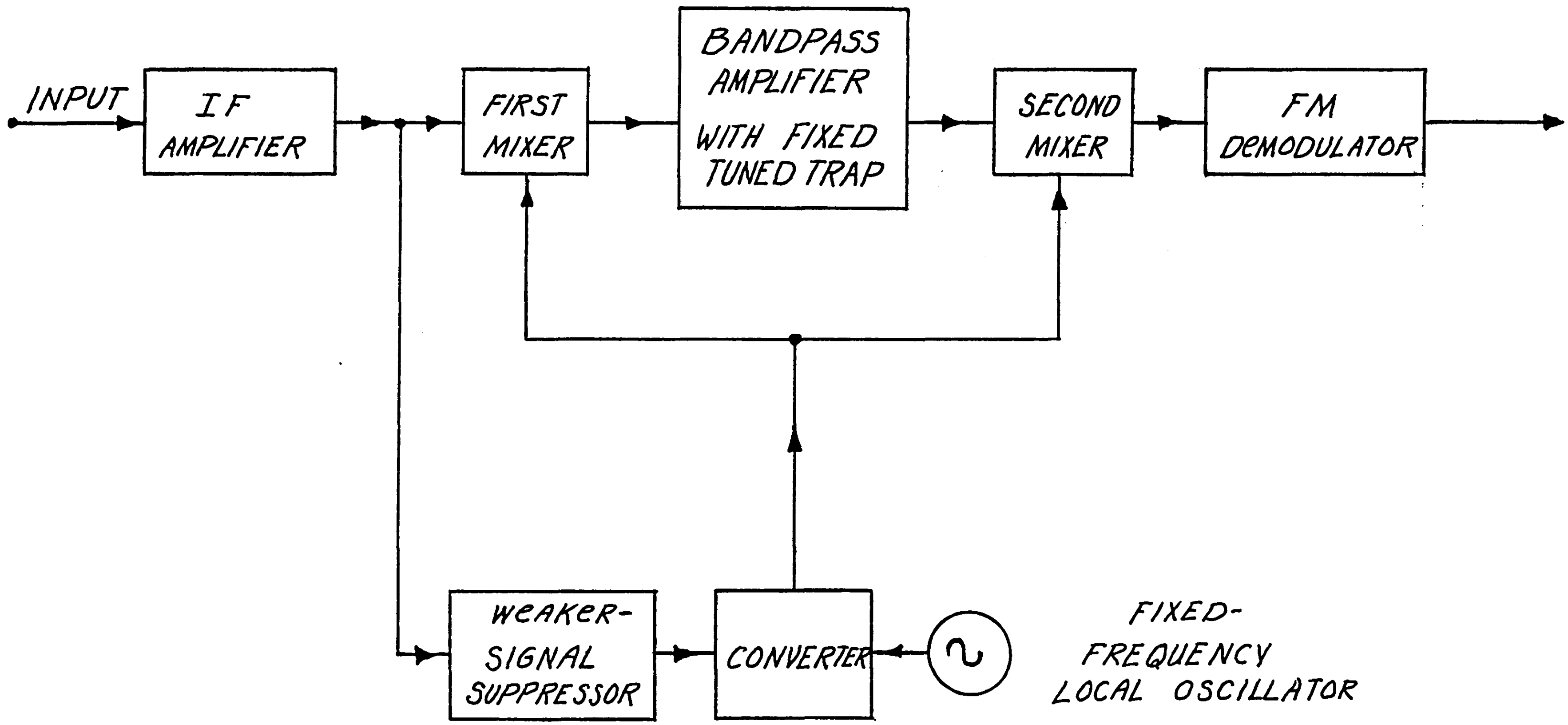


Fig. 2.4 A fixed trap system that does not require prior demodulation of the stronger signal

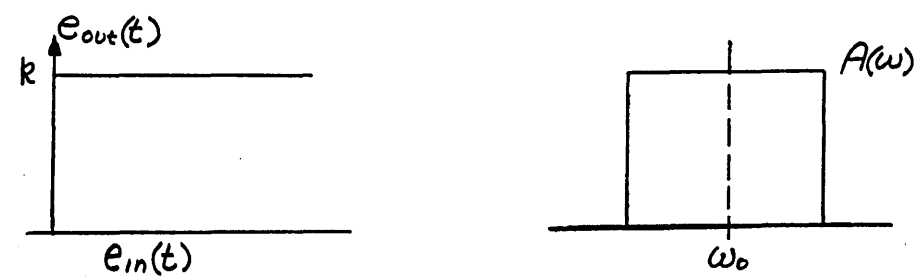
intelligible message from the desired signal in the output of the demodulator following the trap.

Some indication of how the use of non-ideal weaker signal suppressors will effect the capture performance of the receiver can be arrived at by performing a numerical analysis of the system for the practical limiter situation. Let us assume that the interference arises from the simultaneous presence of two signals of amplitudes E_s and $\underline{a}$ E_s ($\underline{a} \leq 1$) and frequencies p and p + r rad/sec ($r \ll p$) within the pass band of the IF amplifier. To simplify the analysis, it will be assumed that these two carriers have constant amplitudes and that their modulations are so slow that, for the time being, the frequencies can be considered stationary. Thus, the resultant signal at the output of the IF amplifier can be expressed as

$$e_{in}(t) = E_s \cos pt + \underline{a} E_s \cos(p+r)t \qquad 2.10$$

with $\underline{a} \leq 1$ and $r \ll p$

This is also the signal appearing at the input to the first mixer. This composite signal is processed by the weak signal suppressor which we will assume to be a cascaded section of narrow band limiters. However, we will idealize the limiters and their associated filters to the extent shown in Fig. 2.5.



Ideal Limiter Amplitude Characteristic

Ideal Filter Characteristic Following Limiter

Fig. 2.5 Idealized Limiter Characteristic

The ideal limiter is, by definition, a device that will operate upon the resultant of the two sinusoids described in Eq. 2.10 and deliver an output signal given by

$$e_{LIM}(t) = k \, Cos[pt + \theta(t)] \qquad 2.11$$

where k is a constant, and

$$\theta(t) = tan^{-1} \frac{a \, Sin \, rt}{1 + a \, Cos \, rt} \qquad 2.12$$

A Fourier analysis leads Eq. 2.11 and Eq. 2.12 to

$$e_{Lim}(t) = k \sum_{n=-\infty}^{\infty} A_n(a) \, Cos(p - nr)t \qquad 2.13$$

If this amplitude limited signal is filtered by a filter with a well defined bandwidth, $(BW)_{lim}$, the result may be expressed as

$$e_{NBL}(t) = k \sum_{n=-N}^{M} A_n(a) Cos(p - nr)t \qquad 2.14$$

Here, N and M represent the number of spectral components whose frequencies are greater and less than p, that are passed by the filter. The numbers N and M will depend upon the position of p within the IF pass band and upon the value of the frequency difference r relative to the IF filter bandwidth.

Eq. 2.14 represents the output from one narrow-band limiter. For our case, we have a group of cascaded limiters. We will express the output from these cascaded limiters by changing the notation slightly

$$e_{CAS}(t) = \sum_{n=-N}^{M} A_n(a') \, Cos(p - nr)t \qquad 2.15$$

In this equation, the arbitrary constant k is assumed to be unity. Use is also made of a different amplitude interference ratio $\underline{a}'$ in the functional notation for the coefficients $A_n(\underline{a}')$. The reason for this is that we are assuming a <u>cascaded</u> section of narrow-band limiters. The calculation of the coefficients $A_n(\underline{a}')$ are pertinent to the output of the last limiter whose effective input interference ratio $\underline{a}$ has been reduced to $\underline{a}'$ by the previous stages of limiting and filtering. An exact numerical analysis of the coefficients $A_n(\underline{a}')$ from the output of the last limiter would be unwieldly if we did not assume an effective input interference ratio $\underline{a}'$ to the last limiter. Making this assumption permits using the tabulated values for the coefficients after one stage of limiting.*

The output of the cascaded limiter section is linearly translated when hetrodyned with the local oscillator frequency, q, of the converter. The output of this conversion operation is expressed as

$$e_{conv}(t) = \sum_{n=-N}^{M} A_n(a') \cos(p-q-nr)t \qquad 2.16$$

It is assumed here that the BW of the filter following this converter is the same as the BW of the filter in the last limiting stage. This signal is mixed with the other input signal to the first mixer yielding an output, after filtering, given by the sum of two series

$$e_{mix\#1}(t) = \sum_{n=-N}^{M} A_n(a') \cos(q+nr)t + \sum_{n=-N}^{M} \underline{a}\, A_n(a') \cos[q+(n+1)r]t \qquad 2.17$$

The output of the first mixer, given by Eq. 2.17, is applied to the trap whose characteristic is shown in Fig. 2.3. However, in this analysis, we will consider discrete situations. The manner in which the instantaneous frequency p and the instantaneous frequency difference r vary, will be incorporated into the analysis by letting the limits of the summation N, M vary. Under these conditions then, it is more appropriate to represent the

---

* These coefficients are tabulated by Granlund in RLE Tech. Report No. 42 and by Baghdady in RLE Tech Report No. 252. (Ref. 6 and 7)

trap transfer function in a discrete manner such as T(nr) since the attenuation of components q, q+r, q+2r, -- q+kr will vary with n and with r.

The output of the trap may then be expressed as

$$e_{TRAP}(t) = \sum_{n=-N}^{M} T(nr)\, A_n(a')\, Cos(q+nr)t + \sum_{n=-N}^{M} a\, T[(n+1)r]\, A_n(a')\, Cos[q+(n+1)r]t \qquad 2.18$$

Note in the second term, the argument of the trap function is expressed as $T(n+1)r$ to coincide with the instantaneous frequency deviation from q of (n+1) r.

The trap output signal (Eq. 2.18) is now remixed with the signal coming from the converter given by

$$\sum_{m=-N}^{M} A_m(a')\, Cos(p-q-mr)t \qquad 2.19$$

This is the same signal as in Eq. 2.16 only we have changed the subscripts from n to m in order to facilitate the series multiplication that follows.

The output of the third mixer after filtering (same BW as limiter filters) is given by two signals represented in series form.

$$\begin{aligned} e_{OUT}(t) = &\sum_{n=-N}^{M} \sum_{m=-N}^{M} T(nr)\, A_m(a')\, A_n(a')\, Cos[p+(n-m)r]t \\ &+ \sum_{n=-N}^{M} \sum_{m=-N}^{M} a\, T[(n+1)r]\, A_m(a')\, A_n(a')\, Cos[p+(n-m+1)r]t \end{aligned} \qquad 2.20$$

This is the input to the FM demodulator.

We note from this expression the following: The desired signal at the frequency p+r rad/sec is contained numerous times in each series; the output is an explicit function of T(nr), the trap transfer function; the output is a function of the input interference ratio $\underline{a}$ explicitly and implicitly since the coefficients $A_n(\underline{a}')$ are determined by $\underline{a}$; the output is a function of the instantaneous frequency separation r relative to the limiter BW since r determines N, M for a fixed limiter BW. Thus we see that there are many variables to contend with in a complete evaluation of this expression. In order to facilitate a numerical analysis we will fix all but one of the variables, r. We will consider the system for a fixed input interference ratio $\underline{a} = 0.5$. For this input interference ratio, it will be assumed that in the process of cascading limiters the effective input interference ratio $\underline{a}'$ appearing at the input to the last limiter is given by $\underline{a}' = 0.2$ (a pessimistic estimate) thus permitting use of the tables presented by Baghdady for obtaining values for the spectral components $A_n(\underline{a}')$. Furthermore, we will also assume that the trap BW is very narrow with respect to the IF bandwidth and consider only situations where the instantaneous frequency difference, r, is much greater than the trap BW. For these assumptions,

$$T(nr) = \delta, \text{for } n = 0, \quad (\delta \cong 0.01)$$
and
$$T(nr) = 1, \text{for } n \neq 0$$

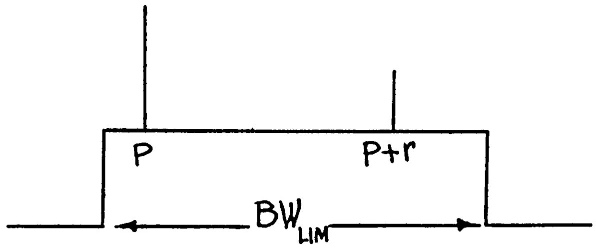
for the second series

$$T\ (n+1)r = \delta, \text{for } n = -1$$
and
$$T\ (n+1)r = 1, \text{for all } n \text{ except } n \neq -1.$$

The numerical evaluation of the demodulator input will be carried out by assuming different values of r and by considering different spectral configurations at the output of the limiter.

A summary of the spectral analysis is as follows:

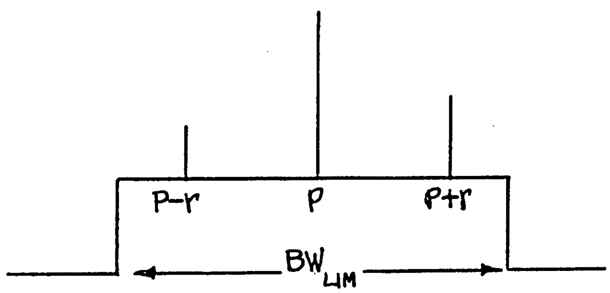
Case No. 1



OUTPUT	
Amplitude	Frequency
.0115	p
.05105	p + r

$$\frac{\text{Signal}}{\text{Interference added in phase}} \doteq \frac{.05105}{.0115} = 4.43$$

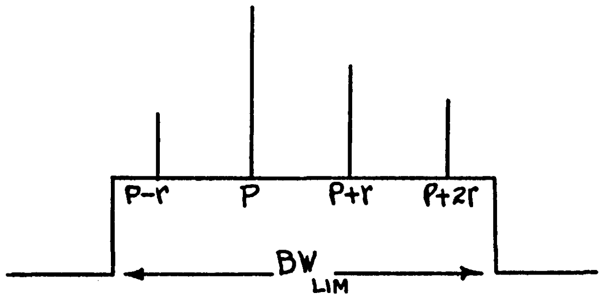
Case No. 2



OUTPUT		
Amplitude	Frequency	
- .0195	p	Interference
.08995	p - r	Interference
.46005	p + r	

$$\frac{\text{Signal}}{\text{Interference added in phase}} \doteq \frac{.46}{.1095} = 4.2$$

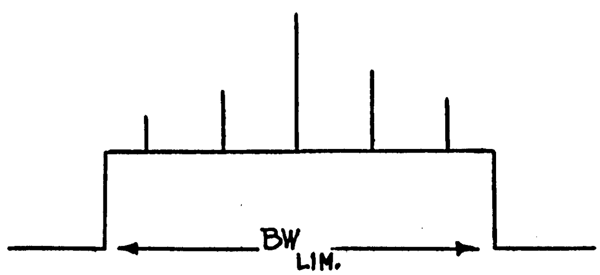
Case No. 3



OUTPUT		
Amplitude	Frequency	
- .019725	p	Interference
.09645	p - r	Interference
- .0099525	p + 2r	Interference
.40556	p + r	

$$\frac{\text{Signal}}{\text{Interference added in phase}} \cong \frac{.4056}{.1261} = 3.22$$

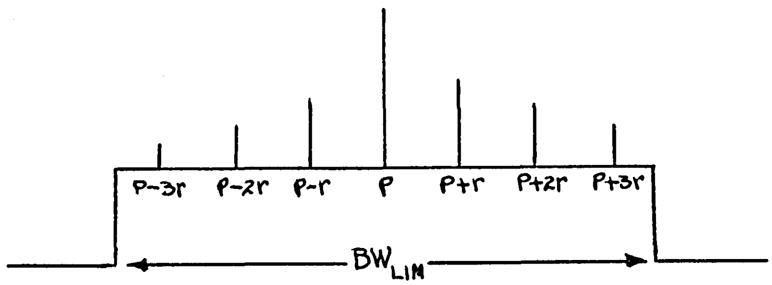
Case No. 4



OUTPUT		
Amplitude	Frequency	
- .0146	p - 2r	Interference
.105	p - r	Interference
- .0203	p	Interference
- .0466	p + 2r	Interference
.4042	p + r	

$$\frac{\text{Signal}}{\text{Interference added in phase}} \cong \frac{.4042}{.1885} = 2.14$$

Case No. 5



OUTPUT

Amplitude	Frequency	
.00241	p - 3r	
- .01530	p - 2r	
- .00509	p - r	
.02004	p	Interference
.00456	p 2r	
- .00083	p 3r	
.40415	p r	

$$\frac{\text{Signal}}{\text{Interference added in phase}} = \frac{.40415}{.04823} = 8.33$$

In all the cases considered we note that the desired signal (the signal at the frequency p + r (rad/sec) dominates the total sum of all the interfering components added in phase by a sufficient margin so that a demodulator would have no trouble capturing the desired message. For calculations involving more spectral components, we would have to take into account the actual shape of a typical trap characteristic.. For such a situation we would expect that the calculated signal to interference ratio would diminish. Another point to bring out is the effect of how the response would vary with a different input a. For a larger a, the relative interference suppression of each limiter diminishes.

However, a larger a also aids the desired component at the frequency $p+r$. For smaller a the relative suppression properties of the limiters improve but now a greater responsibility is placed upon the trap to suppress the rest of the interfering components in order that the signal at the frequency $p+r$ may still dominate. Therefore, we can expect to find a particular a for which the response is optimum (that value of a for which the amplitude of the spectral component at $p+r$ is a maximum). Our aim here in this analysis was not to try to determine this value but rather to examine for a restricted case the feasibility of using narrow band limiters. In this light, the results are encouraging.

Although the use of simple narrow band limiters avoids the problem of critical circuitry in the first demodulator-modulator unit, it introduces a new unavoidable proplem which is difficult to compensate. This problem is the time delay acquired by the stronger signal message as a result of the filtering in the weaker-signal-suppressor and converter. The significance of this time delay is subtle and its effect can be serious. The theory underlying this phenomenon will be presented in Chapter III. Suffice it to say, for now, that because of the accumulated delay through the weaker signal suppressor, we cannot produce without some form of compensation an FM signal in the suppressor-converter channel output whose instantaneous frequency differs from the instantaneous frequency of the stronger undesired signal in the upper channel by a constant frequency difference.

Summary of Sheftman's Investigation of the Fixed Trap System (4)

In his thesis Sheftman investigated the fixed trap system shown in Fig. 2.2. However, he simulated the predection of the stronger signal by arranging the system as shown in Fig. 2.6 (Fig. 5 of Ref. 4.)

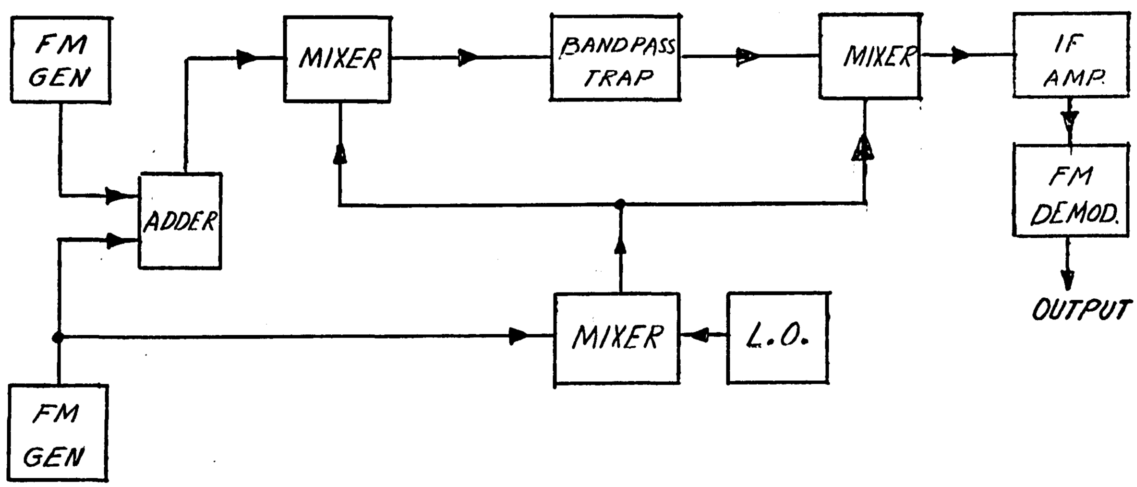


Fig. 2.6 Sheftman's Arrangement of Fixed Trap Circuit Simulating Predetection of Stronger Signal

In the laboratory, the weaker and stronger signals were available from two separate FM generators making it possible to simulate conditions in this manner. Actually, the way the system is set up corresponds to the situation where ideal weaker signal suppressors are used to completely suppress the weaker signal so that the output of this ideal suppressor consists of the strong signal only. This thesis is concerned with carrying this a step forward by investigating the same system but now utilizing practical suppressors rather than ideal weaker signal suppressors.

We have seen earlier in this chapter that the fixed trap system as shown in Fig. 2.2 is capable of capturing the weaker signal for any a in the range $\delta/\rho_2 \leq \underline{a} \leq \rho_1$. However, for the simulated conditions under which Sheftman studied the fixed trap, there is no first demodulator hence there is no upper limit ρ_1 for a in the capture of the weaker signal. Thus capture can be achieved for these simulated conditions for any $\underline{a} > \delta/\rho_2$, where ρ_2 is the capture ratio of the demodulator driven by the trap.

Under these conditions, the principle region of interest in the capture characteristics is the capture transition region centered near $\underline{a} = \delta$. For this idealized situation we should expect excellent results. In general, the results of Sheftman's system were encouraging however the results also indicated some peculiar response trends. A summary of Sheftman's results obtained for the simulated condition of predetection of the stronger signal is presented in tabular form below. (Obtained from Pages 62 through 66 in Ref. 4)

SUMMARY OF SHEFTMAN'S RESULTS

Modulation of Weaker Signal	Modulation of Stronger Signal	Location of capture trans. region centered at $\underline{a} = a_c$	Input interference ratio for which dist. is received weak signal is below 10%
±75kc @ 1kc	unmodulated	.0026	.005
±35kc @ 1kc	±35kc @ 400 cps	.0135	.04
±75kc @ 1kc	±35kc @ 400 cps	.0138	.034
±75kc @ 1kc	±75kc @ 400 cps	.037	.077
±75kc @ 400cps	±75kc @ 1kc	.068	.26

From this summary note first that the capture of the desired signal message, when the interfering signal is CW, is excellent. Consulting Sheftman's curve for this condition (Page 62, Ref. 4) we note that the capture transition region is affected by the following: (1) the trap attenuation characteristic T(r); (2) the center frequency attenuation of the trap, δ; (3) the stronger signal capture characteristic of the trap driven demodulator.

However, as the modulation characteristics of the interfering signal are changed, some peculiar trends are evident. We note first that when the modulation of the desired signal is changed (row 2 and 3 of summary) while the modulation of the undesired signal is fixed, there is no significant change in the response. From this we deduce that the characteristics of the weaker signal modulation do not appear to effect the response. However, when the frequency deviation (modulating frequency fixed at 400 cps) of the stronger FM signal is increased from $\pm$ 35 kc to $\pm$ 75 kc while the weaker FM signal modulation is fixed, we note a degradation of capture response (compare row 3 with 4). When the frequency deviation of both signals are fixed, but the modulating frequencies are interchanged, we note a much greater degradation of response (compare row 4 with 5).

In conclusion, we see that Sheftman's fixed trap receiver is vulnerable to the severity of the modulation of the stronger signal. The response is impaired when either the frequency deviation or the modulating frequency of the interfering stronger signal is increased.

In order that we may investigate the capture capabilities of the fixed trap system by approximating the demodulator-modulator unit with practical weaker signal suppressors and a converter, it is imperative that the fixed trap circuit be responding in an optimum manner for all modulation conditions of the interfering stronger signal. Thus a portion of this thesis is devoted to a detailed evaluation of the fixed trap so that a new circuit can be redesigned which will not exhibit the degradation of response characteristic of Sheftman's circuit.

EVALUATIONS AND OBSERVATIONS OF THE FIXED TRAP CIRCUIT

Before we discuss a specific redesign of the fixed trap circuit, we would like to evaluate in detail and to make further observations of Sheftman's fixed trap circuit. From the results of this evaluation we will offer possible explanations for the degradation of response when either the modulating frequency or the deviation of the stronger signal is increased. This will provide criteria upon which a redesign can be based.

We note first that Sheftman's circuit was certainly free of spurious effects such as parasitic oscillations, stray coupling, extraneous mixer signals etc. as is evidenced by the fact that it performed so well when the interfering signal was unmodulated. If any spurious effects such as these did exist they would certainly manifest themselves for all interference conditions.

As we have seen in the previous chapter, the response of the original fixed-trap circuit is impaired when either the deviation or the modulating frequency of the stronger interfering signal is increased. As either one of these modulation parameters of the stronger signal is altered, the FM signals involved are directly affected by the response of any filters employed in the circuits. Accordingly, the reasons for which the system response deteriorates must lie in the subtleties involved in the instantaneous frequency of the filter response.

Quasi-Stationary Response of Linear Filters *

Consider at the input to a linear filter an FM circuit

* The discussion in this and the next section is a summary of the analysis presented by Baghdady in RLE Tech Report No. 332.

excitation given by

$$i(t) = e^{j\int_0^t \omega_i(x)\,dx} \qquad 3.0$$

where

$\omega_i(t) = \omega_c + \theta'(t)$

ω_c is the carrier frequency

$\theta'(t)$ is some arbitrary function of time representing the frequency modulation.

We then have in the output of this filter, after initial transients have subsided, the forced response denoted by

$$e(t) = E(t)e^{j\int_0^t \omega_i(x)dx}$$

where 3.1

$$E(t) = Z(j\omega_i) = A(j\omega_i)e^{j\phi(\omega_i)}$$

$E(t)$=complex envelope of this response

Expressing the output of the filter in this form at first seems trivial since this is the manner in which we would normally proceed when the input is a constant frequency signal. However, expressing the output in this form when the excitation is an FM signal tacitly assumes that the filter is able to follow the FM excitation through quasi-stationary states. This implies that at any instant of time, t_0, the complex envelope E(t) of the steady state response is approximated by evaluating the sinusoidal-steady-state value of the system function at the frequency $\omega_i(t_0)$ as though this frequency were maintained for a sufficiently long time so as to allow a build up of the response to its sinusoidal-steady-state value.

Thus we see that the prime requirement of all our filters is that they be responding to the FM excitation in a quasi-stationary manner. This is an especially important requirement of the filters in the weaker-signal suppressor and the filters between the converter and the first mixer, otherwise transient distortion of the interfering stronger signal message will result and we will no longer meet the requirements imposed on the translated stronger signal.

For a periodic frequency modulation represented by $\theta(rt)$, where r is the repetition in rad/sec the complex amplitude E(t) of the filter response can be expressed in the form of a series given by

$$E(rt) = \sum_{n=0}^{\infty} G_n(rt)\, r^n \qquad 3.2$$

The coefficients $G_n(rt)$ are listed by Baghdady (Pg. 8, RLE Tech. Report No. 332). The first two will be presented here.

For $n = 0$

$$E(rt) = Z(J\omega_i) \qquad 3.3$$

For $n = 1$

$$E(rt) = Z(J\omega_i) + j\,\frac{1}{2!}\,\theta''(rt)\, Z''(J\omega_i) \qquad 3.4$$

Thus if $\theta(t)$ and $Z(J\omega_i)$ have properties that make the terms for $n \geq 1$ negligible compared with the first term, than the steady state response of the filter is given by Eq. 3.3.

Baghdady has shown that the conditions for which we may approximate the complex amplitude E(t) is given by

$$\epsilon_{rel.} = \frac{1}{2}\,\left|\theta''(t)\right|_{max.} \cdot \left|\frac{Z''J\omega}{Z\,J\omega}\right|_{max.} \ll 1 \qquad 3.5$$

For sinusoidal modulation given by

$$\theta(t) = \frac{\Delta\omega}{\omega_m} \sin \omega_m t \qquad 3.5$$

then we have for $\theta''(t)$,

$$\left|\theta''(t)\right|_{max.} = \omega_m \cdot \Delta\omega \ . \qquad 3.6$$

For any Butterworth filter, the transfer function may be expressed by

$$Z_{bn}(j\Omega) = \frac{1}{F(\frac{\Omega}{\alpha})} \qquad 3.7$$

Ω = deviation from center frequency

$\alpha = BW/2$

Using this response function Baghdady manipulated the expression $\left|\frac{Z''}{Z}\right|_{max}$ into the form

$$\left|\frac{Z''_{bn}}{Z_{bn}}\right|_{max} = \frac{k_n}{(BW)^2} \qquad 3.8$$

where k_n is a constant with the subscript n denoting the order of the filter (number of poles).

Substituting Eq. 3.6 and 3.8 into Eq. 3.5, the error criterion for sinusoidal frequency modulation becomes

$$\epsilon = \frac{k_n}{2} \cdot \frac{\omega_m \cdot \Delta\omega}{(BW)^2} \ll 1 \qquad 3.9$$

If we consider an error of magnitude less that 1/10 as tolerable, then

$$\epsilon = \frac{k_n}{2} \ \frac{\omega_m \cdot \Delta\omega}{(BW)^2} < \frac{1}{10}$$

or solving for the bandwidth we have

$$BW_{min} > \sqrt{5 k_n \ \omega_m \cdot \Delta\omega} \qquad 3.10$$

Thus we have an explicit expression for the minimum bandwidth of the filter determined by the order of the Butterworth filter used, and the modulation parameters of the FM excitation. If we design a filter for a particular FM signal whose BW exceeds this minimum requirement then we will be assured that the complex envelope of the response will be essentially determined by the static filter characteristic. FM transient distortion introduced by the sluggishness of the filter will then be negligible.

Instantaneous Frequency of Filter Response

Now that we have described the minimum BW requirement of the filters, the next step is to examine the instantaneous frequency of the filter response. We are particularly interested in whether or not the instantaneous frequency of the output FM signal has been distorted by factors other than transient distortion.

Under the assumption of quasi-static response, the complex envelope of the filter response E(t) is given by

$$E(t) = Z(j\omega_i) = A(\omega_i)\,e^{j\phi(\omega_i)} \tag{3.11}$$

where $\omega_i = \omega_c + \theta'(t)$

and $A(\omega_i)$ and $\phi(\omega_i)$ are the amplitude and phase characteristics of a bandpass filter.

The response of the filter is then given by

$$e(t) = A(\omega_i)\,e^{j[\int_0^t \omega_i(x)dx + \phi(\omega_i)]} \tag{3.12}$$

The exponent represents the instantaneous phase angle of the filter response. The time derivation of the argument of the exponent yields the instantaneous frequency of the filter response,

$$\omega_{io}(t) = \frac{d[\int_0^t \omega_i(x)dx + \phi[\omega_i(t)]]}{dt}$$

$$\omega_{io}(t) = \omega_i(t) + \frac{d\phi[\omega_i(t)]}{dt}$$

$$= \omega_c + \theta'(t) + \frac{d\phi[\omega_i(t)]}{d\omega_i(t)} \cdot \frac{d\omega_i(t)}{dt}$$

$$= \omega_c + \theta'(t) + \phi'(\omega_i)\cdot\theta''(t) \qquad 3.13$$

The function $\phi(\omega_i)$ may be denoted as the instantaneous time delay of the filter. Thus we can see that even under conditions of quasi-stationary response, the term $\phi'(\omega_i)\cdot\theta''(t)$ introduces distortion into the frequency waveform. This distortion is a function of only the derivative of the filter phase characteristic and the rate at which the frequency is modulated. As we shall presently see, this distortion is a combination of the unavoidable time delay of the filter in addition to distortion of the modulation in the form of message waveform alteration. In order to show this it is necessary to manipulate the expression for $\omega_{io}(t)$ (Eq. 3.13) into a more explicit form.

First consider the function $\phi'(\omega_i)$. We will expand this function in a Taylor series about the frequency $\omega_i = \omega_c$.

$$\phi'(\omega_i) = \phi'(\omega_c) + \phi''(\omega_c)[\omega_i - \omega_c] + \frac{\phi'''(\omega_c)[\omega_i-\omega_c]^2}{2!}$$

$$---- + \frac{\phi^{n+1}(\omega_c)[\omega_i-\omega_c]^n}{n!} \qquad 3.14$$

However $[\omega_i-\omega_c] = \theta'(t)$; substituting $\theta(t)$ for $\omega_i-\omega_c$ and collecting terms we have

$$\phi'(\omega_i) = \phi'(\omega_c) + \sum_{n=1}^{\infty} \frac{\phi^{n+1}(\omega_c)}{n!}[\theta'(t)]^n \qquad 3.15$$

If we now substitute this into the expression for $\omega_{io}(t)$ (Eq. 3.13) we have

$$\omega_{io}(t) = \omega_c + \theta'(t) + \phi'(\omega_c)\theta''(t) + \theta''(t)\sum_{n=1}^{\infty}\frac{\phi^{n+1}(\omega_c)}{n!}[\theta'(t)]^n \qquad 3.16$$

Now consider an arbitary function given by $\theta'(t+\tau)$. Expanding $\theta'(t+\tau)$ about $(t+\tau)=t$ we have

$$\theta'(t+\tau) = \theta'(t) + \theta''(t)\tau + \frac{\theta'''(t)\tau^2}{2!} + ---$$

$$= \theta'(t) + \tau\,\theta''(t) + \sum_{k=2}^{\infty} \frac{\theta^{k+1}(t)\tau^k}{k!}$$

If we let $\tau = \phi'(\omega_c)$ then we have

$$\theta'(t+\phi'(\omega_c)) = \theta'(t) + \theta''(t)\phi'(\omega_c) + \sum_{k=2}^{\infty} \frac{\theta^{k+1}(t)[\phi'(\omega_c)]^k}{k!} \qquad 3.17$$

rearranging, we get

$$\theta'(t) + \theta''(t)\phi'(\omega_c) = \theta'[t+\phi'(\omega_c)] - \sum_{k=2}^{\infty} \frac{\theta^{k+1}(t)[\phi'(\omega_c)]^k}{k!} \qquad 3.18$$

Substituting into Eq. 3.16 we have, finally

$$\omega_{io}(t) = \omega_c + \theta'[t+\phi'(\omega_c)] + \theta'(t)\sum_{n=1}^{\infty} \frac{\phi^{n+1}(\omega_c)[\theta'(t)]^n}{n!} - \sum_{k=2}^{\infty} \frac{\theta^{k+1}(t)[\phi'(\omega_c)]^k}{k!} \qquad 3.19$$

Expanding a few of the terms of each series we obtain

$$\begin{aligned}\omega_{io}(t) = {} & \omega_c + \theta'[t+\phi'(\omega_c)] \\ & + [\theta''(t)\phi''(\omega_c)\theta'(t) + \theta''(t)\frac{\phi'''(\omega_c)[\theta'(t)]^2}{2!} + ----] \\ & - [\frac{\theta''(t)}{2!}[\phi'(\omega_c)]^2 + \frac{\theta'''(t)}{3!}[\phi'(\omega_c)]^3 + ----]\end{aligned} \qquad 3.20$$

Thus the constituents of the distortion present in the instantaneous frequency in the output of a filter even under quasi-static response are manifested in the form

a. an unavoidable time delay

b. a part that brings out the effect of non-linearties in the phase characteristic (second expression in brackets involves higher order derivatives of $\phi'(\omega_c)$)

c. a part that brings out the effect of the slope of phase characteristic (third expression in brackets involves higher order powers of $\phi'(\omega_c)$). This part turns out to be a mathematical phenomenon that we may call a truncation error.

If we assume a perfectly linear phase characteristic $\phi(\omega)$ we have for the instantaneous frequency of the response

$$\omega_{io}(t) = \omega_c + \theta'[t + \phi(\omega_c)] \qquad 3.21$$

Thus we have residual distortion even under quasi-stationary response conditions in the form of an unavoidable time delay of the frequency modulation given by $\phi'(\omega_c)$, the slope of the phase characteristic at the center frequency of the filter.

Baghdady has also derived conditions for which the waveform distortion terms may be neglected. The error in neglecting the distortion due to $\phi'(\omega_c)$ is given by

$$\mathcal{E}_\phi \leq \frac{1}{2} \left| \theta'''(t) \cdot \phi'^2(\omega_c) \right| \qquad 3.22$$

If we assume sinusoidal modulation of the form

$$\theta'(t) = \Delta\omega \sin \omega_m t$$

then

$$\mathcal{E}_\phi \leq \frac{1}{2} \left| \omega_m^2 \Delta\omega \cdot \frac{b^2}{\alpha^2} \right| \qquad 3.23$$

The relative error is

$$\frac{\mathcal{E}_\phi}{\Delta\omega} \leq \frac{1}{2} b^2 \left(\frac{\omega_m}{\alpha}\right)^2 \qquad 3.24$$

where b is a constant dependent on type of filter used.

$$\alpha = \frac{BW}{2}$$

For a single-tuned circuit $b = 1$, $\alpha = 200\ kc$, $\omega_m = (2\pi)\,5\ kc$

$$\frac{\mathcal{E}\phi}{\Delta\omega} \leq \frac{1}{2}\left(\frac{5}{200}\right)^2 \qquad 3.25$$

$$\leq \frac{1}{3200} \approx .0003 \text{ or } .03\ \%$$

For a double-tuned circuit $b = \sqrt{2}$, $\alpha = 100\ kc$

$$\frac{\mathcal{E}\phi}{\Delta\omega} \leq \left(\frac{5}{100}\right)^2 = .0025 \text{ or } .25\ \% \qquad 3.26$$

The error in neglecting the distortion due to the nonlinearties of the phase characteristic is given by

$$\mathcal{E}_{non'L'}\,\phi \leq \frac{1}{2}\left|\theta''(t)\right| \cdot \left|\theta'(t)\right|^2 \cdot \left|\phi'''(\omega_c)\right| \qquad 3.27$$

Assuming sinusoidal frequency modulation as in Eq. 3.23 we have

$$\mathcal{E}_{non'L'}\,\phi \leq \frac{1}{2}\ \omega_m\,(\Delta\omega)^3\,\frac{k}{\alpha^3} \qquad 3.28$$

$$\frac{\mathcal{E}_{non'L'}\,\phi}{\Delta\omega} \leq \frac{k}{2}\left(\frac{\Delta\omega}{\alpha}\right)^2 \frac{\omega_m}{\alpha} \qquad 3.29$$

For a typical single-tuned circuit $k = 2$, $\alpha = 200\ kc$

The relative error is

$$\frac{\mathcal{E}_{non'L'}\,\phi}{\Delta\omega} \leq \frac{(\Delta\omega)^2}{\alpha^2}\ \frac{\omega_m}{\alpha} \qquad 3.30$$

$$\text{for} \quad \Delta\omega = (2\pi)\,75\ kc\ ; \quad \omega_m = (2\pi)\,5\ kc$$

$$\frac{\mathcal{E}_{non'L'}\,\phi}{\Delta\omega} \leq \left(\frac{75}{200}\right)^2 \frac{1}{40} = .0035 = .35\ \% \qquad 3.31$$

For a double-tuned circuit (second order Butterworth)

$$k = 2\sqrt{2}\ ; \quad \alpha = 100\ kc \qquad 3.32$$

$$\frac{\mathcal{E}_{non'L'}\,\phi}{\Delta\omega} \leq \left(\frac{75}{100}\right)^2 \cdot \frac{1}{20} \approx .03 = 3\ \%$$

Evaluation of the Fixed Trap Circuit

We are now able to consider possible reasons for the degradation of response exhibited by the fixed trap circuit. The discussion that follows concerns a redesigned version of the fixed trap circuit based on the original system block diagram Fig., 2.2 This redesigned circuit exhibited the same trends of response when cursory tests were performed for the same simulated conditions studied by Sheftman.

We will begin the evaluation by restating the requirement imposed on the weaker signal suppressor and the converter unit. The function of these circuits is to suppress the desired weaker signal with respect to the undesired stronger signal and to perform a linear translation of the stronger signal to a new IF frequency. This operation produces a stronger signal whose instantaneous frequency differs from the instantaneous frequency of the incoming stronger signal by a constant amount.

The instantaneous frequency of the input stronger signal is expressed by

$$\omega_{ii} = \omega_1 + \theta_s'(t)$$

If we assume complete suppression of the weaker signal, the instantaneous frequency of the stronger signal in the output of the converter is given ideally by

$$\omega_{ic} = \omega_1 - \omega_0 + \theta_s'(t)$$

The process of mixing the input signal and the converter output signal/produces the difference frequency component ω_0. However, we note from the previous analysis that there is always some distortion of the frequency modulation of the stronger FM signal as it is processed by the filters in the weak signal suppressor and the filters and amplifiers between the converter and first mixer. This distortion manifests itself in the form of time delay in the modulation message due to the finite slope of the phase characteristic of the filters. In addition, it is possible to have harmonic distortion and intermodulation terms present in the event that the filters possess nonlinear phase characteristics.

We note also that the stronger FM signal will be amplitude modulated in accordance with the shape of the filter amplitude characteristics.

Thus the translated stronger signal is appropriately described by

$$e_c(t) = A_c(\omega_i)\, e^{j\int_0^t \omega_{ic}(x)dx} \qquad 3.33$$

where $\omega_{ic}(t) = \omega_c + \theta'[t + \phi'(\omega_c)] + f(t - \tau_c)$

$\omega_c = \omega_1 - \omega_0$

$A_c(\omega_{ic})$ amplitude characteristic of the filters in converter channel

$f(t-\tau_c)$ harmonic and intermodulation distortion introduced by nonlinear phase characteristics of all the filters

$\phi'(\omega_c)$ accumulative time delay through all the filters

Note that we are still assuming that all the filters excited by the stronger FM signal have responded in a quasi-stationary manner so that FM transients are neglibible. We are also making the assumption of complete weaker-signal suppression by the limiters.

The signal from the converter $e_c(t)$ (Eq. 3.33) is now mixed with the stronger input signal from the IF amplifier given by

$$e_i(t) = A_i(\omega_{ii}) e^{j\int_0^t \omega_{ii}(x)dx} \qquad 3.34$$

$$\omega_{ii}(t) = \omega_1 + \theta'(t)$$

$A_i(\omega_{ic})$ = amplitude characteristic of IF filter

The process of mixing and filtering yields the product

$$e_T(t) = e_i(t) \cdot e_c(t)$$

$$e_T(t) = A_i(\omega_{ii}) A_c(\omega_{ic}) e^{j\int_0^t \omega_{iT}(x)dx} \qquad 3.35$$

where

$$\omega_{iT}(t) = \omega_0 + \theta'(t) - \theta'[t + \phi'(\omega_c)] - f(t - \tau_c) \qquad 3.36$$

The function $e_T(t)$ represents the signal feeding the trap circuit. To it must be added the frequency modulated weaker signal. For the moment we will not consider the weaker signal since we are only interested in how effective the circuit is in reducing the stronger signal to an unmodulated carrier.

We note from Eq. 3.35 that the first mixer output signal is amplitude modulated in accordance with the product of the amplitude characteristics of the filters in both channels preceding the first mixer. In addition, as the expression for the instantaneous frequency $\omega_{iT}(t)$ (Eq. 3.36) indicates we have not realized the reduction of the stronger signal to an unmodulated carrier because of the unavoidable delay incurred by the stronger-signal modulation as it is processed by the filters in the suppressor-converter channel. Furthermore, the message waveform distortion term $f(t - \tau_c)$ prevents cancellation of the stronger-signal message.

We see that the mixer output signal is frequency modulated in accordance with a residual signal given by

$$g(t) = \theta(t) - \theta(t - \tau_c) - f(t - \tau_c) \qquad 3.37$$
$$\tau_c = -\phi'(\omega_c)$$

From this expression we note that even if we could neglect the waveform distortion term $f(t-\tau_c)$ we still have not achieved complete cancellation of the modulation because of the time delay incurred by the stronger signal message in the suppressor-converter channel. For one or two cascaded Butterworth filters, the examples worked out previously indicated that distortion due to nonlinearties of the phase characteristic can be neglected. However when a number of Butterworth filters are cascaded as they are in the suppressor-converter channel (narrow-band limiter filters, mixer filters, amplifier filters) this distortion due to nonlinear phase characteristics may not be negligible at all. The effect of nonlinearties may also be compounded when some of these filters are slightly mistuned.

Now let us see how this discussion is related to the observed trends of system response degradation when the modulation characteristics of the stronger FM signal are changed. Consider first the effect of increasing the frequency deviation of the stronger signal while the modulation rate is fixed. We note that a greater portion of the filter phase characteristic is swept. From the properties of Butterworth filters, we know that nonlinearties of the filter phase characteristics predominate at the edge of the filter passband. With wide frequency deviations of the FM signal, more of the nonlinear portions of the phase characteristics are swept giving rise to stronger harmonics and intermodulation terms. In addition, a larger deviation causes the FM signal to sweep over more of the filter amplitude characteristic. If the deviation exceeds the flat portion of the passband, (though still within the BW of the filter) a larger percentage of incidental amplitude modulation of the envelope will result. This explains why the response deteriorates with widely deviated signals.

Consider now the effect of increasing the modulation rate while holding the deviation constant. We note first that the incidental amplitude modulation of the envelope occurs at a faster rate. Secondly, the effect of time delays become more noticeable at the higher sweep rates. For example, assume that we had a time delay of 6 μsec. and that we originally started with a sweep rate of 50 cps. At 50 cps, a 6 μsec. delay represents a phase shift of only 0.0019 rad. At 5,000 cps the phase shift is 0.19 rad. or approximately 11°. As Eq. 3.38 indicates, this is certainly enough phase shift to prevent cancellation of the modulations.

$$\theta'(t) - \theta'(t-\tau) = \Delta\omega[\sin \omega_m t - \sin \omega_m (t-\tau)]$$

$$= \Delta\omega [\sin (2\pi) 5000t - \sin(2\pi 5000t - .19 \text{rad})] \quad 3.38$$

Observations of the Time Delay Effect

In order to confirm our expectations of the effect of time delay, the system was arranged as in Fig. 3.1.

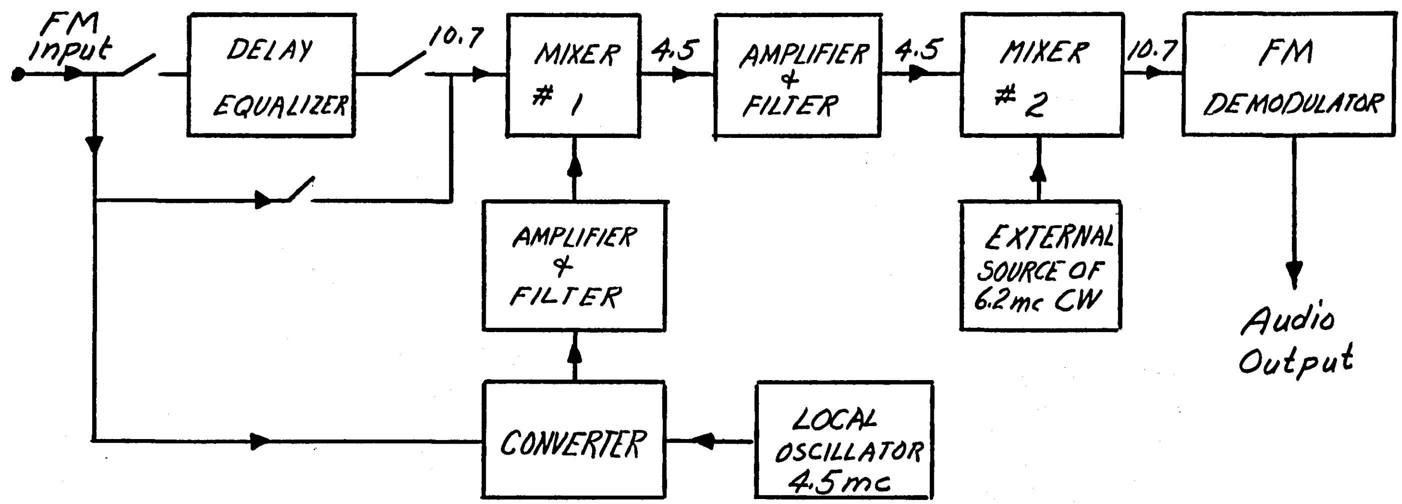


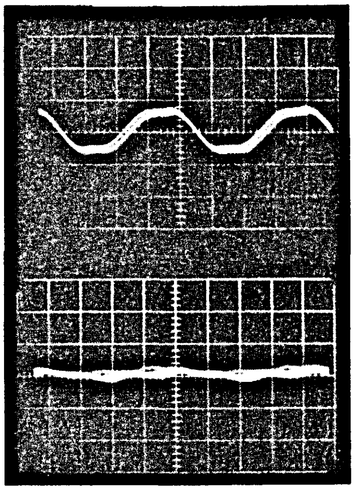
Fig. 3.1 System Arrangement to Study Delay Effect on Stronger-Signal Modulation in Output of First Mixer

For this arrangement, with the delay equalizer out of the circuit, the time delay of the converter channel was approximately 6μ sec. The 4.5 mc signal appearing in the output of the first mixer should theoretically be unmodulated. This signal is remixed with a 6.2 mc carrier and the result is demodulated by the FM demodulator. The oscillograms in Fig. 3.2 represent the supposedly constant frequency signal appearing in the output of the first mixer. Note how the audio amplitude increases with increasing modulation frequencies. The reason for this is that the time delay of 6μ sec. represents a greater phase shift of the message at the higher audio frequencies thereby preventing complete message subtraction.

The second set of oscillograms represent the same demodulated 4.5 mc signal in the output of the first mixer but now a delay

Pic. No. 1

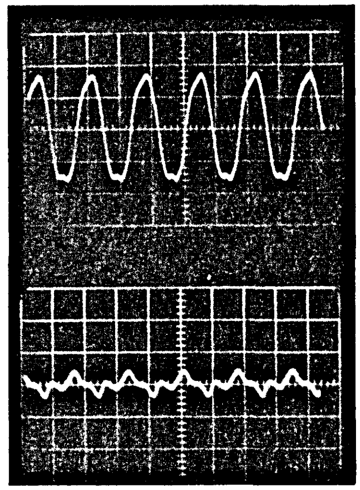
Vert.Axis
0.1 V/cm
Delay
Diff. 6 sec



Vert.Axis
0.05 V/cm
Delay
Equalized

Mod. of Mixer No. 1
Input 75kc @ 1kc

Pic. No. 2

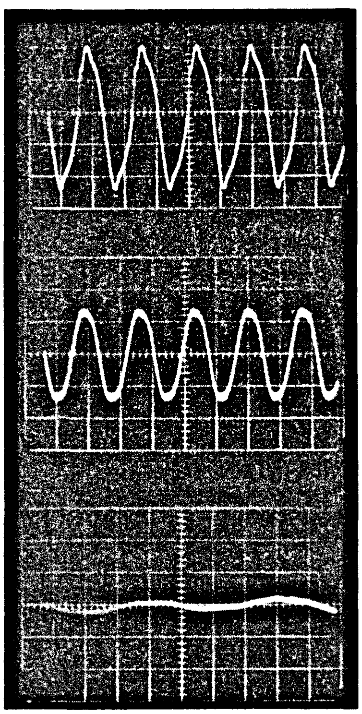


Vert.Axis
0.1 V/cm
Delay
Diff. 6μsec

Vert.Axis
0.1 V/cm
Delay
Equalized

Mod. of Mixer No. 1
Input 75kc @ 5kc

Pic. No. 3



Vert.Axis
0.05 V/cm
100 cps

Vert.Axis
0.05 V/cm
1000 cps

Vert.Axis
0.2 V/cm
5000 cps

Delay Differential is 20μ sec.(Due to Weak Signal Suppressor in Converter Channel) Deviation of Mixer Input Signal is 75 kc

Fig. 3.2 Oscillograms showing the effect of delay incurred by the stronger signal message in the suppressor-converter channel filters. Output of first mixer is demodulated in accordance with circuit arrangement shown in Fig. 3.1. All these oscillograms represent a supposedly carrier frequency signal; however after the process of demodulation we note some residual FM due to time delay and phase nonlinearities of the filters in both channels feeding the first mixer.

equalizer was placed in the signal grid circuit of the first mixer. This delay equalizer consisted of a cascaded section of IF amplifiers employing Butterworth filters similar to those used in the converter channel. There is a noticeable improvement since we note that the frequency modulation of the 4.5 mc output of the first mixer has definitely been suppressed as evidenced by the oscillograms.

The third set of oscillograms, show the effect of increased time delay when the entire weaker-signal suppressor was placed in the converter channel with no delay equalizer in the other channel. The situation has been noticeably worsened. The delay differential between the two channels is now of the order of 20μsec. Thus we need a delay equalizer in the upper channel capable of providing 20μsec. of delay of the stronger signal message. This delay equalizer must have a center frequency of 10.7 mc and a BW (to insure quasi-stationary response) exceeding 200 kc.

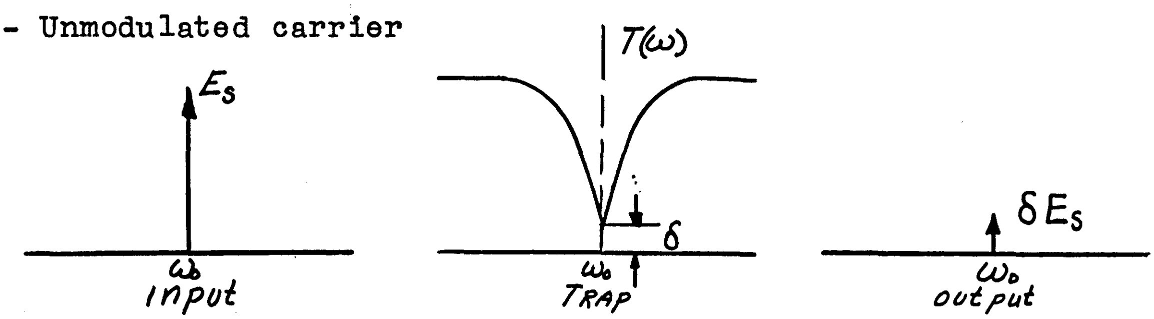
To conclude this section we note that when the upper channel is provided with a delay equalizer, both the stronger and the weaker signal messages are delayed by an equal amount. However, delay of the weaker signal message has no bearing on the reduction of the stronger-signal to a constant frequency signal. The compensation that the delay equalizer provides only ensures that the message of the stronger signal in the upper channel is highly correlated with the message of the stronger signal in the suppressor-converter channel. The only distinction between these two FM signals is their constant instantaneous frequency difference.

Discussion and Observations of Envelope Amplitude Modulation

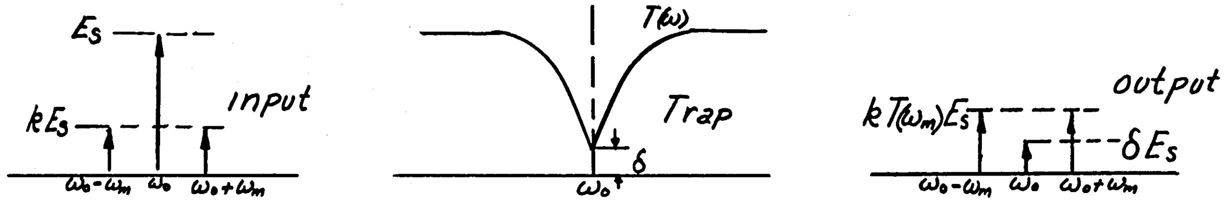
Previously, it was mentioned that the supposedly constant frequency signal was also amplitude modulated in accordance with the filter amplitude characteristic. The effect of AM on this signal is rather subtle. It will be recalled that we intend to filter out the 4.5 mc signal by using a narrow notch at the center frequency of a bandpass filter. However, when the signal is amplitude modulated, we have AM sidebands present on both sides of the 4.5 mc carrier. Although the trap filter will be attenuating the 4.5 mc carrier, the sideband components may be significantly displaced (4.5 + fm) from the center frequency to prevent their suppression. These insufficiently attenuated sidebands will prevent the system from suppressing the stronger signal effectively. The higher modulation frequencies, will result in less sideband suppression because these sidebands will be farther out from the center frequency of the trap. In addition, the wider the deviation of the FM signal, the greater the range of filter response swept and hence the greater the depth of envelope AM developed.

The following sketches show this spurious effect:

A - Unmodulated carrier



B - Amplitude modulated carrier

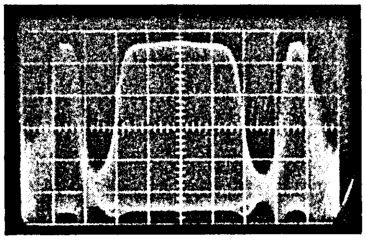


The oscillogram showing this effect is shown in Fig. 3.3. The first two oscillograms show the bandpass characteristics of the amplifier in the signal grid circuit of the first mixer and of the bandpass filter in the converter channel. The out-of-band response of the latter filter is an accumulation of noise as a result of the weak-signal suppressor (limiters) failing to limit the out-of-band noise. The noise peaks occur 250 kc from center frequency and are of no consequence to the inband signal.

The third oscillogram is the RF signal appearing in the output of the first mixer when only the one FM signal is present in both channels. Note the AM that results from the amplitude characteristics of the filters. This AM is only 5% of the carrier magnitude and at a rate of 1 kc since we were using a 1 kc audio sweep signal to frequency modulate the generators. Thus if we assume sinusoidal AM modulation, a 5% amplitude modulated signal results in sidebands components that are 2.5% of the magnitude of the carrier. If the trap bandwidth is sufficiently wide, the sideband components will also be attenuated. However, for low input ratios ($\underline{a} < 1/10$), the effect of the spurious AM and the fact that we have a narrow trap, would account for the deterioration of system performance.

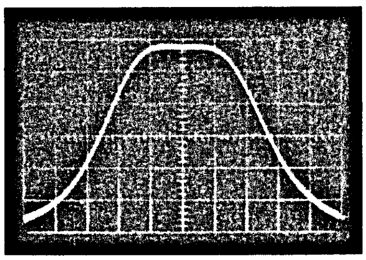
Sheftman attributed the deterioration of performance when

Pic. No. 1



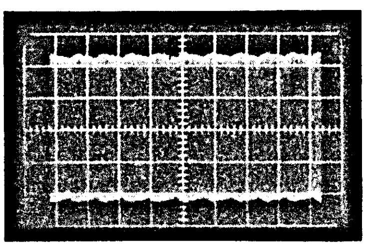
Filter Characteristic of
Converter-First Mixer
Interstage Circuits
Hor. Axis cal. = 75 kc/cm
Sweep Rate = 1 kc

Pic. No. 2



Filter Characteristic of
Delay Equalizer Feeding
First Mixer
Hor. Axis cal. = 75 kc
Sweep rate = 1 kc

Pic. No. 3



RF Output of First Mixer
(input to trap) when one
FM (stronger signal) is
at Input to System
Note AM @ 1 kc rate
% AM ≅ 5%

Fig. 3.3 Oscillograms showing amplitude vs. frequency response of filters in both channels feeding first mixer. Pic. No. 3 shows incidental AM of envelope of constant frequency signal appearing at input to trap.

the strong-signal modulation is increased solely to this incidental AM and the failure of the trap to attenuate sufficiently the sidebands due to this AM. He confirmed his reasoning by applying an AM signal (at the frequency of the trap) to the first mixer and monitored the output of the trap through a peak detector. The results showed that for various modulating frequencies and depth of modulation, the trap signal output varied in accordance with the reasoning given previously. (Pgs. 46 and 47 of Ref. No. 4)

Delay Distortion Introduced by Trap

Consider now the addition of the delay-compensating network as shown in Fig. 3.1. With this delay network, we have seen that we can produce approximately the required constant frequency signal at the input of the trap. However, let us now consider the system with both the weaker and stronger signals applied to the input.

Assuming complete weaker-signal suppression in the limiters and neglecting the spurious AM introduced by the filters, the output of the suppressor-converter channel is given by

$$e_c(t) = \cos[(\omega_1 - \omega_0)t + \theta_s'(t - \tau_c)] \qquad 3.39$$

This signal is applied to the first and last mixer.

With the delay equalizer in the upper channel, the process of mixing the converter output signal with the delayed incoming signals produces in the first mixer output the familiar constant frequency component and the FM component given by

$$e_T(t) = \cos\omega_0 t + a\cos[\omega_0 t + \theta_w'(t - \tau_d) - \theta_s'(t - \tau_c)] \qquad 3.40$$

τ_d = time delay of equalizer

τ_c = time delay of suppressor-converter channel

This signal is processed by the trap bandpass filter. This filter delays the modulation of the FM component so that the signal appearing at the output of the trap (neglecting FM transient distortion and phase nonlinearities) is given by

$$e_{T_0}(t) = \delta \cos \omega_0 t + a \cos[\omega_0 t + \theta_w'(t-\tau_d-\tau_T) - \theta_s'(t-\tau_c-\tau_T)] \qquad 3.41$$

where τ_T represents the accumulated delay through the band pass trap filter. This is mixed with the FM signal (Eq. 3.39) from the converter yielding

$$\begin{aligned} e_{mix\#3\,out}(t) &= \delta \cos[\omega_1 t + \theta_s'(t-\tau_c)] \\ &+ a \cos[\omega_1 t + \theta_w'(t-\tau_d-\tau_T) \\ &- \theta_s'(t-\tau_c-\tau_T) + \theta_s'(t-\tau_c)] \end{aligned} \qquad 3.42$$

Note that now in the reappearing weaker signal, we have not removed completely the stronger-signal modulation. The residual stronger signal modulation appearing in this term will manifest itself as intermodulation when the weaker signal is demodulated. The higher the stronger signal sweep rate, the greater the effect of this intermodulation. Thus we have established another possible reason for the degradation of system response.

In concluding this chapter, we see that from the discussion of the instantaneous frequency response we were able to pinpoint the reasons for the original system shortcomings. The primary source of trouble is the time delay acquired by the stronger-signal message in the suppressor-converter channel. As we have seen, this may be compensated by an equalizer in the other channel. The time delay accumulated by the weaker-signal through the trap is another cause for failure though not as serious, since the delay differential is small (less than 2μ sec.).

The problem of nonlinear phase characteristics is the most difficult to solve. This is a very common problem but by no means trivial in some military FM applications where it is necessary to minimize intermodulation distortion to less than a few percent. The only way to alleviate this problem is to design wider BW filters, however for our application this reduces the weaker-signal suppression properties of the limiters and also prevents the filtering of spurious responses from the converter. Therefore, we see that the nonlinearties in the phase characteristics will have to be tolerated since their harmful effects will be less than the effect of increasing the bandwidth of the suppressor and converter filters.

REDESIGN OF FIXED TRAP CIRCUIT

Improvements to be Incorporated into Redesign

At the outset, one of the objectives of this thesis was to design a prototype of the fixed trap circuit that stressed simplicity both in design and in the number of components used. In lieu of the results of Chapter III and the further improvements required of the circuit listed below, this objective was not fulfilled. However, another more important objective was to determine the feasibility of the scheme depicted in Fig. 2.4; this the author felt was achieved.

Before embarking on the actual design, it is necessary to list the necessary improvements to be made in the redesigned version.

1 - A bandpass trap circuit will be designed whose null is produced passively so that it will not be subject to tube parameters and stage gains which the differential amplifier trap circuit of Sheftman's and Rubissow's circuits depended heavily upon. Furthermore, flexibility will be incorporated into the design of this circuit to provide for separate bandwidth control, trap positioning control, and trap attenuation control. An attempt will be made to make these controls independent of one another for quick and easy changes. Most of these controls can be dispensed with after the feasibility has been demonstrated.

2 - We will make provisions to minimize the effects of incidental envelope amplitude modulation resulting from the filter amplitude characteristics by broadening the trap BW. Special

limiter and filter stages will also be incorporated into the design to minimize the generation of this spurious AM.

3 - Delay equalizers will be incorporated into the design to compensate for the spurious delays encountered by the signals, (considered in next chapter).

4 - The problem of oscillator stability will be alleviated by using crystals.

5 - Precautions will be taken to minimize the effects of oscillator leakage and radiation.

6 - Redesign of the mixer units will be undertaken so as to minimize spurious mixer responses, fluctuation noise, and stray coupling via the mixers.

7 - Careful layout and sheilding will be stressed in order to prevent parasitic oscillation and to minimize stray coupling between the different stages.

Design of a Bandpass Filter with a Tuned Trap at the Center of the Band

Theoretically, all we require of our trap is to attenuate a constant frequency signal. However, as we have seen in the previous chapter, it would be beneficial to provide attenuation over a small band so as to minimize the effect of any incidental AM sidebands. This would require widening the trap bandwidth while still maintaining a sufficient degree of attenuation at the center frequency. Although it would be desirable to widen the trap in the stop band, this must be achieved with a minimum amount of pass band disfiguration in order to restrict amplitude modulation of the FM component to the stop band of the trap only. We have seen from the discussion in the second chapter that a wider trap bandwidth exerts its greatest influence on the capture of the weaker signal in the capture transition region centered near $\underline{a} = \delta$. In order to improve the capture transition, a narrow BW for the trap is required. However, we also require a trap BW that would minimize FM transient distortion. In Chapter III we stated conditions based on quasi-stationary requirements derived by Baghdady for the minimum BW required of a filter when excited by an FM signal. It should be recalled that the FM signal sweeping the trap is frequency modulated in accordance with the algebraic sum of the stronger and weaker signals. If $\theta_T(t)$ represents the frequency modulation and if we assume that both weaker and stronger signals are sinusoidally modulated then

$$\theta_T(t) = \frac{\Delta\Omega_1}{\omega_1} \sin \omega_1 t - \frac{\Delta\Omega_2}{\omega_2} \sin \omega_2 t \qquad 4.1$$

from which we obtain straight forwardly

$$\left|\theta_T''(t)\right|_{max.} = \omega_1 \Delta\Omega_1 + \omega_2 \Delta\Omega_2 \qquad 4.2$$

If we assume that both weaker and stronger signals are both fully deviated to $\pm$ 75 kc at their maximum rates, 3 kc (3 kc is the highest sweep rate used in the lab experiments that follow) then

$$\left|\theta_T''(t)\right|_{max.} = 2\,\omega_{max.}\,\Delta\Omega_{max} \qquad 4.3$$

where

$$\omega_{max} = (2\pi)\, 3\ kc$$

$$\Delta\Omega_{max} = (2\pi)\, 75\ kc$$

Thus we have for the minimum BW requirement

$$BW_{min.} = \sqrt{5 \cdot k_n \cdot 2 \cdot \Delta\Omega_m\, \omega_m}$$

If we assume that the trap attenuation characteristic can be approximated by a single tuned circuit (an optimistic estimate) then consulting Baghdady's analysis (Pg. 24 Ref.No.5) we have $k_n = k_1 = 8$. Substituting this into Eq. 3.10 we have for the minimum trap BW

$$BW_{min.} > \sqrt{5.8.2.75.3} \quad kc \tag{4.4}$$

$$BW_{min} > 134 \; kc$$

This BW is much too large since we are only contemplating a 300 kc BW for the overall pass band of the filter. Thus it appears that FM transient distortion will have to be tolerated. Sheftman used a 3db BW of approximately 30 kc and he shows oscillograms that do not indicate excessive transient distortion (Pg. 57 Ref. No. 4). It appears from his oscillograms that when the FM deviation was reduced, the distortion became more excessive. The reason for this is that with smaller deviations a greater portion of the FM signal falls within the stop band of the trap. The trap attenuation was sufficient to prevent the demodulator limiters from saturating during this portion of the modulating cycle thus giving rise to noise bursts of longer relative duration. Based on Rubissow's and Sheftman's results, a 3db BW, for the trap, between 30 kc and 60 kc would minimize this form of distortion for FM signals whose frequency deviations exceed $\pm$ 35 kc while still being wide enough for an acceptable degree of transient distortion.

The second trap parameter of importance is the center frequency attenuation. The main requirement imposed on this parameter is that the attenuation at the center frequency be sufficient so as to permit capture of the weaker signal over a wide range of amplitude interference ratios. From the discussion in Chapter II, we note that the location of the stronger to weaker signal capture transition region is centered approximately about $\underline{a} = \delta$, where δ is the trap center frequency attenuation. It is desirable to have this region below amplitude interference ratios of $\underline{a} = 0.01$. Thus to be on the safe side, the design will strive for a center frequency attenuation of 60db ($\delta = 0.001$).

The third parameter associated with the trap filter shown in Fig. 2.3 is the overall BW of the pass band portion of this filter. If deviations of $\pm$ 75 kc are used for both FM signals, the full deviation of the FM signal that sweeps the trap will require a 300 kc BW since the modulation of this FM signal consists of the algebraic sum of both the stronger and weaker signal modulations.

We thus have established the three bandpass trap filter parameters. In summarizing, we have:

a - 3db attenuation of trap - 30 kc to 60 kc

b - center frequency attenuation of the trap $\delta = 0.01$

c - overall BW of the pass band portion of the filter at 3db pts $\doteq$ 300 kc

The next consideration given to the trap design was to investigate some of the possible methods for achieving the desired trap attenuation characteristics shown in Fig. 2.3. A few considered were:

1 - A passive crystal bandpass notch filter

2 - A bandpass notch filter utilizing only lumped passive elements

3 - A feedback amplifier employing a rejection circuit

Possible Trap Design Utilizing Crystals

The first possibility, that of designing a crystal bandpass notch filter, was quickly abandoned because of the complexities and costs involved. First of all, it is not a problem easily solved in the laboratory since it requires special knowledge, skills and tools. The possibility of having the filter designed commercially was investigated; however, the cost was excessive and too much time was required for delivery. In connection with using crystals, the possibility of using two single sideband bandpass crystal filters in the manner shown in Fig. 4.1 was considered.

resultant when both are cascaded

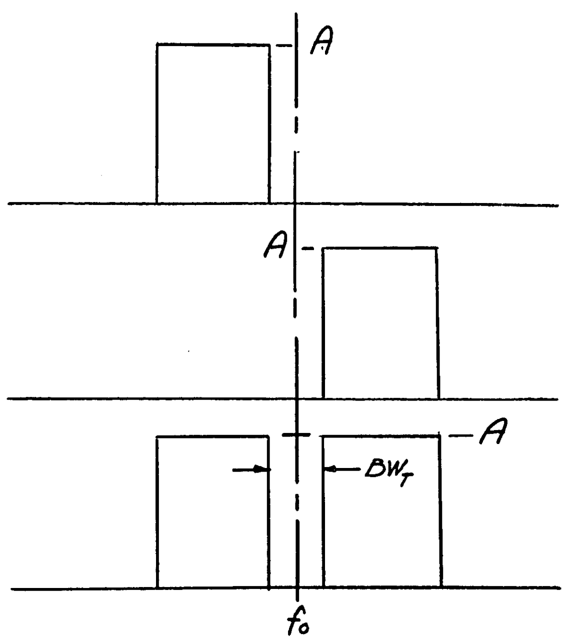


Fig. 4.1 Possible Method of Obtaining Bandpass Trap Circuit with Single Sideband Crystal Filters

One of the difficulties encountered by this scheme is that the spurious resonant response out of band of each single side-band crystal filter will influence the response of the other half. This is especially true of the effect of the spurious, high-frequency, out-of-band, resonant, response of the lower frequency, single-sideband, bandpass crystal filter. This requires further design considerations thus adding more to the cost.

Another difficulty that would inevitably be encountered by designing a trap filter in this manner is the excessive FM transient distortion that would prevail. The lumped parameter, LC equivalent of a crystal filter is usually a very complicated lattice network involving many poles and high energy (high Q) storage elements. From Baghdady's quasi-static analysis (Ref. No. 4) we note that the greater the number of poles that a filter has, the more difficult it is to meet the conflicting bandwidth requirements imposed on this filter. Previously we calculated a minimum bandwidth requirement for the trap assuming a series resonant circuit as an approximation to the trap response. However, single tuned circuits are the least sluggish of any filter that could provide the trap response required and the minimum bandwidth calculation still resulted in an exorbitant bandwidth requirement. Thus we see that the crystal filter, with its complex equivalent network, would require even more of a minimum bandwidth.

A second scheme was considered which seemed simpler; however, the cost in having the filter made was still exorbitant. This scheme employs the cascading of a simple LC bandpass filter with a crystal notch filter. If time permitted, this scheme would have been used in the thesis for the sake of stability (but at the expense of control flexibility.) Crystal filters for this application also have a bandwidth handicap. It is very difficult to achieve a BW greater than 5 or 6 kc at the 3db attenuation point. This trap BW would be rather narrow and the FM transient distortion would be excessive. However, using crystals, it is no problem to achieve 60db attenuation at a BW of roughly 1/4 the 3db bandwidth (the ratio of 3db BW to 60db BW in crystal notch filters is usually of the order of 3 to 4). Therefore, for future applications of the fixed trap circuit, it is felt that specially designed crystals would be the ideal trap circuit since the attenuation characteristics would be excellent and the drift stability problem nonexistent. The only disadvantage would be FM transient distortion; however, this could be circumvented by using narrow-band FM signals.

<u>Passive LC Trap Circuit</u>

The second method considered, obtaining a trap circuit entirely from passive elements, was not too successful. This method depended upon cascading a bandpass filter with a narrow notch filter to produce the overall trap characteristic. The greatest difficulty with this scheme was achieving sufficient attenuation together with the proper notch BW. Fig. 4.2 shows the network that was designed in an attempt to produce the required trap characteristics. The best attenuation that was achieved for this circuit was 30db at a center frequency of 1 mc with a 3db BW in excess of 60 kc.

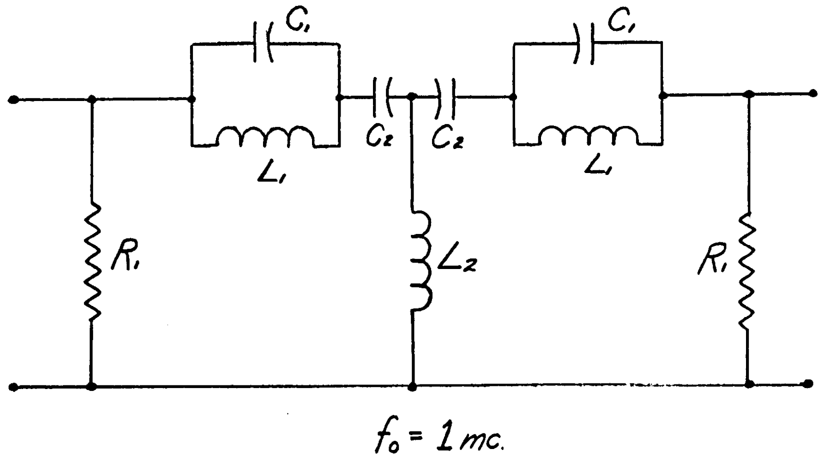


Fig. 4.2-Passive Small Percentage Band-Reject Circuit

The greatest difficulty encountered with this network was realizing the necessary Q's for the inductors. At high frequencies, suitable core material was not available for achieving Q's in excess of 150. These filters also require a high degree of shielding which tends to lower the Q of the inductors. One solution to this problem of designing a passive trap circuit is to go to a lower center frequency (below 1 mc). At these frequencies, high quality core material is available which can meet the required Q demands.

Feedback Amplifier Employing a Rejection Circuit

An amplifier utilizing a rejection circuit in a negative feedback loop is a common method for improving the selectivity of low frequency amplifiers. Fig. 4.3 depicts the arrangement commonly used, and also shows the selectivety curves that result using the rejection circuit in the feedback loop. The negative feedback rejection circuit used in conjunction with the amplifier

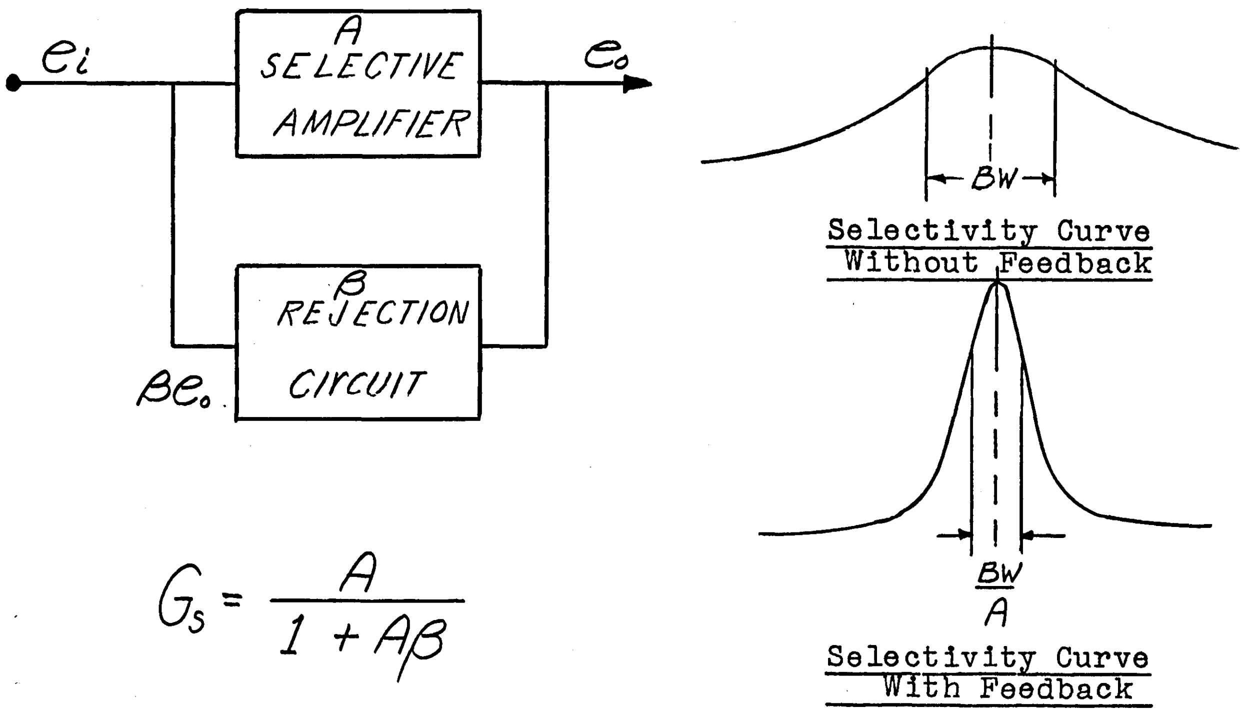


Fig. 4.3 Selective Amplifier Utilizing a Rejection Circuit in a Negative Feedback Loop

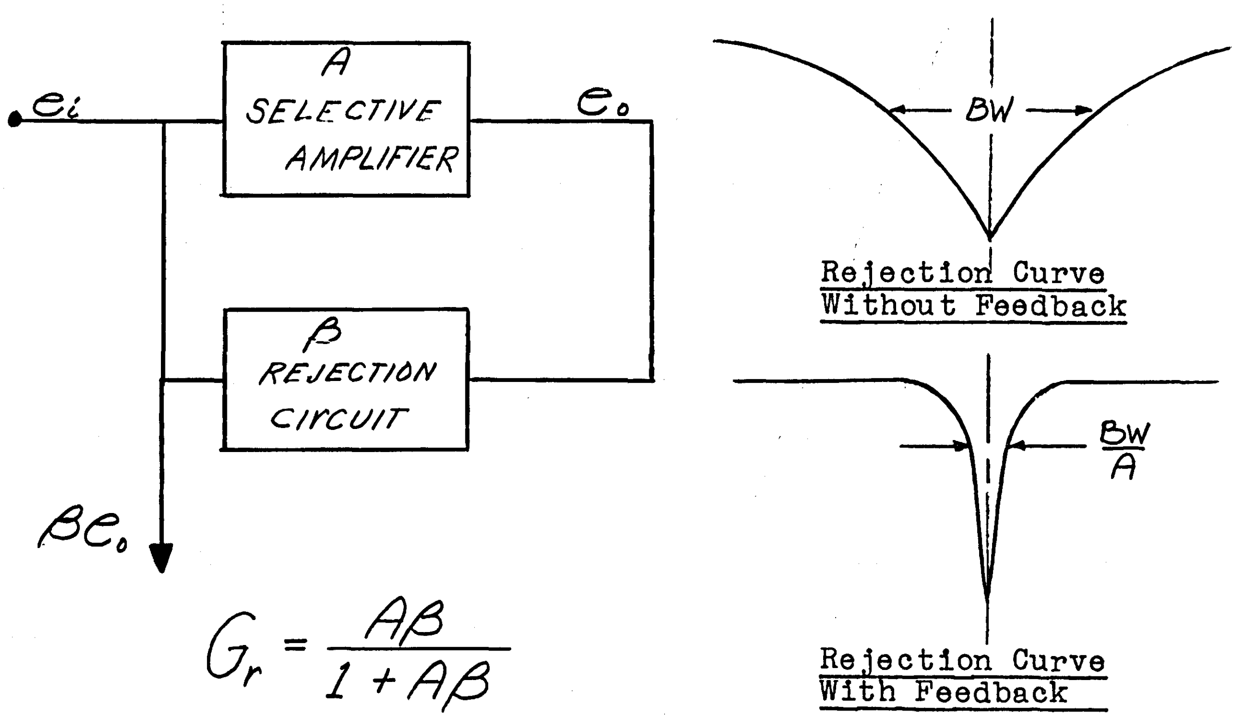


Fig. 4.4 Rejection Amplifier Utilizing Negative Feedback

is essentially a form of Q multiplication. For this arrangement, the overall gain of the system is given by

$$G_s = \frac{A}{1 + A\beta}$$

where A = gain of amplifier without feedback

β = transfer function of rejection circuit

If instead of taking the output from the selective amplifier, output terminals are placed in the output of the rejection circuit, the resulting configuration results in a sharply tuned rejection circuit, as shown in Fig. 4.4. The overall gain for such a system is given by

$$G_r \frac{A\beta}{1+A\beta}$$

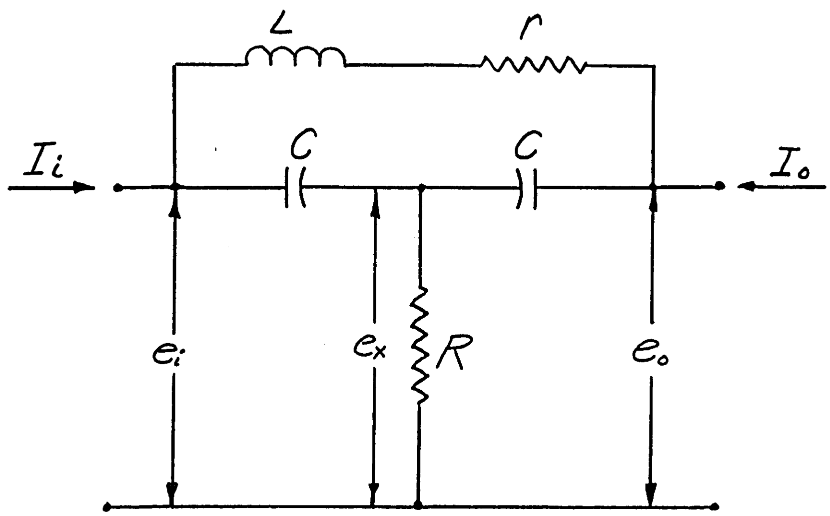
It is important to note from this equation that the null at the center frequency is solely determined by the transfer function of the rejection circuit, because if β goes to zero, G_r goes to zero also. However, the shape of the attenuation characteristic is made more selective by use of feedback. Essentially this is what is desired in the trap circuit, to have the null produced passively but to have some control over the shape of the characteristic.

Both Rubissow and Sheftman discounted the possibility of using feedback for fear of oscillation at the RF frequencies employed. In addition, most of the literature available on the type of commonly used rejection circuits is confined to low frequency applications. However, the response that can potentially be achieved is excellent and on the basis of this it was decided to try it at high frequencies. Fortunately, most of the problems that were anticipated were not as insurmountable as originally believed and the circuit to be presented later based on the principle of negative feedback utilizing a rejection circuit worked exceptionally well.

Analysis of a Bridged-T Rejection Circuit (13)

Three very common rejection circuits used in the application described previously are the twin-T, the bridged-T and the Wien-bridge. All of these circuits exhibit similar amplitude vs. frequency transfer characteristics. The twin-T involves six elements in which pairs of elements must be varied synchronously to change transfer characteristics. The bridged-T requires only four elements, two of which can be made independently variable to change the transfer characteristics. Thus, it was decided that the bridged-T is best suited for the application intended. The Wien-bridge rejection circuit requires only four elements, two of which can be made independently variable. However, the Wien-bridge is balanced with respect to ground making it vulnerable to stray capacitive effects especially for the high frequency application intended. Both the bridged-T and the twin-T are unbalanced with respect to ground, making them both applicable.

Fig. 4.5 shows the bridged-T in the form that it was used in the receiver trap circuit. The input drive is a voltage source. We are interested in the voltage transfer function, β, for the case of no loading in the output.



$$\beta = \frac{e_o}{e_i}$$

no load assumption

$I_o = 0$

<u>Fig. 4.5-Bridged-T Rejection Circuit</u>

Because of the circuit configuration, it is more expedient to use determinant theory in finding a relation for the voltage transfer ratio. Using nodal analysis, the admittance matrix is given by

$$\begin{bmatrix} I_i \\ 0 \\ 0 \end{bmatrix} = \begin{bmatrix} sC + \frac{1}{r+sL} & -sC & -\frac{1}{r+sL} \\ -sC & \frac{1}{R} + s2C & -sC \\ -\frac{1}{r+sL} & -sC & sC + \frac{1}{r+sL} \end{bmatrix} \begin{bmatrix} e_i \\ e_x \\ e_o \end{bmatrix} \qquad 4.5$$

solving for e_o and e_i we have

$$e_o = \frac{\Delta_{13}}{\Delta} I_i \qquad e_i = \frac{\Delta_{11}}{\Delta} I_i \tag{4.6}$$

$$\beta = \frac{e_o}{e_i} = \frac{\Delta_{13}}{\Delta_{11}} \tag{4.7}$$

Δ_{11}, Δ_{13} are the appropriate co-factors of the admittance matrix.

The null conditions are obtained by letting the output voltage go identically to zero while the input voltage remains finite, or that $\Delta_{13} \equiv 0$ while $\Delta_{11} \neq 0$.

$$\Delta_{13} = \begin{vmatrix} -sC & \frac{1}{R} + 2sC \\ -\frac{1}{r+sL} & -sC \end{vmatrix} \equiv 0 \tag{4.8}$$

Expanding Δ_{13}, and substituting for s, $j\omega$, we have

$$\Delta_{13} = \frac{1 - \omega^2 C^2 Rr + j(2RC\omega - \omega^3 LC^2 R}{R(r + j\omega L)} \tag{4.9}$$

Equating reals to zero and imaginaries to zero (since we have an identity) we have two conditions that must be satisfied in order to produce a null

$$\omega_0^2 = \frac{1}{rRC^2} \quad , \quad \omega_0^2 = \frac{2}{LC} \tag{4.10}$$

The Q of the coil at ω_0 is more convenient to use as a parameter

$$Q_0 = \frac{\omega_0 L}{r} \tag{4.11}$$

substituting in Q_0, the two null conditions become

$$Q_0 = \omega_0 R 2C \ , \ \omega_0^2 = \frac{2}{LC} \tag{4.12}$$

We see that resonance is determined by the bridged inductance and the series combination of the two capacitances.

The co-factor Δ_{11} is given by

$$\Delta_{11} \quad \begin{vmatrix} \frac{1}{R} + 2sC & -sC \\ -sC & sC + \frac{1}{r+sL} \end{vmatrix}$$

Expanding Δ_{11} we have

$$\Delta_{11} = \frac{1 - \omega^2 RCL - \omega^2 RrC^2 + J(\omega rC + \omega RrC - \omega^3 LC^2)}{R(r + J\omega L)} \qquad 4.13$$

Forming the ratio $\beta = \frac{\Delta_{13}}{\Delta_{11}}$ from Eqs. 4.9 and 4.13 we have

$$\beta = \frac{(1 - \omega^2 C^2 Rr) + J(2RC\omega - RLC^2\omega^3)}{1 - \omega^2 RCL - \omega^2 Rrc + J(\omega rc + \omega Rrc - \omega^2 LC^2)} \qquad 4.14$$

Making use of high Q approximations and substituting $\rho = \frac{\omega}{\omega_0}$ the expression for β can be manipulated into the form

$$\beta = \frac{1}{1 - J\frac{\rho}{\rho^2 - 1}\frac{2}{Q}} \qquad 4.15$$

if we let $2\delta = \rho - \frac{1}{\rho}$

then $\frac{\rho^2}{\rho - 1} = \frac{1}{2\delta}$

where $\delta = \frac{\Delta\omega}{\omega}$, the fractional frequency deviation from center frequency. Substituting these relations into Eq. 4.15 we have for β

$$\beta = \frac{1}{1 - J\frac{1}{Q\delta}} \qquad 4.16$$

The overall gain for the feedback rejection amplifier shown in Fig. 4.4 is given by

$$G_r = \frac{A\beta}{1 + A\beta} \qquad 4.17$$

The open loop gain of the amplifier is A. We will assume that this amplifier is a wideband (BW $\approx$ 300 kc) single tuned amplifier whose gain may be expressed by

$$A = \frac{A_0}{1 + J2Q_s\delta} \qquad 4.18$$

A_o = midband gain of amplifier

Q_s = Q of single tuned circuit

δ = fractional deviation from center frequency,

Substituting Eqs. 4.16 and 4.18 into Eq. 4.17 we have for the overall gain of the rejection amplifier

$$G_r = \frac{\dfrac{A_o}{1+J2Q_s\delta} \cdot \dfrac{1}{1-J\frac{1}{Q_o\delta}}}{1 + \dfrac{A_o}{1+J2Q_s\delta} \cdot \dfrac{1}{1-J\frac{1}{Q_o\delta}}}$$

$$G_r = \frac{A_o}{A_o + (1+J2Q_s\delta)(1-J\frac{1}{Q_o\delta})} \qquad 4.19$$

Further simplifications can be made by assuming $A_o > 10$, and by assuming the single tuned circuit to be wide band. This justifies the assumption that $2\ Q_s\delta \ll 1$. With these assumptions the overall gain of the rejection circuit reduces to

$$G_r = \frac{1}{1 - J\frac{1}{A_oQ_o\delta}} \qquad 4.20$$

We note from this equation that at resonance, $\delta = 0$ and $G_r = 0$. At frequencies sufficiently displaced from the center frequency $Q_o\ A_o\delta$ becomes much greater than unity and the gain of the rejection circuit, G_r, approaches unity. By using feedback, we have also accomplished Q multiplication, since the effective Q of the rejection circuit is now given by

$$Q' = A_oQ_o$$

Thus, a very sharp attenuation characteristic can be achieved by using simple RF coils. A typical value for a well shielded coil is Q = 100. If we assume A_o = 10, we effectively have a trap Q = 1000.

Fig. 4.6 shows some curves plotted for the response that theoretically can be achieved. The bandwidth of the circuit as shown on the curves is 9 kc. Theoretically, at the center frequency we have a complete null. Curve B depicts the response when the out-of-band response of the amplifier is taken into account. As expected, the trap attenuation response is not effected. The only effect of the single tuned circuit is that the response out of the trap band never reaches its maximum of unity. At the extreme edges of the response curve, the single tuned circuit exerts its greatest effect and the response starts to diminish. However, in order to improve the skirt selectively at the edges of the band a double-tuned bandpass amplifier will be cascaded with the trap circuit to produce the desired filter characteristic as shown in Fig. 2.3.

Practical Design of a Bandpass Trap Filter

Since the trap will immediately follow the first mixer, the bandpass filter in the output of this mixer will provide the overall pass band shape of the trap response.

In the design of the rejection amplifier, numerous considerations had to be taken into account because of the high frequency application. The first problem was to provide an arrangement for isolating input and output terminals. The basic block diagram of the rejection circuit, Fig. 4.4, shows that the output of the rejection circuit is tied directly to the input terminals. Obviously,

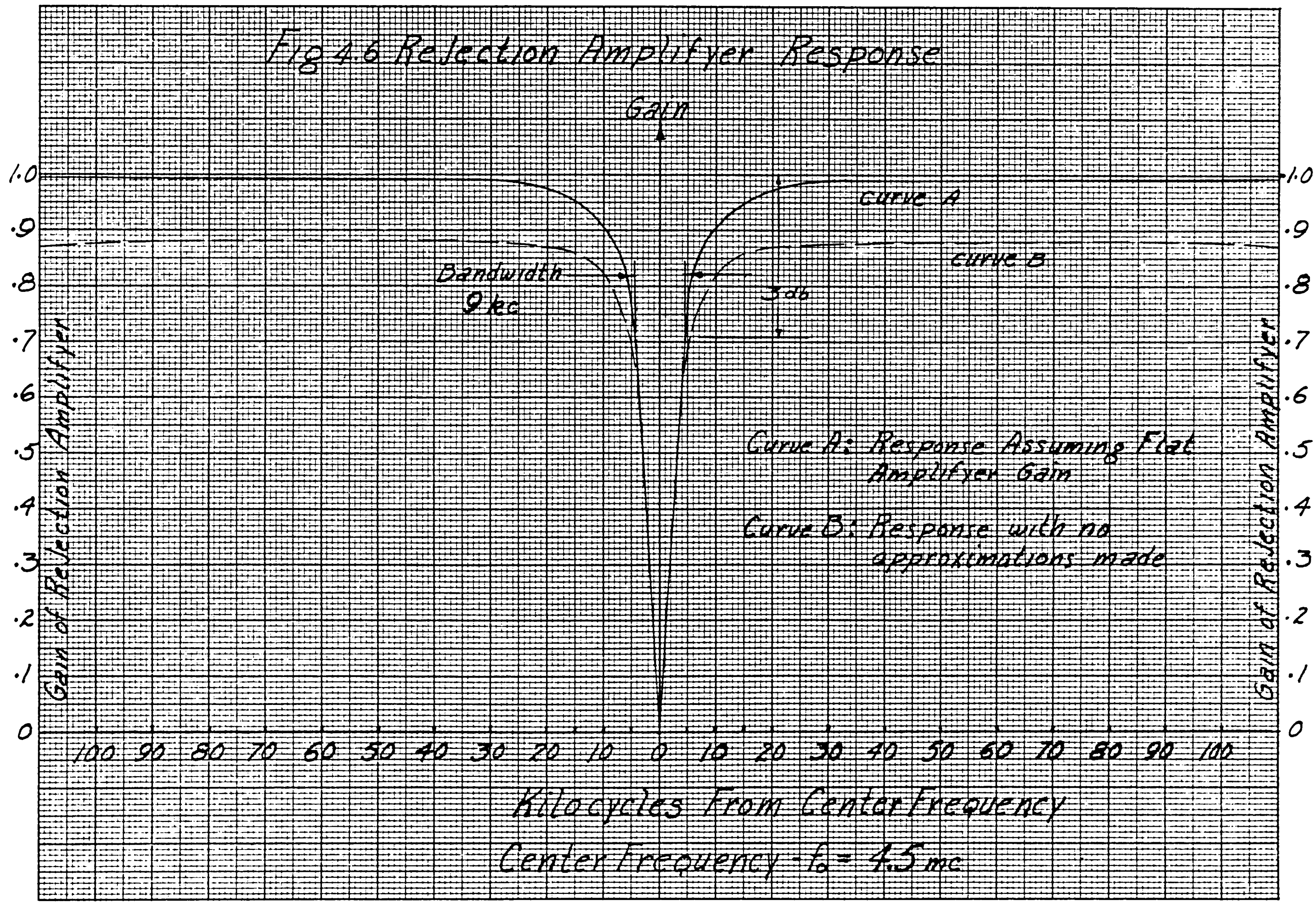


Fig 4.6 Rejection Amplifyer Response

some circuit arrangement is necessary that will provide direct electrical coupling of output to input without short circuiting the input terminals to the output terminals. This suggests driving the amplifier from the cathode and looping back through the plate circuit impedance and bridged-T network to the control grid of the amplifier. This amplifier control grid would also serve as the output terminal for the entire rejection circuit. We thus achieve in a practical manner through the electronic coupling between control grid and cathode, that which is shown ideally in the rejection circuit block diagram (see Fig. 4.4).

Since the input impedance of a cathode driven amplifier is very low; we require some means of impedance matching if we intend to drive from the output of the mixer bandpass filter. This can be best accomplished by using a cathode coupled amplifier arrangement which amounts to inserting a cathode follower between the mixer filter and the cathode driven amplifier.

The next problem in using the bridged-T arises mainly because of the high frequency application. We recall that the analysis and resulting gain equation (Eq. 4.20) is valid only for the conditions of no loading in the output of the bridged-T. Special precautions are necessary then for the high frequency application. Feeding the output of the bridged-T into a grid of a pentode and also taking the output from this point and feeding into the grid of a pentode stage that follows, loads the bridged-T capacitively. This loading capacitance, including wiring strays, can easily amount to as much as 20 mmfd. At 4.5 mc, 20 mmfd of capacity is only 1800 ohms which is too low an impedance to justify a no loading assumption. This loading capacity may be reduced to approximately 1 mmfd of capacity (35,000 ohms) but only at the expense of reduced feedback and circuit complexity. Reduced negative feedback would not effect the null condition since this is produced passively but it will effect the shape of the trap characteristic. We could compensate for this by using a higher gain selective amplifier in the feedback loop. This requires a higher plate load impedance in the amplifier while still maintaining the required broad band response. (This amounts to using a coil of higher inductance since the Q is fixed by the bandwidth requirements.) In order to reduce the capacitive loading, the output of the bridged-T was fed through a 1 mmfd capacitor into a cathode follower. The output of this second cathode follower is fed back to the control grid of the selective amplifier thus completing the feedback loop. Another terminal is tied to this second cathode follower which serves as the output terminal of the rejection amplifier. It was found necessary at this point to include a stage of gain because of the attenuation produced by the 1 mmfd capacitor. Thus the second output terminal from this cathode follower feeds directly into a single tuned amplifier forming the last stage of the trap circuit. The tuning of this last single tuned amplifier was helpful to the alignment since it helped compensate for any dissymetry in the pass band of the bandpass trap filter. It has no effect on the notch, however.

The next problem encountered in the design of the bridged-T was that the driving point impedance of the bridged-T, as seen by the plate load impedance of the selective amplifier, was not constant. Assuming that the output terminals of the bridge-T are open, the driving point impedance is given by

$$Z_{in} = \frac{S^3RC + S^2\left(1+\frac{rRC}{L}\right) + S\left(\frac{r}{L}+\frac{2R}{L}\right) + \frac{1}{LC}}{CS\left(S^2 + S\frac{r}{L} + \frac{2}{LC}\right)} \qquad 4.21$$

Using the two null conditions of the circuit, this equation may be put in terms of Q_0 and ω_0. Making some high Q approximations, the input impedance may be manipulated into a form valid for small frequency changes (δ<1/10).

$$Z_{in} = 2R\left[\frac{(1+3\delta) + JQ_0\delta}{1+2\delta + J2Q_0\delta}\right] \qquad 4.22$$

where R = resistance to ground of center shunt arm of bridged-T

Q_0 = Q of coil

$\delta = \frac{\Delta\omega}{\omega_0}$ fractional frequency deviation

At resonance, Z_{in} is a maximum = 2R. Off resonance Z_{in} decreases monotonically to R at the extreme edges of the band. In cascading the bridged-T to the selective amplifier it would be desirable to have a plate load impedance of roughly one fifth to one tenth that of R; otherwise, the gain of the amplifier would vary with frequency in accordance with the combined impedance of plate load and input impedance of the bridged-T. However, it is preferrable to keep R as low as possible in order to minimize the effect of stray capacity from the center node of the bridged-T to ground (in shunt with R). This last effect is just as detrimental as the variations of gain would be if a low R were used so a compromise was made in the choice of plate load impedance and bridged-T shunt resistance R. The values chosen in the final design were $-R \approx 6$ kilohms, and the resonant impedance of the tuned circuit was approximately 2 kilohms.

The last consideration given to the design of the trap was that of providing flexible control. As mentioned, the mixer band-pass filter determines the shape of the overall passband of the trap. This filter was overcoupled producing peaks at the extreme edge of the band thus compensating for the rounding effect of the single tuned circuits employed later. The location of the null was made variable by utilizing a variable inductance in the bridged-arm of the bridged-T circuit. A fine control of the null location was also provided by padding the capacitance arm with a synchon-ously tuneable butterfly condenser. The null attenuation was made variable by using a rheostat in place of the fixed R in the shunt arm of the bridged-T. Actually two rheostats were used in series, the second, lower-resistance rheostat provided fine control of the null attenuation. The trap bandwidth was also made variable by placing a variable capacitor pad across the tuned circuit in the selective amplifier that drives the bridged-T. In order to pro-vide bandwidth control, in this manner, it was necessary to in-crease the selectivity of this amplifier slightly since it was originally too broad. Varying the tuning of this amplifier effects the gain which has a direct bearing on the bandwidth of the trap. However, varying the bandwidth in this manner causes an unsymetrical

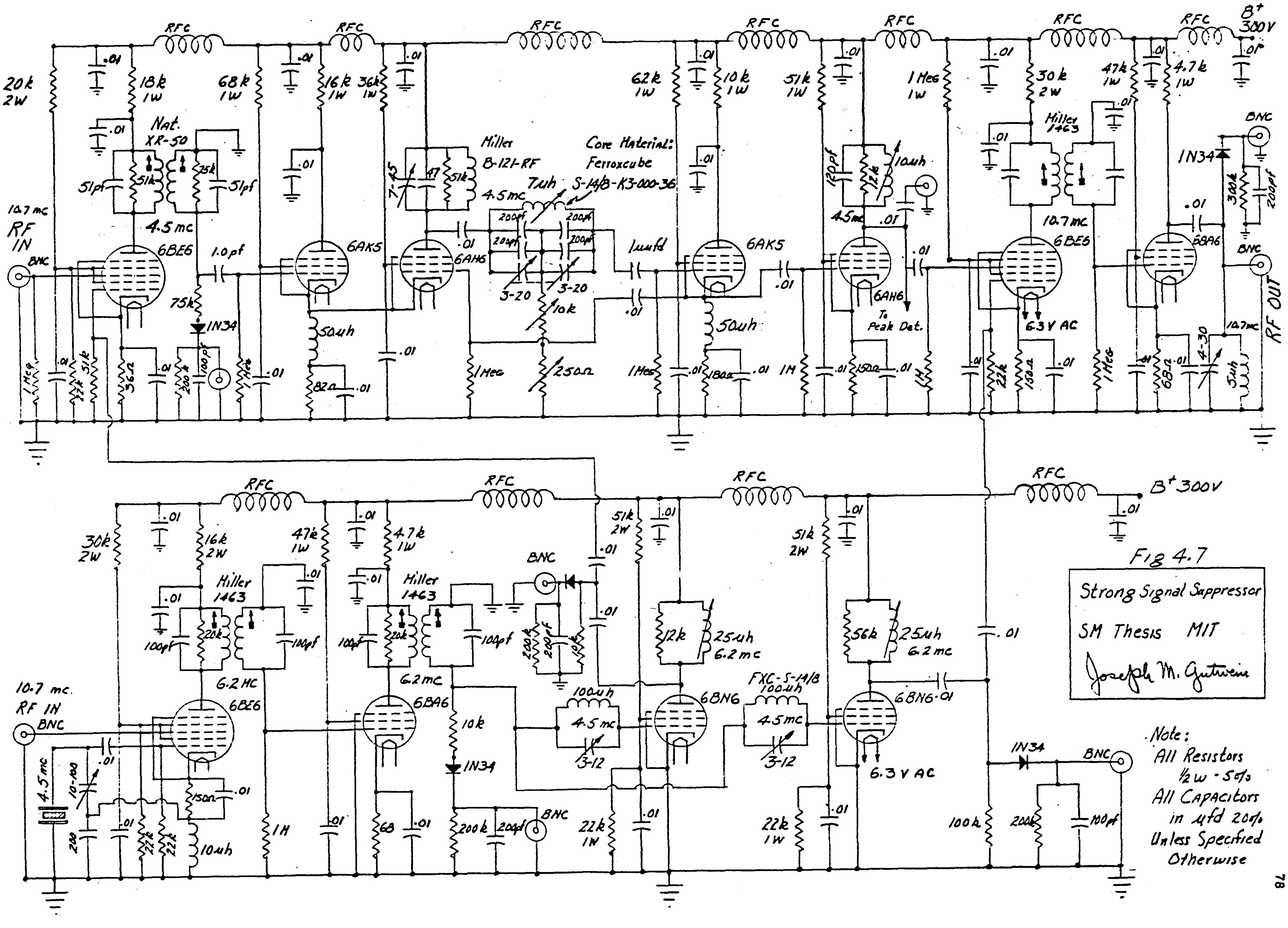


Fig 4.7

Strong Signal Suppressor
SM Thesis MIT
Joseph M. Gutwein

Note:
All Resistors ½w - 5%
All Capacitors in μfd 20%
Unless Specified Otherwise

pass band which is compensated by tuning the very last single tuned stage.

As might be expected, all these controls are somewhat interdependent when large changes are made. However, for small changes, each control is approximately independent of the other, making for extreme ease in alignment. As might also be expected, mistuning of the circuit will render it oscillatory. This is especially true of the shunt R if in the event that this resistance is reduced to the point of short circuiting the center node to ground. A second oscillatory trend was noticed after the selectivity of the amplifier was increased to provide bandwidth control of the trap. Tuning this amplifier to exactly the center frequency also causes the circuit to oscillate, but when detuned slightly, its only effect is as described previously. Other than these two defects, the circuit showed no other oscillatory trends and gave no trouble (other than drift and tube ageing) for the duration of the thesis experiments.

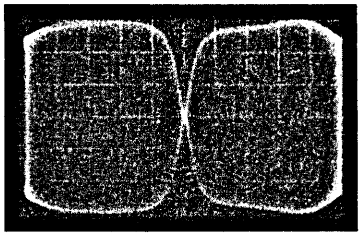
The upper half of Fig. 4.7 shows the redesigned circuit of the first mixer and the trap filter employing the rejection amplifier. The rest of the fixed trap circuitry will be explained later. In the layout and the construction of the trap circuit, precautions were taken to eliminate stray coupling between the selective-amplifier-tuned-circuit and the bridged-T inductor. The selective amplifier coil is a low Q, commercial, RF coil. The bridged-T inductor is made from a high quality pot core which is self shielding. The coil is wound on a small bobbin and is sealed between two halves of a pot core assembly. Variable tuning is achieved by a small slug whose position effects the air gap between the two core halves, thereby producing the required inductance changes.

Also in the trap design, 6AH6 pentodes were used for their high gain properties. For the cathode followers, 6AK5 pentodes were used, mostly for their low input and cathode capacities. In the design of the cathode followers employing these 6AK5 tubes, coils were used to provide the AC cathode impedance. These coils are approximately self resonant with the surrounding stray capacities.

Fig. 4.8 shows oscillograms of the overall trap response for various trap bandwidths. One of the disadvantages of this circuit is the slight dissymetry of the trap; however, this effect is by no means serious. This was especially noticeable when the trap BW was made very narrow. In the course of conducting the tests, it was found that the optimum BW (3db) was approximately 60 kc and this was set and held fixed for the duration of the tests. For this BW the overall trap characteristic is approximately symetrical. It is not shown in these photographs, but the maximum BW obtainable by the simple adjustments described exceeded 100 kc.

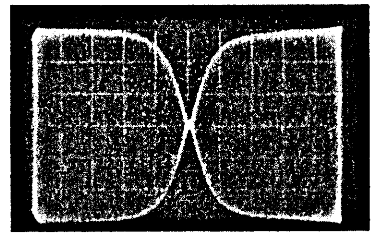
The null did not photograph too well in these oscillograms; however, when the sweep rate of the scope was expanded and the scope gain maximized, there appeared to be a perfect zero at the null. It was difficult to measure the actual attenuation at the null for the lack of high gain RF equipment; however, the attenua-

Pic. No. 1



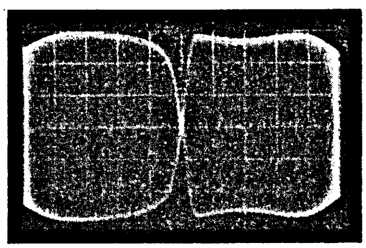
Hor. Axis Cal.
30 kc/cm
Trap BW_{3db} = 25 kc
Min. Trap BW
maintaining symetrical
pass band

Pic. No. 2



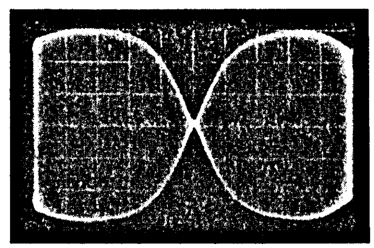
Hor. Axis Cal.
15 kc/cm
Trap BW_{3db} = 25 kc
only Hor. Scale
expanded to show
null

Pic. No. 3



Hor. Axis Cal.
30 kc/cm
Trap BW_{3db} = 15 kc
Min. Trap BW
attainable using
rejection circuit

Pic. No. 4



Hor. Axis Cal.
30 kc/cm
Trap BW_{3db} = 60 kc
Trap BW used
for testing circuit
performance

Center Frequency of Trap is 4.5 mc

Fig. 4.8 Amplitude vs. Frequency Response of Bandpass Fixed-Tuned Trap tion was sufficient to permit a thorough investigation of the response of the fixed trap receiver.

Design of Mixer Stages

In their normal application, mixers or converters are usually required to perform a simple linear translation of the spectrum of a signal to a new IF frequency. However, the hetrodyning operation of the mixers in the fixed trap circuit is unorthodox since it involves the simultaneous mixing of two, sometimes three, frequency modulated signals. In two of the mixing stages employed, the theory of the fixed trap requires undistorted shifting of the modulation from one FM signal at one input of the mixer to a second FM signal at the other input of the mixer, producing in the output, a third entirely different FM signal. These are stringent requirements imposed upon these mixers; however, the results of Sheftman's circuit do not indicate difficulties that would arise from malfunctioning of these mixer stages. With this in mind, the same basic mixer circuits will be employed with the inclusion of a few minor improvements.

In the design of the mixer stages, 6BE6 pentagrids were used to perform the mixing operation. One of the assumptions made in the analysis of the mixing properties of pentagrids is that the control grid (oscillator grid) is driven with a signal that is much larger than the signal appearing on the signal grid. Driving the control grid with a large signal and providing proper bias for this grid (usually a self bias grid charging circuit similar to that used in pentode limiters provides the bias) are conditions necessary for which a complete sweep is made of the signal-grid-to-plate vs. control-grid-transconductance characteristic. Maximum conversion gain is achieved in this manner providing for a maximum signal to harmonic interference ratio in the output. By making a judicious choice of input signal frequencies, the effect of harmonics and other spurious response can be further minimized after filtering. The frequencies chosen for this fixed trap circuit are as follows: input and output signal frequencies - 10.7mc; local oscillator and trap center frequencies - 4.5 mc. With this choice of frequencies the interstage frequency between the converter and two mixers is 6.2 mc. The only frequency harmonic that might create a problem is the second harmonic of the oscillator frequency, 9.0 mc, since it is only 1.7 mc from 10.7 mc. However, proper care in mixer-filter design and alignment could minimize this effect and the effect of other spurious responses.

A third consideration given to the design of the mixers was to minimize coupling between the signals on the control grid and the signal grid. Special care was given to this problem, especially in the first mixer, since the performance of this mixer stage has the greatest effect on the overall performance of the system. Signal coupling between the control and signal grid arises from direct intergrid capacity, both external and internal, and from space charge coupling. External coupling can be minimized by shielding the signal grid from the control grid as was done in the construction of this circuit. Nothing can be done about the internal intergrid capacity.

The space charge coupling is the source of greatest coupling between signal grids. Coupling of this nature creates many problems and is difficult to compensate for. According to theory, (12) this coupling is in the form of a control-grid-to-signal grid, unilateral, negative capacitance. Different schemes of suggested neutralization techniques were tried in an effort to neutralize this effect. One of these consisted of connecting the signal grid to the control grid through a small variable resistance in series with a small variable capacitance. By varying either one, or both of these parameters, and noting the response of the signal grid, the space charge coupling could be partially neutralized. The improvement noticed was insignificant, so the scheme was abandoned. The only other possible solution was to provide low impedance in the signal grid circuit to the frequencies of the signals in the control grid circuit. Thus, the tuned circuits in the signal grid are tuned to a different frequency than the frequency of the signals in the control grid. This also guided the choice of frequencies employed in the fixed trap circuit.

A fourth consideration given to the design of the mixers was the noise problem. Mixers are notoriously noisy and some effort was expended in an attempt to reduce the noise generated. To reduce the relative effect of this noise, the signal grids were driven with as large a signal as possible (one to two volts) but still keeping the level within the linearity limits of the mixer tube. Another safeguard taken was to determine experimentally the optimum screen and plate voltages necessary to provide a signal in the output with a minimum amount of added noise. Usually this resulted in a reduced value of screen voltage than that recommended in the tube manual. The reason for this is that the fluctuation noise is proportional to the DC plate current drawn by the tube, reducing the screen voltage reduces the plate current, hence the noise. Since gain was not of importance, and since it was desirable to minimize the noise, the screen voltage of the third mixer was reduced to a very low value by insertion of a 1 megohm resistor. This seemed to reduce the noise generated by this tube (which was excessive) to a tolerable limit.

Design of Converter

The function of the converter is to linearly translate the stronger signal to a new IF frequency, 6.2 mc. The same considerations given to the mixer design apply to the converter. In this design, the local oscillator was incorporated directly into the 6BE6 pentagrid thus avoiding to some extent the problem of radiation and leakage from signal leads that would come from an external oscillator. The entire converter unit was shielded properly so as to minimize the radiation effect even more. The converter oscillator (4.5 mc) employed a crystal thus providing a stable local oscialltor frequency. Incorporated into the design was means for controlling the oscillator amplitude by a capacitive voltage divider. This variable capacitive voltage divider provided a range of oscillator amplitude of approximately five to forty volts. It was found, experimentally, that fifteen volts was an optimum value.

Design of Intermediate Stages Between Converter and Mixers

Reference to the block diagram of the fixed trap shows that the output of the converter should feed directly to the mixers. However, Sheftman reported that this at one time rendered his circuit oscillatory. Instead of feeding directly into the mixers, he split the output of the converter and fed through two buffer amplifiers to each mixer. This apparently cleared up the trouble. The basic idea of using buffers was also used in this design, however with certain modifications.

There are three things we wish to accomplish with the converter-mixer interstage circuitry.

a - To clear up part of the incidental AM discussed in Chapter III.

b - To provide a good selective filter for the converter in order to minimize any spurious responses.

c - To minimize any oscillator signal that may be directly leaking through to the other stages.

From Chapter III we saw that the incidental AM appearing in the constant frequency signal in the output of the first mixer was caused by the filter amplitude characteristic between the converter and first mixer. It would be desirable then to drive the first mixer control grid from a wide-band single tuned circuit in order to minimize the incidental AM. Since we are driving the control grid of the first mixer with a large signal, the input impedance at this grid is very low and a wide-band single tuned circuit is not only desirable, but necessary. To minimize the AM even more, a stage of wide-band limiting was provided and the single tuned circuit placed in the plate circuit of this limiter. The limiter also reduced the AM caused by the fluctuation noise generated by the converter. The filter characteristic following this limiter driving the mixer is shown in Fig. 3.3. In order to ensure limiting, a stage of gain was provided between the limiter and the converter. Using double tuned circuits in the output of the converter and the output of this amplifier provided essentially a four pole bandpass filter following the converter. This filter provided for sharp skirt selectivity which we required in order to eliminate some of the spurious response from the converter.

In order to check any oscillator leakage, simple 4.5 mc antiresonant chokes were placed in series with each limiter grid lead. The effectiveness of these chokes was placed in evidence, when it was noticed that without these chokes some oscillator signal was appearing at the center frequency of the trap circuit. Tuning of the choke to 4.5 mc reduced the oscillator signal in these stages. Further evidence of the effectiveness of these chokes was noted when it was observed that with a 10.7 mc signal applied to the converter, with no signal applied to the first mixer, a 10.7 mc also appeared in the output of the third mixer. The reason for this was found to be a result of the 4.5 mc oscillator signal directly leaking through to the third mixer and mixing with the 6.2 mc signal from the converter to produce a 10.7 mc signal in the output. Tuning the choke in this channel to 4.5 mc minimized most of this trouble.

The last provision made in the design was to incorporate a buffer amplifier following the third mixer. This buffer amplifier was necessary in order that we may couple out of the entire fixed trap circuit at a relatively low impedance. It also provided means by which a good filter response was achieved in the last mixer output directly preceding this buffer amplifier.

Fig. 4.7 is the entire redesigned fixed trap circuit appropriately entitled <u>Strong Signal Suppressor.</u> In concluding this chapter, one further mention is the actual construction of the circuit. Each unit was separately shielded by aluminum partitioning; where possible and where magnetic shielding was required, commercial coils were used. The only exception is the limiter filters and the filter following the first mixer. A large degree of overcoupling was necessary in the first mixer filter requiring special design considerations not afforded by the commercial cans. It was also desirable to have as high an inductance as possible in the limiter filters in order to drive the mixer control grids with a large signal.

We might add, that most of the precautions taken in the new fixed trap circuit redesign represents considerable overdesign. Once the feasibility of the system is established, many simplifications can be incorporated into the design providing for a less complicated appearing receiver. As it stands now, the only justification for all the components is that the system works well.

DESIGN OF WEAKER-SIGNAL SUPPRESSOR, DEMODULATOR AND DELAY EQUALIZER

In this chapter, the design of the weak signal suppressors, FM demodulator, and the delay equalizer will be discussed. Both Baghdady and Granlund have presented most of the theory pertinent to the design of limiters and demodulators. (6), (7) The design of these circuits is based on these theories.

Design of Weaker-Signal Suppressors

In this weaker-signal receiver, narrow-band limiters will be employed to provide the preliminary function of suppressing the weaker signal. The discussion in Chapter II typifies their operation as intended for this application. The considerations given to these narrow-band limiters apply equally as well to the limiters employed in the demodulator unit.

Reliable operation of the limiters is necessary for good overall performance of the systems. Certain basic requirements must be fulfilled by these limiters. We demand, therefore, realization of the following:

1 - The limiters must have a very low limiting threshold in order to handle a wide range of amplitude-interference ratios. We are especially concerned with handling interference ratios approximating unity ($\underline{a} \approx 0.90$). Under these conditions the limiters must have a threshold below the lowest possible input signal level. If we assume that the amplitude of the stronger signal is E_s (the median signal level) and the amplitude of the weaker signal is $\underline{a}\,E_s$, then the amplitude modulated resultant of these two signals varies between $(1-\underline{a})E_s$ and $(1+\underline{a})E_s$. For an $\underline{a} = 0.90$ this range

becomes 0.10 E_s to 1.9 E_s. Of principal concern is to ensure that the limiter threshold is below .10 E_s to ensure this condition by providing sufficient gain in the IF amplifiers. However, we then have to concern ourselves with providing a wide, dynamic linear range at large signal strengths in these amplifiers since we do not want these amplifiers to limit. For example; consider a typical limiter threshold to be 2.0 volts. For an $\underline{a} = 0.9$, we would then have to provide a minimum signal strength of $(1-\underline{a})\ E_s = 2.5$ volts in order to be safely above the limiter threshold. For these conditions, $E_s = 25$ volts which would still be within the linear range of the output of the last amplifier assuming we had enough impedance to provide for 25 volts out. However, the resultant signal for an $\underline{a} = 0.9$ would swing to approximately 50 volts. We would then have to design a linear amplifier capable of providing a dynamic linear range of output from 2.5 to 50 volts. This is a requirement difficult to meet in the design of an amplifier. But, if we could lower the threshold of the limiter by a factor of 5 or even 10, the amplifiers would only have to provide a maximum signal of 10 volts, which poses no design problems.

2 - The limiters must be rapid acting and not fail as a result of abrupt, input-amplitude variations. It is well known that the presence of two co-channel signals causes rapid amplitude fluctuations of the resultant envelope. The limiters must be made insensitive to these rapid amplitude fluctuations; otherwise, they will fail to limit.

3 - The fluctuations in the output amplitude of the limiter

must be kept to a minimum over the largest possible range of input amplitudes. In other words, the limiter characteristic should be as flat as possible beyond the threshold.

In this narrow-band application, there are other requirements which are desirable but not absolutely necessary. These are as follows:

1 - The filters following the limiters should be provided with a well-defined pass band, a flat top, and steep skirts.

2 - For purposes of cascading, either to an IF amplifier or to another limiter, it is desirable that the limiters have both, a high impedance input and a low impedance output. Good IF filter response and good limiter filter response can best be achieved and easily maintained when the limiter is provided with a high impedance input so as not to load the preceding filter. A low impedance output provides flexibility in interchanging the limiter circuits if the need arises.

With these requirements in mind, three common limiter circuits were tested; a double diode limiter, a pentode limiter and the 6BN6 limiter. The arrangement used to test these limiter circuits is shown in Fig. 5.1. The limiter response of each of these circuits is shown in Fig. 5.2.

The diode limiter, although rapid acting, exhibited the poorest limiter characteristic. This characteristic could have been improved by biasing the diodes; however, this complicates the design. Another disadvantage is that to produce one stage of limiting, two amplifiers are necessary. The first amplifier provides the signal

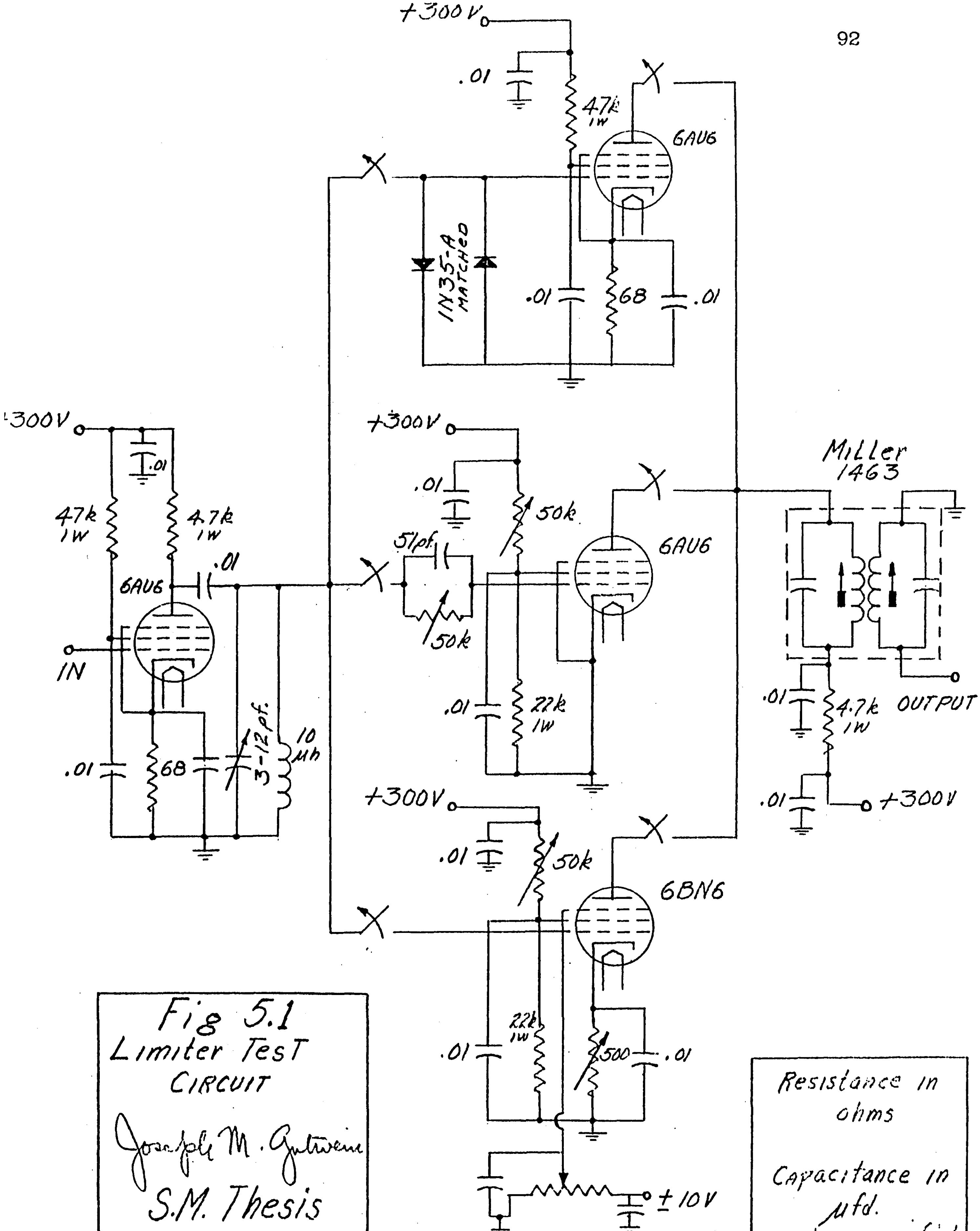


Fig 5.1
Limiter Test
Circuit

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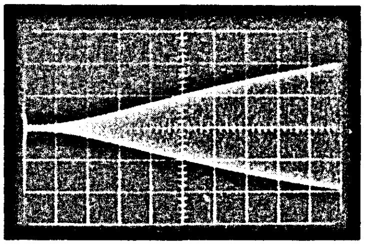
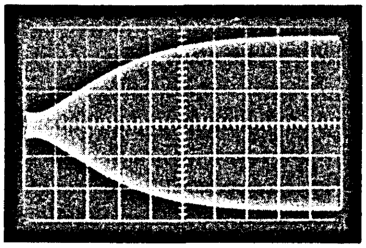
M.I.T.

Resistance in ohms

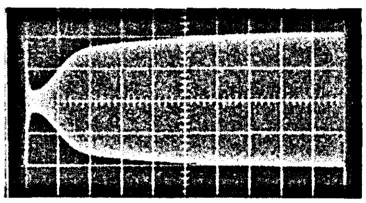
Capacitance in μfd. unless specified



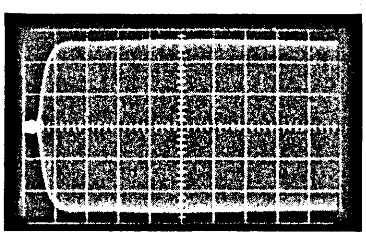
Sawtooth AM to Limiter
Vert. Axis Calibration
1 volt/cm



Diode Limiter Response
Hor. Axis Calibration
0.2 volt/cm



Pentode Limiter
Hor. Axis Calibration
0.2 volt/cm



6BN6 Limiter
Hor. Axis Calibration
0.2 volt/cm

Note: One amplifier precedes each limiter.
Schematic of Test circuit shown in Fig. 5.1

Fig. 5.2 Limiter Response of Various Limiter Circuits level to drive the diodes into cut off while the second amplifier, driven by the diodes, provides buffering between the limiting operation and the filtering.

The second limiter tested was the common pentode limiter. From the second oscillogram we see that the limiter characteristic for this type of limiter is much better than that achieved with the diodes. The pentode limiter depends upon a self-biasing arrangement to achieve its limiting properties. Unless special care is taken in the design to minimize the time constant of the grid-charging bias circuit, the limiter is not rapid acting. One advantage of this limiter though, is that it is not absolutely necessary to provide an interstage buffer. However, incorporating the rapid acting feature into the grid charging circuit has a direct bearing on the input impedance. Lowering the time constant of this grid bias circuit also lowers the input impedence. If an $\underline{a} = 0.90$ is to be handled, the grid bias circuit necessary provides an input impedence less than 10 kilohms, which in most applications is too low. This is an unfavorable feature of pentode limiters and presents difficulties in obtaining good response from filters immediately preceding the limiter.

The third limiter circuit tested was the 6BN6 limiter. As evident from the third oscillogram, the limiter characteristic is excellent. This limiter works on the plate-current-saturation principle when biased properly and unlike the diode or the pentode limiters, it experiences definite saturation. However, two of its disadvantages are the necessity of having to preset the biases, experimentally, and the variations of response exhibited from tube to tube. Nevertheless, where the emphasis is on performance, its excellent limiting properties warrant its use in laboratory designed equipment where these necessary accomodations can be made quite easily.

The 6BN6 limiter also has an advantage over the pentode limiter in that its input impedance is higher. It is difficult to measure this impedance but from experience gained in the lab, it was found to be approximately 15 to 20 kilohms which is not prohibitively low to cause excessive loading. One big disadvantage (as with the pentode limiter) is that this input impedance is non-linear with signal level. However, the range quoted is a lower limit and the loading was tolerable when used with low impedance (approximately 15 kilohms) commercial coils. This input impedance also varies with screen grid voltage, as might have been expected, but a peculiar trend was observed experimentally. Lowering the screen grid voltage resulted in an <u>increase</u> in the input impedance. This confirms other investigations that have been made on 6BN6 limiters.

Regarding these other investigations, the results of which were given pessimistic consideration by those reporting (3) (11), it was felt that the 6BN6 limiter, aside from its disadvantages, exhibited the best limiter characteristic obtainable for limiters using relatively simple circuits. This warranted its exclusive use as a limiter in both the weaker-signal suppressor and in the demodulator.

Fig. 5.3 depicts the circuit used to obtain one stage of narrow band limiting in the weaker-signal suppressor. Four of these circuits were cascaded to provide the desired weaker-signal suppression properties. Certain circuit components require further discussion. The arrangement used to provide signal grid and quadrature grid (suppressor) bias is as shown. This is the same bias arrangement used by McLaughlin in his design of the 6BN6 limiter. (11) It is not a good arrangement since the controls are interdependent. A better scheme would have been to use negative bias obtained directly from a negative voltage supply. However, after the circuit was built, satisfactory response was still achieved by grounding the cathode thus short-circuiting the bias potentiometers. Actually, as it was later determined, the only critical adjustment for this limiter circuit was the screen grid voltage. The screen grid voltage had the greatest effect on all three of the limiter characteristics; threshold, flatness beyond threshold, and output gain. An average value for the screen grid voltage which resulted in good limiting properties for various 6BN6 tubes, was found to be approximately 50 volts.

In an effort to lower the threshold beyond that which could be achieved by just providing bias control of the 6BN6, a stage of gain was provided preceding each limiter stage. This amplifier was a 6AH6 high gain pentode employing a wideband single tuned circuit in its output. With this stage of gain preceding each limiter the effective threshold for each circuit was about 0.2 volts. Later, after some circuit compromises had to be made, the

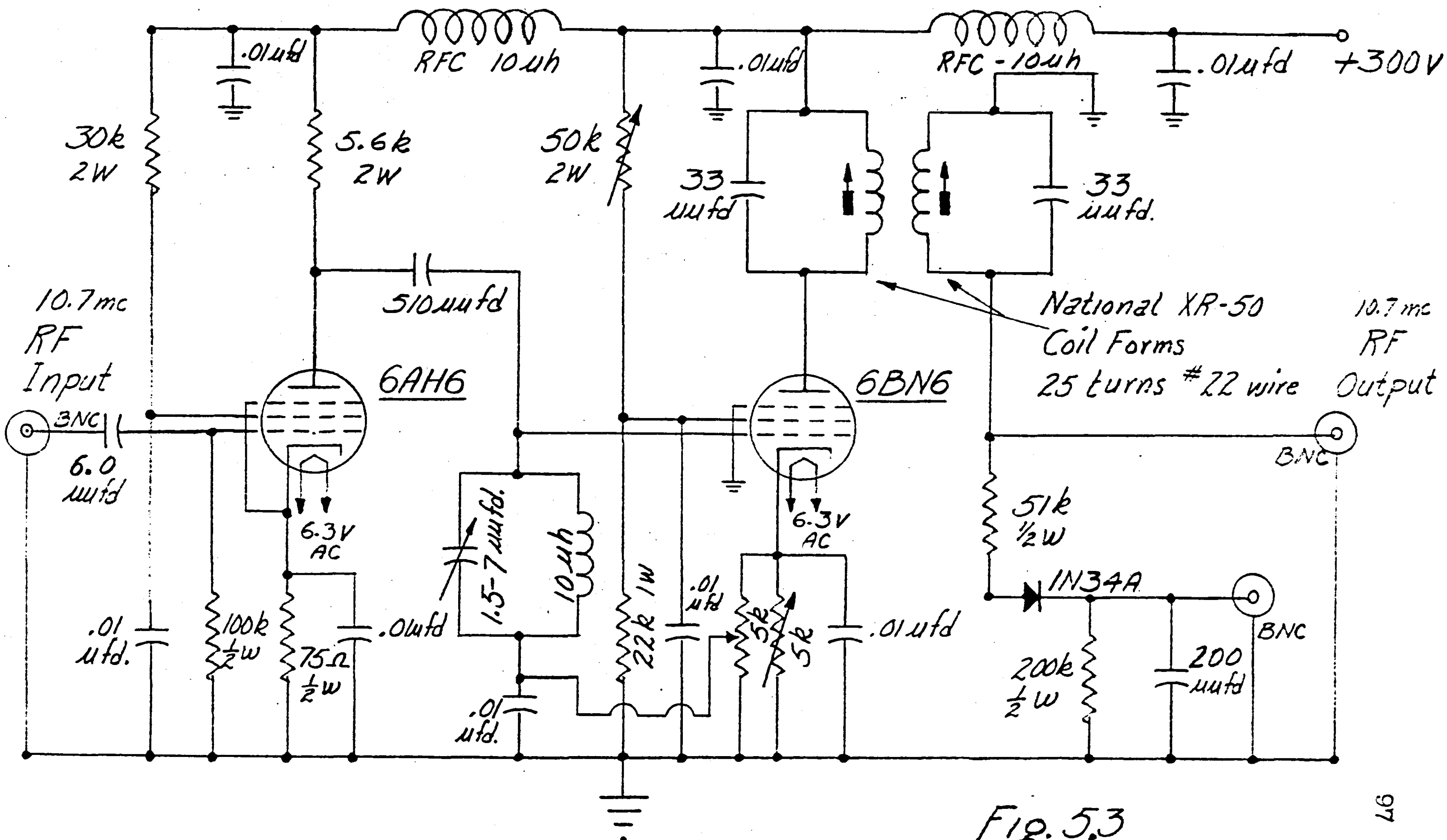


Fig. 5.3

Low Threshold
Narrow - Band Limiter

Joseph M. Gutwein

limiter-unit* threshold was increased to 0.5 volts. This threshold is still low enough for a given median signal level (E_s), to ensure limiting under most all conditions of signal interference.

The narrow-band filters used in the limiter output consisted of mutually-coupled, double-tuned circuits. Each double-tuned circuit was slightly overcoupled in an effort to improve both the skirt selectivity and the flatness of the inband response. Since each limiter unit was enclosed in a separate chassis, shielding was not necessary and the coils were made in the laboratory. This permitted use of good quality coil forms to provide high inductances and high Q. Using these laboratory wound coils facilitated ease in alignment to achieve the desired filter response.

One disadvantage of the limiter-unit was its high output impedance. With this high output impedance, the voltage out was in excess of 10 volts. This provided too much drive for the 6AH6 amplifiers and the high impedance input desired of the limiter-unit could not be realized without certain modifications. Rather than couple directly into the amplifier from the preceding limiter-unit or from the IF amplifier, a small capacitor (6 mmfd) was placed in series with each 6AH6 grid lead. This small capacitor provided part of the tuning capacitance of the preceding filter. However, inter-unit coupling in this manner attenuated the input signal. Since the limiter-unit threshold was already very low, some attenuation could be afforded in order to preserve the filter response. With this small capacitor in the input, the limiter-unit threshold was increased to 0.5 volts. One further mention in using

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* By limiter-unit we mean the entire limiter circuit shown in Fig. 5.3.

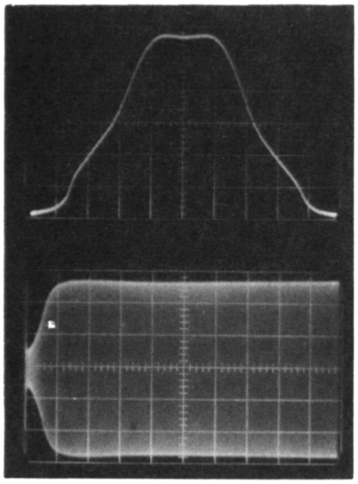
this form of inter-unit coupling is that the signal appearing directly at the grid of the 6AH6 is still sufficient to drive the grid into conduction. Actually then, this tube is performing a preliminary limiting operation and the self bias circuit consisting of this small capacitor and the grid-leak resistor had to be designed accordingly. This accounts for the use of the 100 kilohm grid-to-ground resistor providing a grid-charging circuit of 0.6μ sec.

Fig. 5.4 illustrates the limiter and filter characteristics of each limiter-unit. The thresholds of the limiters are all approximately 0.5 volts and the BW of each filter is roughly 200 kc wide (which is also the IF bandwidth). Each limiter was designed to handle an interference ratio of $\underline{a} = 0.85$ for a median signal level of E_s = 5 volts. For larger $\underline{a}$, the BW of the filters in the cascaded limiters would have to be tapered from about three IF bandwidths down to about one IF bandwidths. Rather than have to upset the tuning of the filters an $\underline{a} = 0.85$ was the upper design limit.

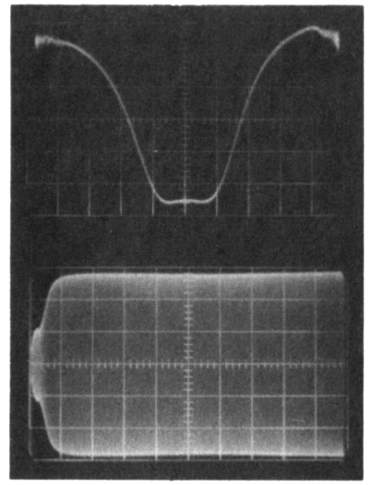
Design of Demodulator

It was desirable to have a high capture-ratio demodulator for the feasibility study of this weaker-signal receiver. In order to achieve this result, both Baghdady's narrow-band theory and Granlund's (6), (7) wide-band theory were incorporated into one design.

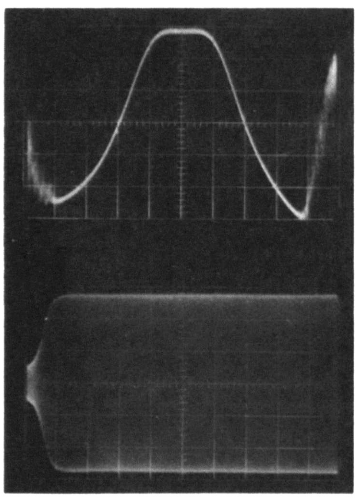
The demodulator consisted of two IF amplifiers, two stages of narrow-band limiting, a wideband low time constant discriminator



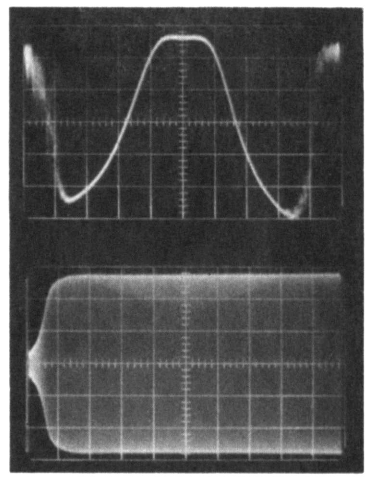
Narrow-Band
Limiter No. 1



Narrow-Band
Limiter No. 2



Narrow-Band
Limiter No. 3



Narrow-Band
Limiter No. 4

Note: Limiter Amplitude Response
Hor. Axis Calibration = 0.5 volts/cm

Limiter Filter Response
Hor. Axis Calibration 75 kc/cm

Fig. 5.4 Weaker Signal Suppressor Limiter Amplitude and Filter Response and an audio section employing a 3 kc sharp cut off, low-pass filter; this is all shown in the schematic of Fig. 5.5.

The IF amplifier was conventional providing a gain of about 400 over a BW of approximately 300 kc.

The two narrow band limiters were of the 6BN6 type described in the preceding section. The optimum screen voltage providing for the lowest threshold, but still maintaining sufficient signal level in the output was obtained as shown in the schematic. Commercial cans were used as interstage double-tuned filters. Their response was not as good as the specially designed filters in the weaker-signal suppressor unit; however, it was still possible to achieve a flat-topped response in both limiter filters using these cans. The bandwidths of each of these filters is roughly 400 kc and 250 kc respectively.

In a first attempt to achieve a high capture-ratio, the manufacturers suggested circuit was used for a narrow-band, Foster-Seely discriminator. Fig. 5.6 is the capture plot of the demodulator for this design. For this receiver, the capture ratio was only 0.5 which was unacceptable. (The capture ratio is defined here as that amplitude ratio for which the distortion rises to 10%.)

In an effort to improve the capture ratio of the demodulator, a commercially available wide-band discriminator having a BW of 900 kc peak-peak was incorporated into the design. Instead of using the manufacturers suggested circuit for the discriminator, a modification was made based on the discriminator theory published by Baghdady. (6) This change consisted of lowering the time

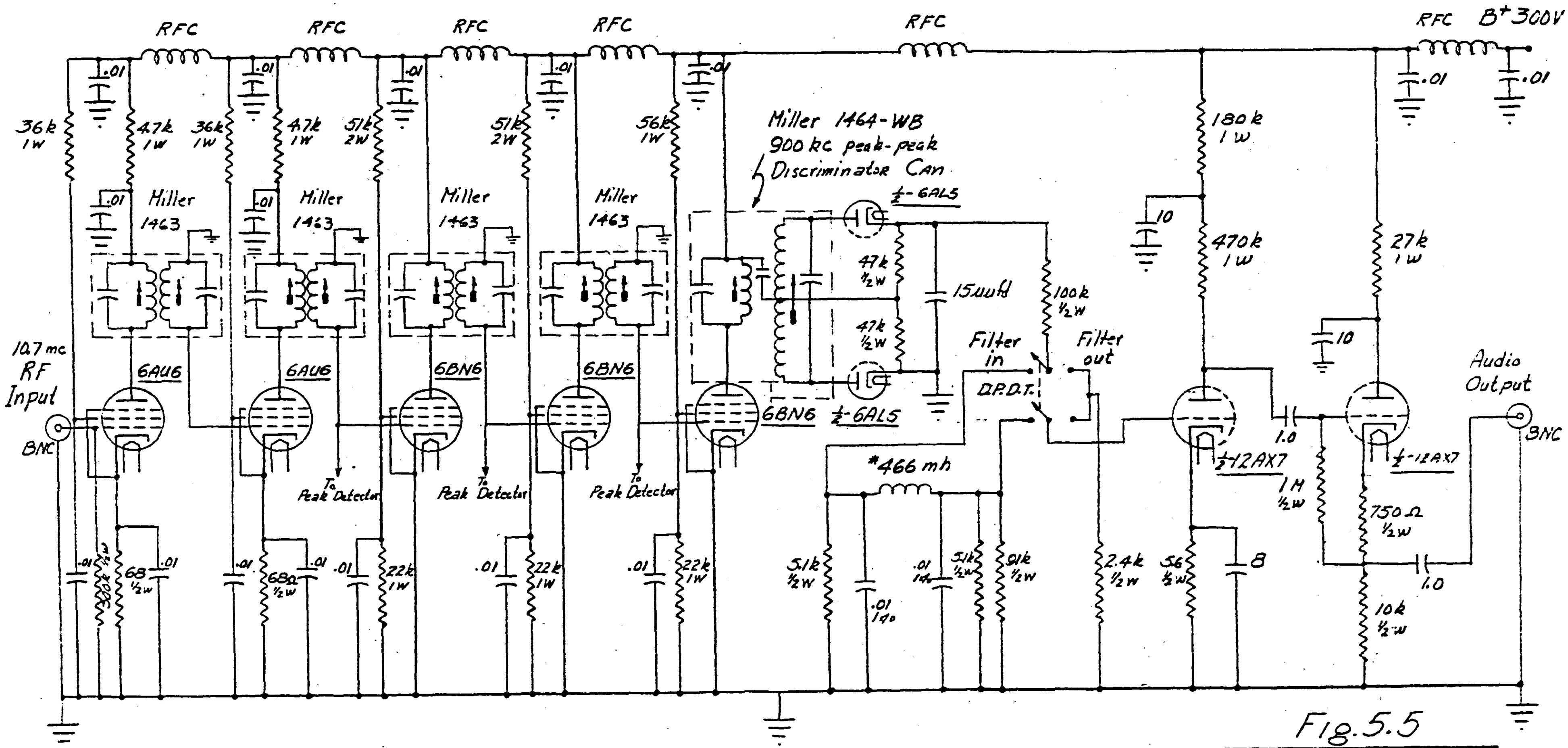


Fig. 5.5

Note: All Capacitors in μfd. (±20%)
All Resistors in kilohms (±5%)
unless specified otherwise
* Sharp Cutoff 3kc Low Pass Filter (20db/octave)
Inductor Core Material:
Ferroxcube S-35/23-K3-001-04

High Capture Ratio
FM Demodulator
SM Thesis MIT
Joseph M. Gutwein

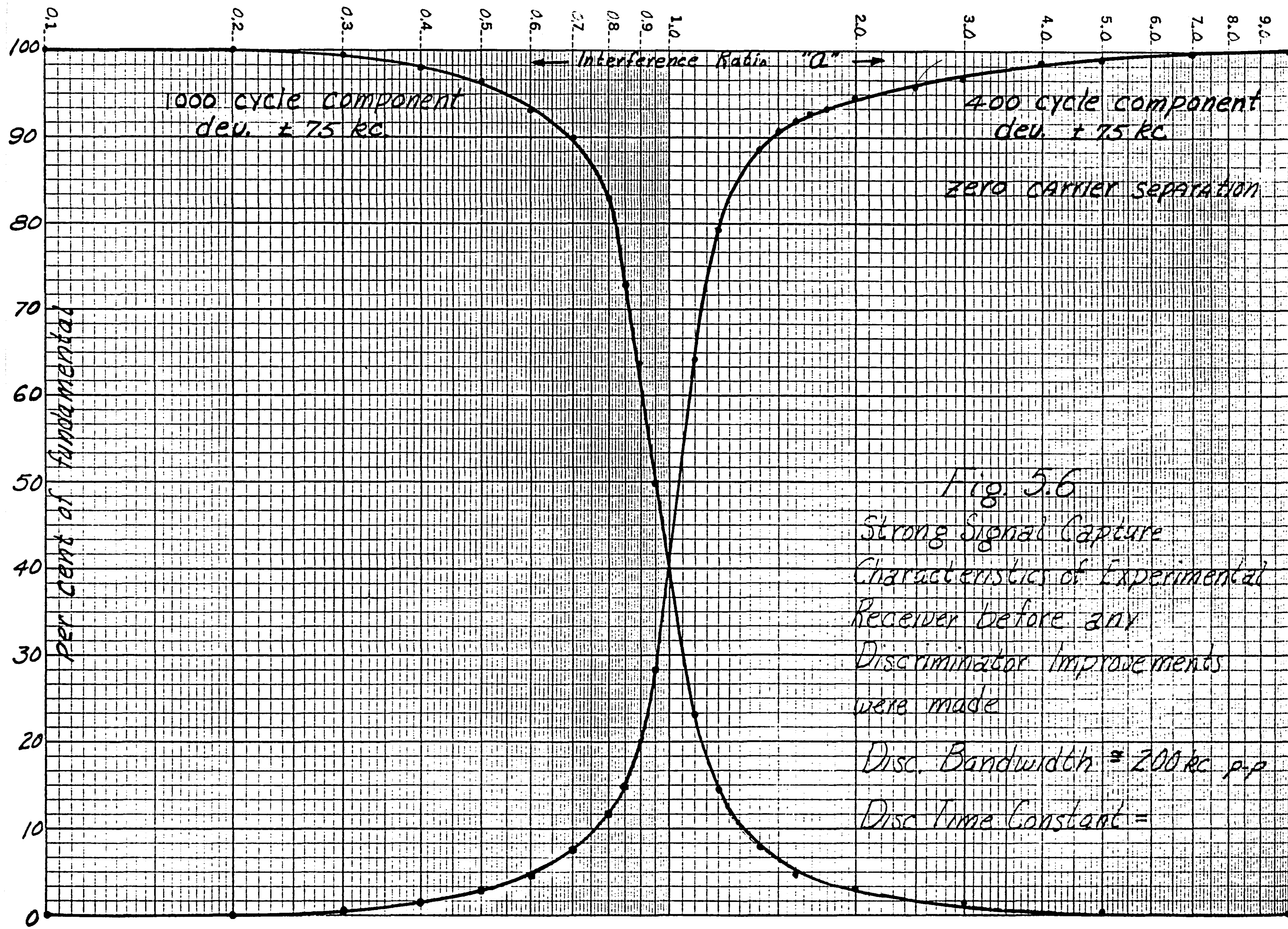


Fig. 5.6
Strong Signal Capture Characteristics of Experimental Receiver before any Discriminator Improvements were made

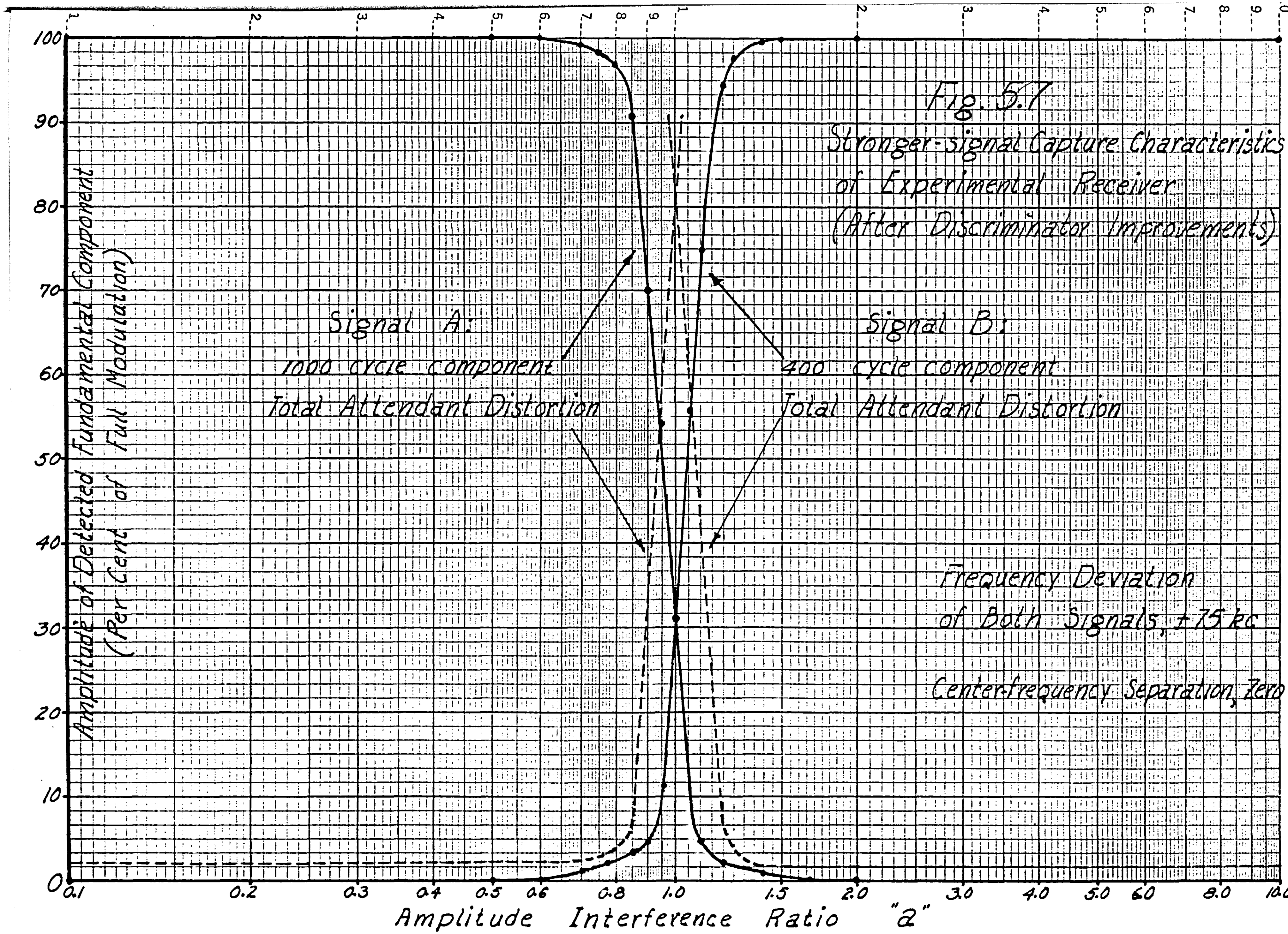


Fig. 5.7
Stronger-signal Capture Characteristics
of Experimental Receiver
(After Discriminator Improvements)
Signal A:
1000 cycle component
Total Attendant Distortion
Signal B:
400 cycle component
Total Attendant Distortion
Frequency Deviation
of Both Signals, ±75 kc
Center-frequency Separation, Zero
Amplitude of Detected Fundamental Component
(Per Cent of Full Modulation)
Amplitude Interference Ratio "a"
0
10
20
30
40
50
60
70
80
90
100
0.1
0.2
0.3
0.4
0.5
0.6
0.8
1.0
1.5
2.0
3.0
4.0
5.0
6.0
8.0
10.0

104

constant of the discriminator detector circuits.

Fig. 5.7 shows the stronger-signal capture characteristics of the demodulator after the discriminator improvements were made. The capture ratio as defined previously is now 0.86 which represents considerable improvement.

The stages following the discriminator consist of a 3 kc low-pass filter, an audio amplifier and a cathode follower. The audio stages are standard and need no explanation. However, the 3 kc low-pass filter was incorporated into the design to provide adequate filtering of the demodulated FM signal. The cut off characteristics are a little more abrupt (approximately 20 db per octave) than that usually found in standard FM demodulators using 75μ sec de-emphasis filtering. The reason for using this filter was to minimize the high audio-frequency distortion accompanying the capture of the weaker signal message when the demodulator was used in conjunction with the weaker-signal supressors and fixed-trap circuit. A special switching arrangement is used to switch this filter in or out. In order to prevent disrupting of the discriminator time constant and to provide the same audio signal level in the output with the filter in or out, the switch circuit is arranged as shown.

Design of Delay Equalizer

The function of the delay equalizer is described in Chapter III. Unfortunately, the need for this delay equalizer was not fully realized until late in the time alloted for this thesis; thus, proper consideration could not be given to a good design.

However, an equalizer was improvised which served the purpose adequately.

Before embarking on the design of any equalizer, one must know beforehand exactly what he is to equalize, and then to design accordingly. With this in mind, the time delay of the circuits whose delay we had to compensate was first calculated and then later measured. The circuits whose delay had to be equalized consisted of the entire cascaded section of narrow-band limiters, the converter and the converter-first mixer interstage circuits. Totaling the filters employed in all these circuits we have; six double-tuned circuits having a BW of approximately 200 kc each and five single tuned circuits having a BW of 400 kc each. According to Baghdady (5) and the theory presented in Chapter III, each of these filters introduces a time delay into the message modulation equal to the slope of the filter phase characteristic at the center frequency of the filter.

We will assume that the double-tuned circuits are maximally flat so that their response can be described by a second-order Butterworth response function. For this filter then, the phase characteristic is given by

$$\phi(\omega) = -\tan^{-1}\left[\frac{\sqrt{2}\,\frac{\omega-\omega_0}{BW/2}}{1-\left(\frac{\omega-\omega_0}{BW/2}\right)^2}\right] \qquad 5.1$$

Differentiating this expression and evaluating at $\omega = \omega_0$, we have for the time delay at the center frequency

$$\phi'(\omega_0) = 2\sqrt{2}/BW$$

(Note that the BW must be expressed in radian per second!)

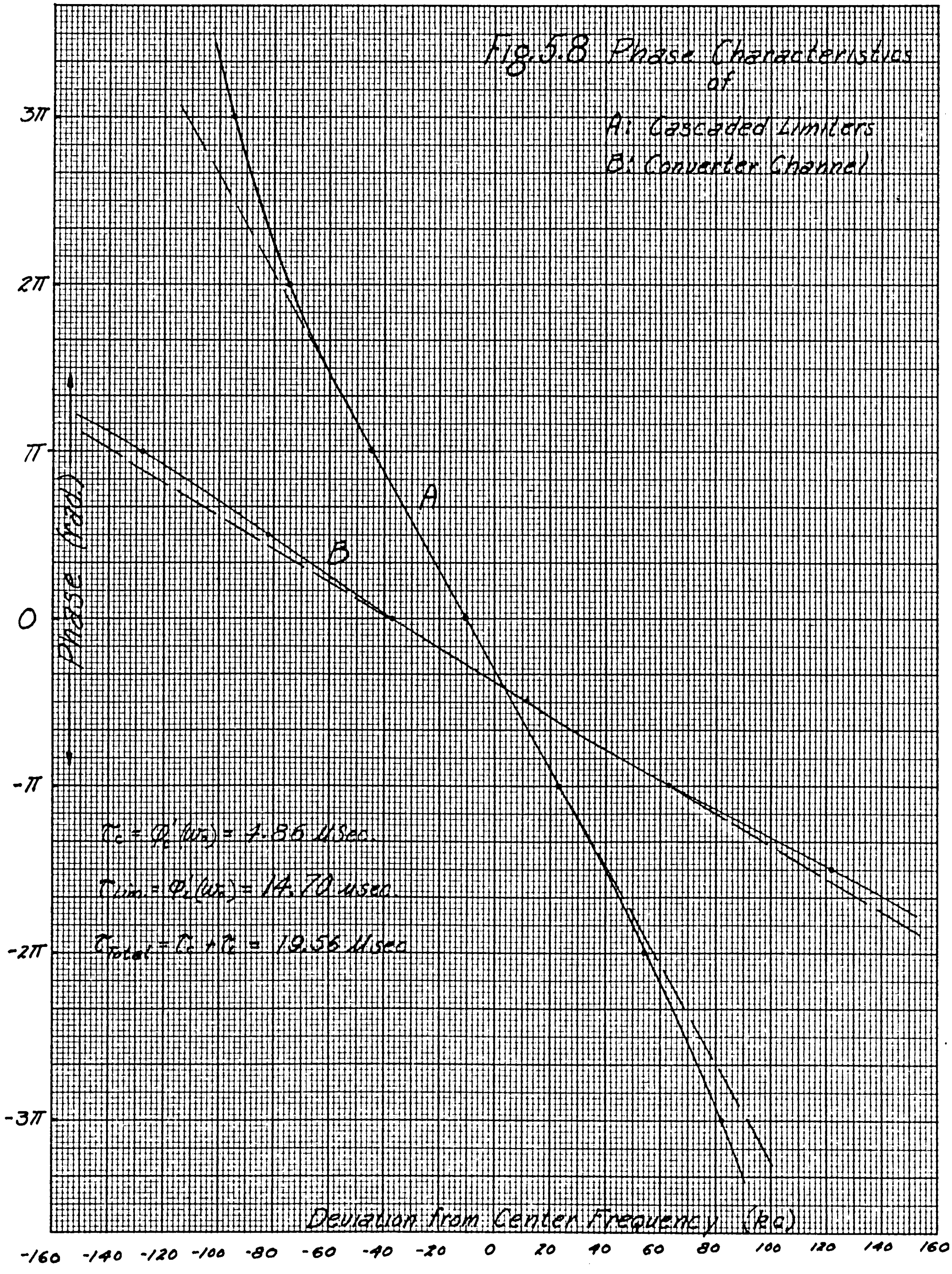


Fig.5.8 Phase Characteristics
of
A: Cascaded Limiters
B: Converter Channel
3π
2π
π
0
-π
-2π
-3π
Phase (rad)
A
B
$\tau_c = \Phi_c'(\omega_0) = 4.86$ μsec.
$\tau_{lim.} = \Phi_L'(\omega_0) = 14.70$ μsec.
$\tau_{Total} = \tau_c + \tau_L = 19.56$ μsec
Deviation from Center Frequency (kc)
-160 -140 -120 -100 -80 -60 -40 -20 0 20 40 60 80 100 120 140 160

For a BW = 200 kc, the time delay is

$$\phi'(\omega_0) = 2.25\,\mu sec.$$

Similarly, for a single tuned circuit whose phase characteristic is given by

$$\phi(\omega) = -\tan^{-1}\left[\frac{\omega-\omega_0}{BW/2}\right] \qquad 5.2$$

Differentiating, we have

$$\phi'(\omega_0) = 2/BW$$

For a BW = 400 kc

$$\phi'(\omega_0) = 0.8\,\mu sec.$$

Totaling the theoretical delay from each filter we have

$$\phi'_{TOTAL}(\omega_0) = 17.5\,\mu sec.$$

It is apparent that these delay calculations are solely dependent upon the BW of the filters. Since it is quite difficult to accurately measure the BW of the filters at RF frequencies, it was decided to obtain the actual phase vs. frequency characteristic by the lissouious method. A good balanced-amplifier high-frequency, Tektronix scope and the GR standard signal generator, set to the center frequency of the filters, were used to obtain the phase characteristics. The phase characteristics for the four narrow-band limiters and the converter channel are shown in Fig. 5.8. From these characteristics we note slight departures from phase linearity especially at the edges of the pass band. The slope of these curves is the delay which we wish to compensate.

Realization of a delay line at a frequency of 10.7 mc with a BW in excess of 200 kc is a thesis in itself. It is almost impossible to achieve passively with lumped LC elements since the

Q required of the inductors is exorbitant and the number of poles required to produce the phase characteristic is excessive. These problems are also compounded with alignment difficulties at 10.7 mc. Nevertheless, two different delay equalizers were built in an attempt to compensate for the delay acquired by the strong signal message in the limiter and converter channel filters. The first delay equalizer that was designed is shown in the block diagram of Fig. 5.9. The circuit used to produce the delay consisted of a special delay cable capable of providing a delay of 1μsec per foot. This delay line was tapped at one foot intervals with each tap connected to a rotary selector switch making it possible to select any delay to within a micro-second from 10μsec to 20μsec. The output of the rotary switch was connected to a second delay line which was continuously variable from 0 to 1μsec, thus providing a vernier control. In this manner, we had provision for continuously varying the delay from 10 to 21μsec.

Two disadvantages associated with this delay equalizer were encountered. First, the delay cable had a 6 db cut off frequency at approximately 5 mc thus necessitating the circuits shown. In these circuits, the 10.7 mc signal was hetrodyned to 2.2 mc by a crystal local oscillator at 8.5 mc. The 2.2 mc signal was then delayed by the delay line by whatever value selected. The output of the delay line was hetrodyned back to 10.7 mc, the frequency of operation of the existing equipment.

The second difficulty encountered, which led to the abandonment of this scheme, was that although this circuit had linear

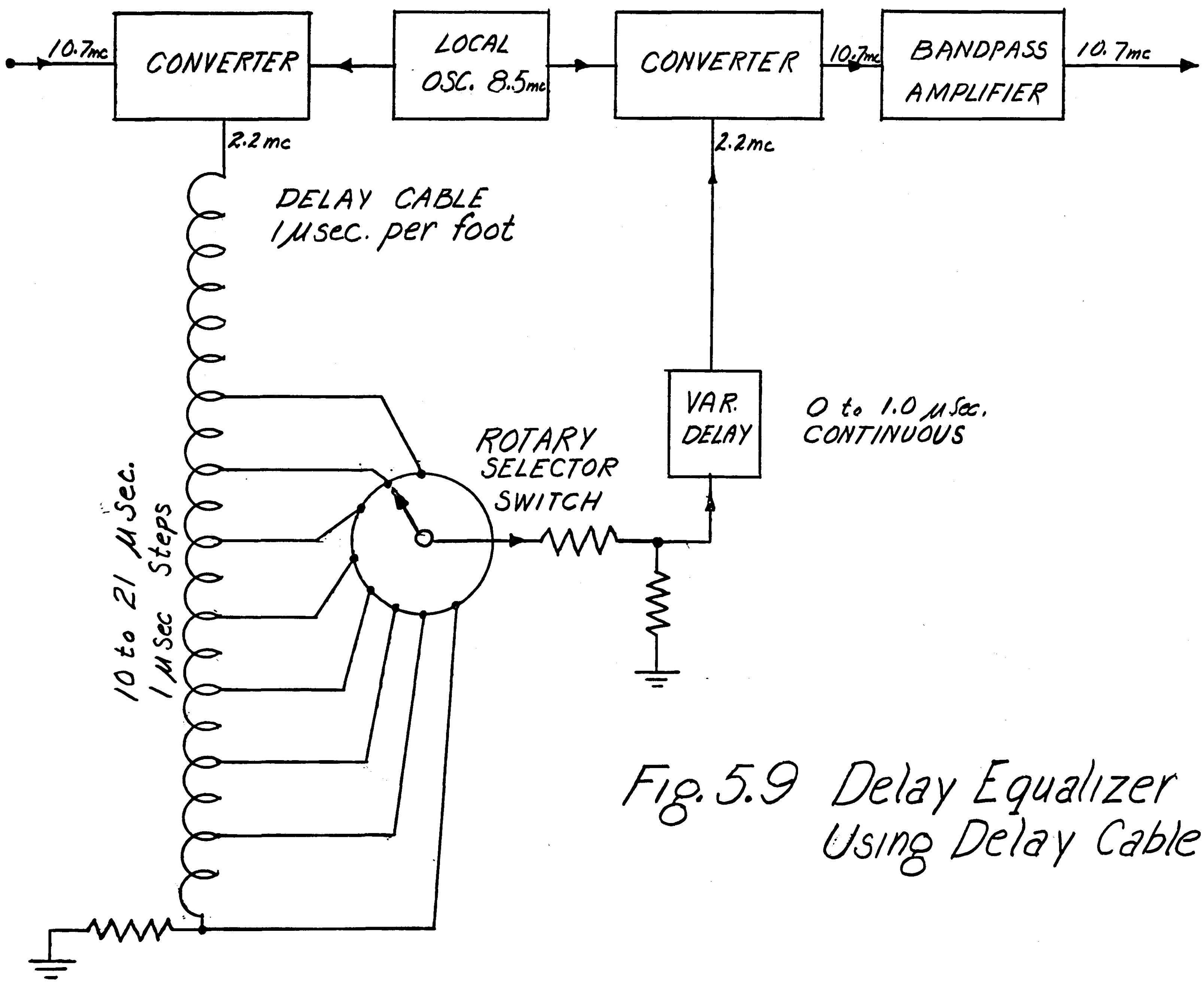


Fig. 5.9 Delay Equalizer Using Delay Cable

phase response, the amplitude vs. frequency response was not flat over the BW of interest (200 kc). To make matters worse, the amplitude vs. frequency response of the output of the entire delay network changed whenever the delay was varied. This amplitude variation could possibly have been a result of improper terminations for either the cable or the variable line (which both required different terminations). A stage of limiting could also have been provided in order to eliminate the resulting AM of the signal. However, there was greater than a 20 db loss of signal strength through the entire circuit necessitating two or more stages of amplification to provide enough signal level to drive the limiters. This would have made this equalizer too complicated and it was decided to abandon this particular scheme entirely.

In lieu of these difficulties and since time was running out, the brute-force approach was taken. This required cascading nine identical IF amplifiers (which were available in the lab) to provide for the delay equalization. Each amplifier was provided with a double-tuned circuit and an attenuator. The attenuators were used after each stage of amplification and filtering. Two factors made it necessary to use these attenuators. First, it was required that the entire cascaded section of filters be linear; therefore, the attenuators were used to reduce the signal level after each stage of amplification and filtering. Secondly, the attenuators reduced interstage regeneration making it possible to tune each filter independently of the others.

In essence, the tubes and attenuators act only as interstage,

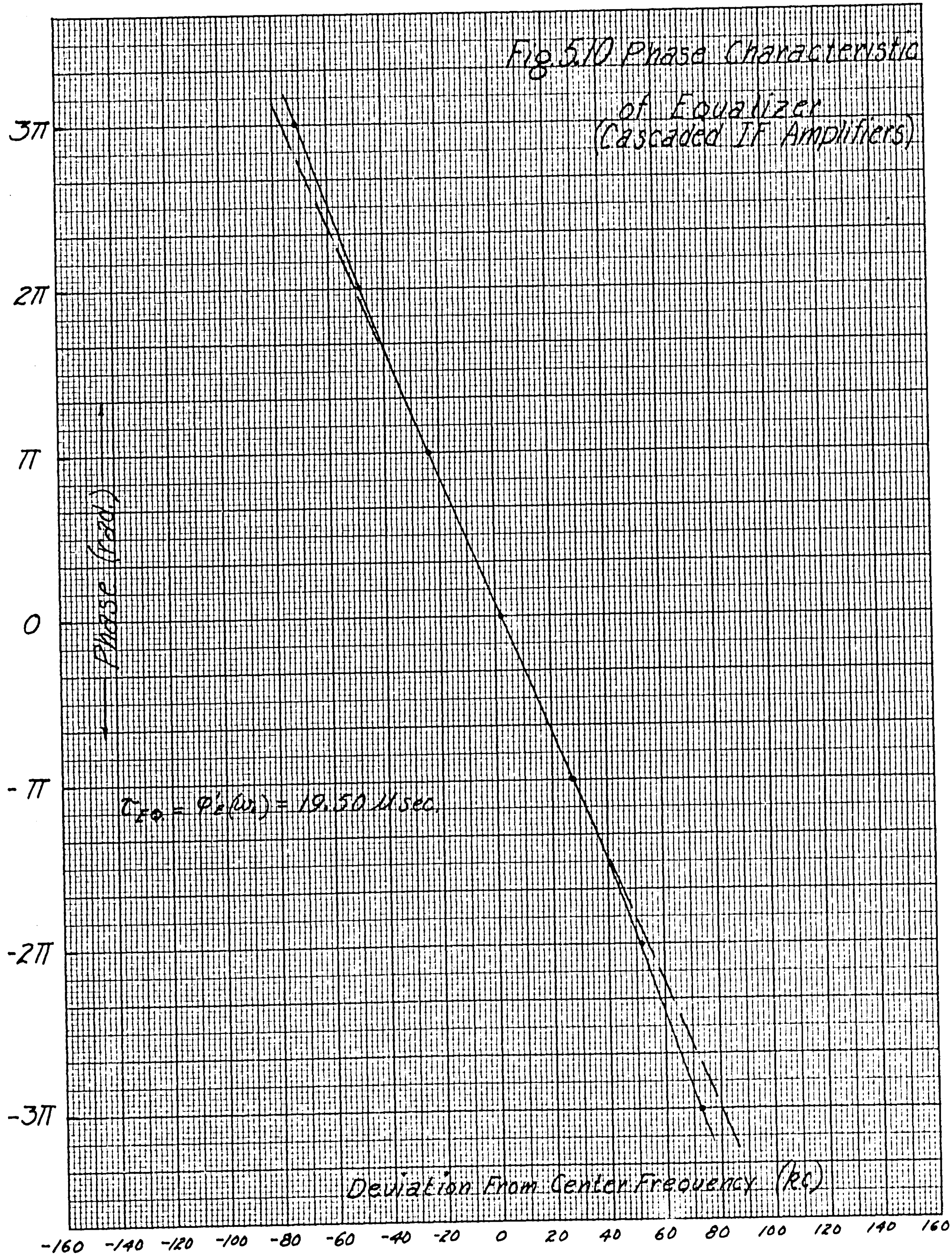


Fig 5.10 Phase Characteristic of Equalizer (Cascaded IF Amplifiers)
3π
2π
π
0
- π
-2π
-3π
Phase (rad.)
$\tau_{E\phi} = \varphi'_E(\omega_1) = 19.50\ \mu sec.$
Deviation From Center Frequency (kc)
-160 -140 -120 -100 -80 -60 -40 -20 0 20 40 60 80 100 120 140 160

high-impedance, coupling devices for the synchronously tuned filters. From input to output of each stage and from input to output of the entire cascaded section there is approximate unity gain.

Although this scheme involved a great deal of redundancy, its one justification was that it provided a phase characteristic which was approximately the same shape as that of the limiter and convertor channel filters. Fig. 5.10 shows a plot of the equalizer phase characteristic. In comparing Fig. 5.10 to Fig. 5.8 it can be seen that the phase nonlinearties of each channel were almost identical.

One of the disadvantages of this equalizer besides the fact that it was cumbersome, was that it did not provide any flexibility for selecting various delays without having to completely realign the entire filter. It was also difficult to provide for exact delay equalization by this method whence the delay equalizer compensates only to within one microsecond. However, this was found to be a close enough approximation to improve the performance of the fixed-trap circuit.

Chapter VI

CONSTRUCTION AND ALIGNMENT OF SYSTEM

Some Comments on Design and Construction

In this chapter a few comments pertinent to the overall system design, construction and alignment are mentioned.

In an effort to facilitate simple construction of most of the circuits, commercially available coils and transformers were used wherever possible. These coils were provided with their own shields; hence it was not necessary to shield each stage individually by providing partitions. However, when high Q open coils were used there was always the possiblity of regeneration because of stray coupling; thus, special shielding precautions were taken. Although shielded coils were predominantly used in the fixed trap circuit, partitioning of each stage was still provided. This was done in an effort to minimize the spurious effects that could arise because of the presence of so many different FM signals. However, partitioning was not necessary in the demodulator or the delay equalizers where there was only one signal channel and where commercial cans were used extensively.

Although use of the commercial coils facilitated the mechanical construction and layout of the circuits, their use created other problems. First of all, these coils rely mainly upon the stray capacities for their tuning. Because of this dependence on stray capacitance, they are susceptible to both mistuning because of thermal drifts and regeneration due to the grid-to-plate capacity of the following stages which is augmented by the familiar Miller effect. As a result of this interstage feedback, synchronous tuning of cascaded identical stages was difficult to achieve.

In the fixed trap circuit, where partitions enclose and shield each stage, it was necessary to use feed-through connectors through each partition. These provided each stage with signal, supply voltage and filament voltage. The feed-through connectors used for the signal were of the low capacity to ground type while the supply and filament voltage feed-through connectors were of the bypass type providing 1500 mmfd to ground.

Decoupling networks in all the filament and supply circuits were used extensively. These decoupling networks were series inductance, shunt capacitance, low pass pi-filters. Commercial choke coils were used for the decoupling networks in the plate circuit. These varied in inductance from 10 to 50 micro-henries depending upon the RF frequency of the signals in the stages that were being decoupled. The filament chokes used in the filament decoupling networks were wound in the lab and were roughly 5 micro-henries each.

In general all the bypass capacitors used were a .01 mfd miniature ceramic disc capacitors. Actually this value of bypass capacitance was a little high since at 10 mc only .025 micro-henries is required to resonate with .01 mfd. Long bypass leads alone may contribute this much inductance. As a result, parasitic oscillations were occasionally encountered, especially in the high gain IF amplifier. However, relocating the bypass ground connections, shortening the leads, and then reducing the value of bypass capacitance from .01 to .001 mfd usually cured the trouble if the circuits had a tendency to oscillate.

A center partition was erected with no filament or supply circuit cross-overs to either channel. The only penetrations in this partition were made at opposite ends of the chassis in order to accomodate the signal. The filaments of all the tubes were bypassed directly to ground at the tube socket. Good bypassing arrangements were conducive to trouble free operation of the circuits. Where possible, all bypass capacitors were grounded to the same point of the chassis and all bypass capacitor leads were kept short. All signal leads were also kept to a minimum length when possible, to avoid stray coupling.

Method Used to Align Filters

Essential to the good performance of the system is the requirement of having all the filters accurately aligned to the proper center frequency with the correct amplitude vs. frequency characteristic. In the discussion that follows, a method most suitable for aligning the filters will be described after which a general alignment procedure for the entire receiver will be presented.

Fig. 6.1 illustrates the manner in which the test equipment is arranged to align the system. The audio signal used to frequency modulate the FM generator is also applied to the horizontal input of a scope with the horizontal selector switch of the scope in the external (or driven) sweep position. With the frequency deviation of the FM generator set to a convenient level,

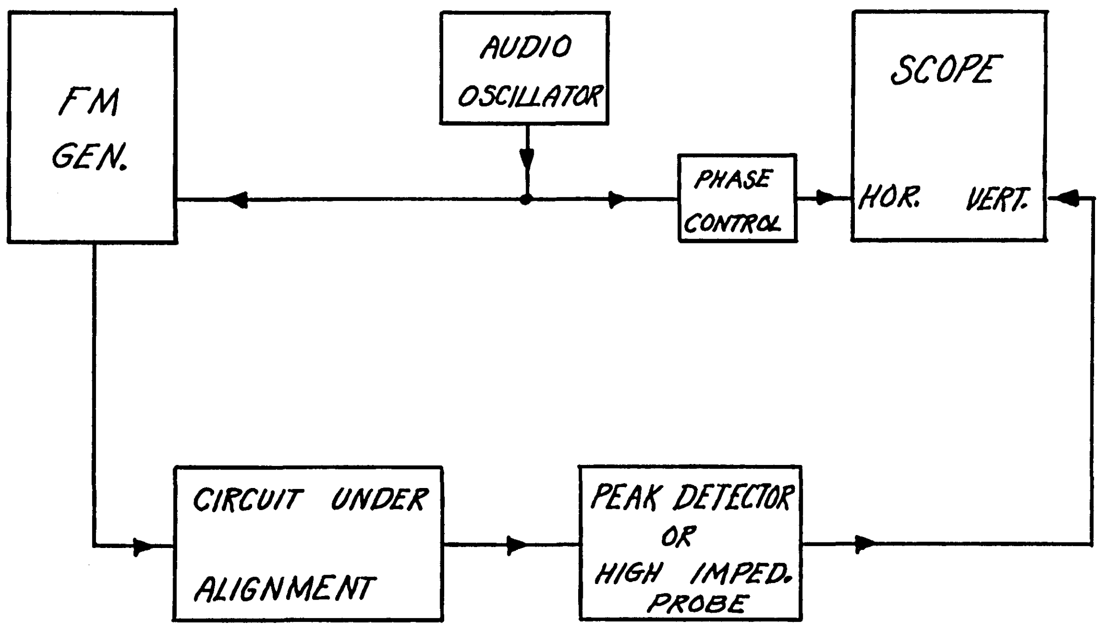


Fig. 6.1 Equipment Arrangement for Filter Alignment

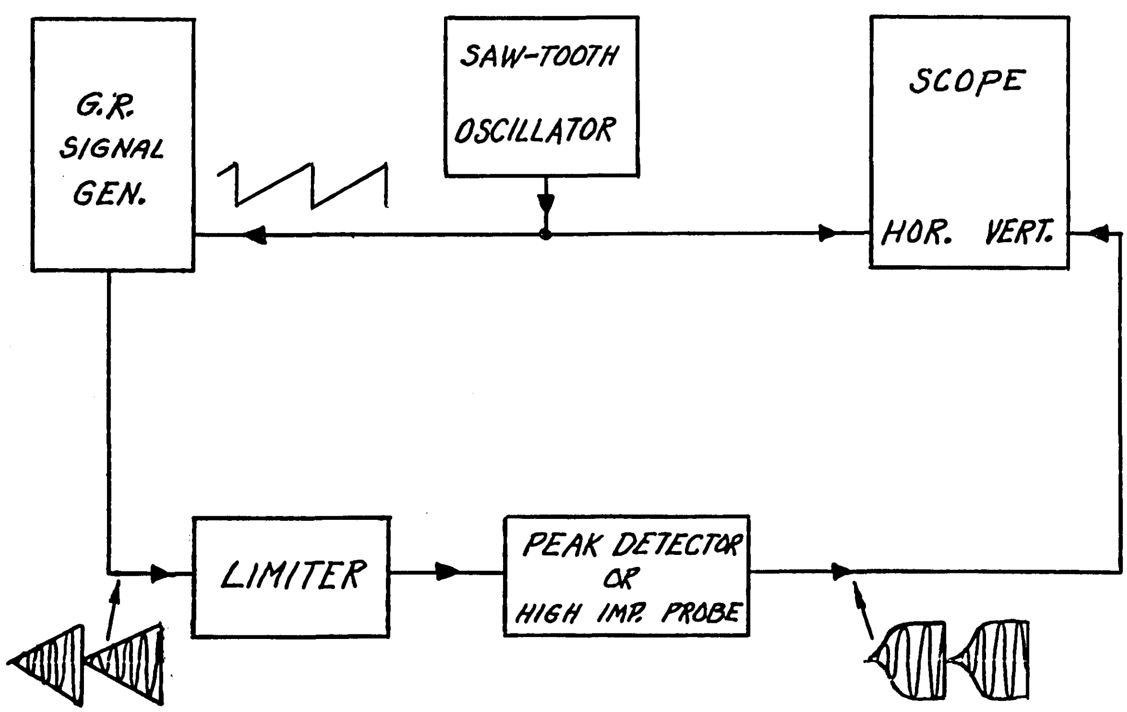


Fig. 6.2 Equipment Arrangement for Displaying Limiter Amplitude Characteristic

the line trace on the scope is centered and adjusted to the desired length. The center of the line trace represents zero deviation or the center frequency of the FM signal that is applied to the circuit being aligned. The length of the line trace represents the total frequency deviation of the FM generator. In accordance with this deviation the scope sweep is calibrated in frequency of deviation in kilocycles per unit length of the trace in cm. For example, assume that the FM generator is frequency deviated to $\pm$ 150 kc. This represents a total change of frequency of 300 kc. If the line trace on the scope is centered and adjusted to 10 cm. in length, the scope reads 300 kc of frequency deviation for 10 cm. or 30 kc/cm.

Once the scope is calibrated to the FM deviation, the lowest audio signal frequency is selected to modulate the generator (recalibration is usually necessary when audio frequency changes are made in the FM generator). The lowest audio frequency should be used because in this alignment procedure it is absolutely necessary that the filters be responding to the FM excitation through quasi-stationary states. If the filters are not responding to the FM excitation through quasi-stationary states, the characteristic displayed on the scope would immediately indicate this. Reduction of either the sweep frequency or the frequency deviation would correct matters.

The FM signal is applied to the circuit under alignment. The RF amplitude of this signal is adjusted to permit linear operation of all the stages in this circuit. If the circuit under test is provided with a peak detector that is permanently wired in, the output of this peak detector can then be applied directly to the vertical input of the scope. The image appearing on the scope is the amplitude vs. frequency characteristic of the filter. The circuit can then be tuned to provide for the desired filter shape.

If the circuit under test is not provided with a peak detector, a high impedance scope probe may be used. However, use of a probe has the disadvantage of loading the circuit capacitively. Removal of the probe results in a mistuning of the filter.

In the output of most of the filters in this receiver, peak detectors are permanently wired in as shown in the schematics. The advantage of using these peak detectors is that a quick check can be made of the filter without disturbing the filter or its environment. So as not to load the filters with the diode and RC circuit, a high resistance (usually 51 kilohms) was inserted in series with each diode.

Method of Displaying Limiter Characteristics

In order that we may adjust the bias for each limiter, it was necessary to remove all ambiguity from the procedure by displaying the actual limiter characteristic on a scope. This characteristic is a plot of the amplitude of the limiter response vs. the amplitude of the limiter input. The method used to display the limiter characteristic is closely related to that used for displaying the amplitude vs. frequency response of the filters described in the previous paragraph. The block diagram for the equipment arrangement is shown in Fig. 6.2.

In displaying the limiter characteristic, an audio frequency (approximately 300 cps) sawtooth oscillator (horizontal sweep output from another scope) is used to amplitude modulate (to 100%) a GR standard signal generator. The RF frequency of this generator is adjusted to the center frequency of the limiter filter that was previously aligned. The sawtooth signal is also applied to the external horizontal input of a high frequency scope. The scope line trace is centered and adjusted to the full width of the scope face. To calibrate this line trace in volts per cm of deflection, the sawtooth signal is applied to the vertical input of the display scope (which for the moment is internally swept). The peak to peak amplitude of the sawtooth signal is adjusted to some convenient value. Knowing this amplitude of the sawtooth signal, the line trace of the display scope when swept externally is now calibrated to volts drive per cm. of deflection. For example, assume that we had a sawtooth signal equal to 5 volts peak to peak and that the length of the line trace on the display scope was 10 cm. The deflection of the trace is calibrated at 5 volts for 10 cm of deflection or 0.5 volts/cm.

The amplitude modulated output of the RF generator is now applied to the input of the limiter. The limiter response is obtained through a peak detector or from a high impedance probe in the output of the limiter filter. This signal is applied to the vertical input of the display scope whose horizontal sweep is driven by the external sawtooth signal. The image appearing on the scope (see Fig. 5.4) is the limiter characteristic. Since the horizontal scale was previously calibrated, we can immediately ascertain the limiting threshold of the limiter.

With the limiter characteristic displayed in this manner, the effect of any bias adjustments can be directly observed and the limiter response optimized.

Alignment of the System

In any alignment procedure, it is good practice to start with the last stage first and then work back towards the input. With this in mind, the demodulator is the first circuit to be aligned. Once this is aligned, we have established a frequency reference. The remainder of the system *must* be aligned to the center frequency of the demodulator filters.

The first step in aligning the demodulator is to align the IF amplifier and limiter filters. A check at this point is simultaneously made of the input drive level to the IF amplifier. The input level must be adjusted so as to provide sufficient drive for the limiters (5 times the limiter threshold for a good demodulator) but not so high so as to over drive the linear IF amplifier. After the amplitude check is made, and the IF and limiter filters are aligned, the discriminator can then be aligned.

The alignment of the discriminator is best achieved by expanding the FM deviation of the signal generator to its maximum since the discriminator employed in the demodulator in this thesis is wide band. The discriminator is aligned by obtaining a symetrical "S" curve about the center frequency (center of the scope trace). The linearity of the discriminator "S" curve was aligned in this thesis by setting the deviation of the FM signal generator to ± 150 kc and applying the output of the discriminator to a distortion analyzer. The linearity control of the discriminator was then adjusted to produce a minimum distortion reading.

Once the demodulator is aligned, the filters in the rest of the circuits are aligned to the same center frequency. Since peak detectors are provided in most all filter outputs it is very simple to align this receiver. Only the order in which stages are aligned and the method used to align the trap require mentioning. The order of stage alignment is as follows:

1 - Demodulator
    a - IF amplifier
    b - Limiters
    c - Discriminator
2 - Third mixer and output single tuned circuit
3 - Input IF amplifier
4 - Narrow band limiters
5 - Converter filter
6 - Converter-first mixer interstage amplifier and limiter filters
7 - Converter-third mixer interstage amplifier and limiter filters
8 - Delay equalizer
9 - First mixer and trap filter

The trap was aligned using two generators. The first generator supplies the FM signal (10.7 mc) and is applied directly to the input of the first mixer. A second signal generator (the GR) operating CW at the center frequency of the first FM generator (10.7 mc), is applied to the converter input. The signal level of the GR should be set to approximately 1 volt. The output of the trap response is now available at the trap peak detector. The first mixer filter is aligned. This determines the shape of the overall pass band of the bandpass trap filter. This mixer filter is a pair of high Q, high inductance, mutually-coupled, coils that are over-coupled. The peak to peak BW of this filter is adjusted to approximately 250 kc. Next in the trap alignment, is to roughly center the rejection amplifier single tuned circuit response and the single tuned circuit response in the last stage of the trap. The null is now produced by varying the coarse rheostat control in the bridged-T. The trap notch can then be centered by tuning the bridged-T inductor. In order to obtain the exact trap characteristic it is necessary to "touch up" the following controls: (1) Control of overall pass band by re-tuning the mixer filter. (2) Bandwidth of the notch by re-tuning the rejection amplifier tuned circuit (this provides a control of trap BW from 40 kc to 100 kc). (3) Symetry about a vertical center line by adjusting the last single tuned circuit in the trap amplifier. (4) Trap position and attenuation by adjusting the variable bridged-T inductance and the rheostat control.

The final positioning of the null and the final adjustment of the null attenuation of the trap is made when the system is connected for operation by tuning out the stronger signal.

In conclusion, it should be mentioned that before attempting any alignment, the system should be given sufficient time to warm up to thermal equilibrium - one hour is sufficient. Occasional checks and adjustments are necessary from time to time due to thermal drifts.

Chapter VII

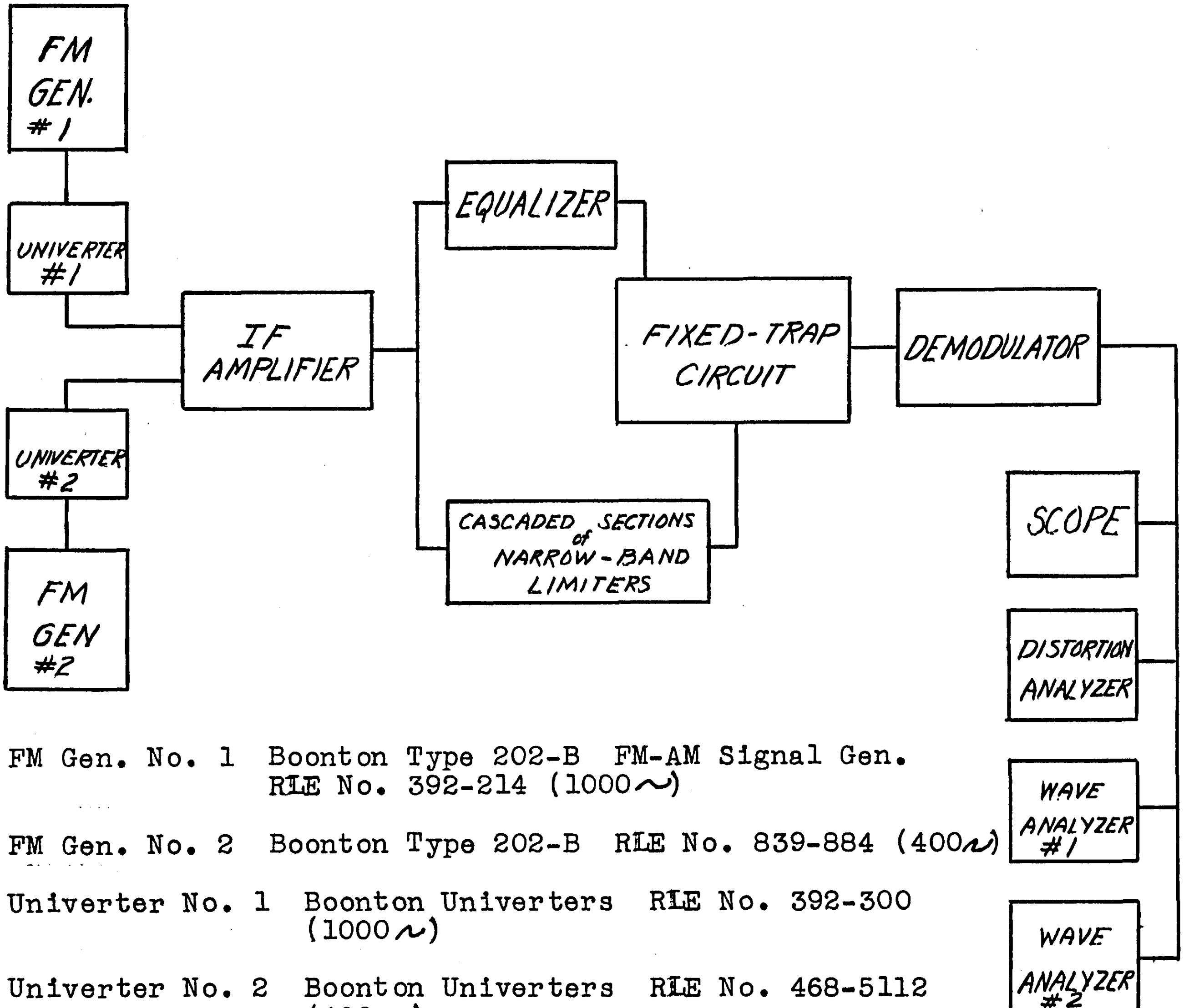
EXPERIMENTAL PROCEDURE

Prior to presenting the data, it is necessary to describe the method used to test the system. Fig. 7.1 depicts in block diagram form, the arrangement and the type of equipment used for testing the system.

Calibration of FM Generators-Frequency and Amplitude

The FM interference situation is created by two separate FM generators. These generators have the advantage of having accurately calibrated attenuator controls. If one generator is set at any given median signal level, and the level of the second generator equalized to this median signal level, then the attenuator control of this second generator may be used to read _a_ directly. Since most of the tests will be conducted for zero carrier-frequency separation, it is necessary to equalize both the carrier frequencies of the generators as well as their amplitudes at some convenient median signal level.

A very precise method of equalizing frequency and amplitude simultaneously is one based on the properties of the demodulated instantaneous frequency variation of the resultant of two unmodulated co-channel carrier signals. The demodulated instantaneous frequency variation of the resultant of two unmodulated carriers in the same IF band is in the form of a periodic train of pulses. The repetition frequency of these pulses is the frequency of separation between the two co-channel carrier signals. The amplitude of the pulses is roughly inversely proportional to the difference in amplitude between the RF levels of these two carrier



FM Gen. No. 1 Boonton Type 202-B FM-AM Signal Gen. RLE No. 392-214 (1000~)

FM Gen. No. 2 Boonton Type 202-B RLE No. 839-884 (400~)

Univerter No. 1 Boonton Univerters RLE No. 392-300 (1000~)

Univerter No. 2 Boonton Univerters RLE No. 468-5112 (400~)

Oscilloscope Tektonic 515-A RLE No. 7522-243

Distortion Analyzer Hewlett Packard Type 330-C USN 7131-293

Wave Analyzer No. 1 General Radio Type 736-A RLE 392-128 (1000~)

Wave Analyzer No. 2 General Radio Type 736-A RLE 468-2000 (400~)

GR. Signal Gen. General Radio Type 805-C RLE 836-793

Fig. 7.1 Arrangement of System and Test Equipment signals.

With this in mind, each generator is connected to the input terminals of the IF amplifier that precedes the weaker-signal suppressor. The output of the weaker-signal suppressor is connected directly to the input of the demodulator. The modulation of each generator is turned off and their RF frequencies are approximately equalized according to the FM generator frequency dial. Each generator amplitude is adjusted so that the attenuator dials both read 10,000 micro volts. The generator used to align the system is used as the reference generator since its carrier frequency is the center frequency to which the entire system was previously aligned. The amplitude of this generator is also used as the reference amplitude level. With this arrangement, the output of the demodulator (with the 3 kc low pass filter switch in the off position) is displayed on a scope with a slow time-base internal sweep. The signal that is observed on the scope is the periodic train of pulses. The frequency of the second generator is "zeroed in" on the reference frequency by varying the univerter dial pertinent to the second FM generator. As the carrier frequency separation decreases, the period between successive pulses increases (repetition frequency of pulses decreases). When this period is a maximum (theoretically infinite), the carrier frequencies of the two signals are identical.

Next, the amplitude monitor control of the second FM generator is varied in order to produce the greatest pulse amplitude. At the point of greatest pulse amplitude, a slight turning of the amplitude monitor control in the same direction will cause the pulses to flip over, thus changing polarity. This phase reversal indicates that the amplitudes of the generators are identical and the amplitude monitor control of this generator is left in that position.

Both generators are now equalized both in amplitude and frequency. Due to drifting of both the amplitudes and the carrier frequencies of each generator, this calibration procedure must be repeated once every half hour.

Calibration of Measuring Test Equipment

The next calibration required is to calibrate each of the GR wave analyzers to their respective demodulated audio signals. Two wave analyzers are provided since it is necessary to measure the demodulated, fundamental audio component of both the stronger and weaker signal message simultaneously. The calibration of each wave analyzer is accomplished by the following procedure:

One generator at a time is connected directly to the input of the demodulator. The audio modulating frequency to frequency modulate this generator is selected and the frequency deviation of the FM signal adjusted to the desired amount. The demodulator output then is applied to both wave analyzers, scope and the distortion analyzer. One of the wave analyzers is balanced, calibrated and tuned to the demodulated audio signal and the reference level of this wave analyzer is then adjusted to 100 percent. This procedure must be repeated again for the other FM generator and wave analyzer.

Only one distortion analyzer was available, hence this instrument had to be used to measure the total distortion of both audio signals. Having to calibrate the distortion analyzer at each setting of a is a handicap since it is time consuming and prevents the taking of quick measurements. This allows time for the unavoidable drifting of the generators and the trap setting. In an effort to minimize the time required to calibrate the distortion analyzer at each setting of a, note was made during an initial balance of the instrument as to the location on the frequency dial of the audio frequency component that was to be measured. Thus, although the reference level had to be recalibrated at each setting of a, it was then only necessary to turn the distortion analyzer frequency dial to the noted frequency position to quickly obtain the distortion reading. Fortunately, the instrument only required an initial balancing which was carried out at the outset of each run. This procedure, although still time consuming, minimized to some degree the time required for a complete run.

A second painstaking, time consuming operation was the necessity of having to vary the fine tuning dial of the initially precalibrated wave analyzers. The reason for this is that the audio modulating frequencies of both generators and the filters in both wave analyzers had a tendency to drift. The fact that these instruments drifted in their tuning can account for some of the anomalies encountered in the procurement of the data. However, if the curves plotted from the data exhibited an unusual trend in the response of the system, the test was usually repeated.

Final Adjustment of Trap

Prior to actually making a test run, the trap had to be preset in order to suppress the stronger signal. This was best accomplished by first removing one signal (the weaker signal) from the input and then lowering the amplitude slightly of the remaining FM signal. The trap position and attenuation was then tuned while observing the demodulator output. Theoretically, if the trap attenuation was sufficient, there should have been no output from the demodulator. However, there was some finite trap transmission and the demodulator threshold was low enough so that there was still some audio signal in the output but badly distorted by noise. When maximum distortion was observed, this indicated that the trap position and attenuation was set to suppress the stronger signal.

An alternate method (the method used in the following tests) was to apply both FM signals to the input with their RF amplitude levels preset as described. Then the weaker signal was reduced in amplitude as far as possible ($a \approx .01$) while observing the demodulated output. The trap was then set for minimum distortion of this demodulated weaker signal. At this relative interference ratio of $a \approx .01$ the trap tuning was very sensitive requiring that this procedure be repeated in steps of reduction of a.

Method Used to Vary Interference Ratio

One last point concerning the experimental procedure was the manner in which the amplitude interference-ratio a was varied.

If the reader will consult any one of the curves for which the weaker-signal suppressors were used, he will note that the range of a for which each test was performed was $.001 \leq a \leq 100$. In all of these tests, when $a \leq 1$, the weaker (desired) signal was captured and when $a > 1$ this same signal became the stronger (undesired) signal whence it was suppressed. The point which we wish to call attention to is that the input median signal level is always determined by the stronger signal. Thus for $a \leq 1$, the median signal level, E_s, was 10,000 micro volts. However, for $a \geq 1$ the median signal level was $a\ E_s = a$ (10,000) micro-volts if the same generator was used to vary the interference ratio a. When $a = 100$, this median signal level became 1 volt. The difficulty here is twofold. First, a one volt signal is not available from the FM generators. Secondly, even if we had a 1 volt signal, this magnitude of signal would certainly overdrive the IF amplifiers preceding the weaker-signal-suppressors. This difficulty was avoided by keeping the median signal level fixed. To ensure this condition of constant median signal level, the generator that provided the weaker desired signal was always the generator whose amplitude was varied. Thus for $a \leq 1$, the generator that provided weaker signal was used to vary a by lowering its amplitude while the generator that provided the stronger signal was fixed at $E_s = 10{,}000$ micro-volts. For $a \geq 1$, the generator that provided the previously weaker signal (but now the stronger) was fixed at $E_s = 10{,}000$ micro volts while the generator that provided the previously stronger signal (but now the weaker) was used to vary a by lowering its amplitude. This is the manner in which the complete range of a was covered keeping the median signal level fixed. We were then assured that the amplifier was never over-driven and also that the limiters were driven with a constant median signal level.

Summary of Experimental Procedure

To conclude this chapter a summary will be given of the experimental procedure followed prior to and during the accumulation of data for each test run. One test run takes roughly 30 to 45 minutes. The actual time consumed in accumulating the data varies from 20 to 30 minutes during which time the trap does not drift significantly so that the data is indicative of the system response. A quick check of the validity of the data accumulated for one complete test run was made after the completion of a run by rechecking the response at the particular a for which the trap setting was optimized.

The steps and time necessary for each run are as follows:

15 Minutes

1 - Balance, calibrate, tune, and set level of each wave analyzer to their respective audio signals.

2 - Initially balance each distortion analyzer.

3 - Check center frequency of reference generator by centering the filter response of the IF preamplifier in the center of the scope sweep (scope set up to view characteristics of this filter)

4 - Connect system for equalization of the FM generator carrier frequencies and RF amplitudes.

5 - Reconnect system properly.

6 - Adjust trap position and attenuation for optimum response for an a $=0.01$.

15 - 30 Minutes

7 - Set amplitude interference ratio to desired amount.

8 - Peak both wave analyzers readings.

9 - Set level of distortion analyzer and obtain total distortion for both fundamental audio components.

10 - Repeat steps 6, 7, 8 until complete range of a has been covered ($.001 \leq a \leq 100.$)

11 - Recheck trap setting for a $=0.01$.

PERFORMANCE TESTS OF WEAKER-SIGNAL RECEIVER

The capture characteristics of the fixed-trap weaker-signal receiver with photographs of typical output waveforms are presented in Fig. 8.2 through Fig. 8.33. These capture characteristics fall into the following three categories:

1 - Those intended to display the performance of the system for the condition of simulated predetection of the stronger signal, Fig. 8.2 through Fig. 8.7. These capture characteristics were obtained for various modulation conditions of interference and desired signal. They are intended to emphasize the improvement in the response of the redesigned fixed-trap circuit and the necessity of providing delay equalization.

2 - Those intended to display the performance of the system for unsimulated conditions. These capture characteristics shown in Fig. 8.8 through Fig. 8.17 were obtained using the full degree of weaker signal suppression in conjunction with a delay equalizer. They are intended to display the system performance for various modulation conditions of interference and desired signal.

3 - Those intended to display the dependence of the performance of the system upon the degree of weaker signal suppression required and upon the degree of delay compensation required. These capture characteristics shown in Fig. 8.18 through Fig. 8.33 were all obtained for the same modulation conditions of interference and desired signal.

Performance of Fixed Trap System for Simulated Conditions

The first series of tests were conducted by simulating predetection of the stronger signal. A block diagram showing the manner in which this system was connected for the simulated conditions is shown in Fig. 8.1. These are the same conditions assumed by Sheftman in his study of the fixed trap system. A summary of his results appear in Chapter II.

Because of the simulated condition of predetection of the stronger signal, the fixed trap system will capture the weaker signal for all $\underline{a} > \delta/\rho_2$ where δ is the trap transmission at center frequency and ρ_2 is the capture ratio of the trap-driven demodulator. The region of principal interest in these capture plots is the capture transition region centered about $\underline{a} = \delta$. We are particularly interested in the uniformity of response of the fixed-trap system near the capture transition region for all conditions of modulation of interference and desired signal. It will be recalled from Chapter II that the capture transition region for the response of Sheftman's system was to a large extent effected by the modulating properties of the stronger signal message. When either the deviation or the modulating frequency of the stronger signal message was increased, the response of his system became impaired.

Each capture characteristic shown in Fig. 8.2 through Fig. 8.6 was obtained for different modulation conditions of the interference and the desired signal messages. From these curves we note that all the capture plots have transition regions that are centered about interference-ratios in the range $0.0033 \leq \underline{a} \leq 0.0042$. The response of Sheftman's fixed-trap system exhibited capture transition

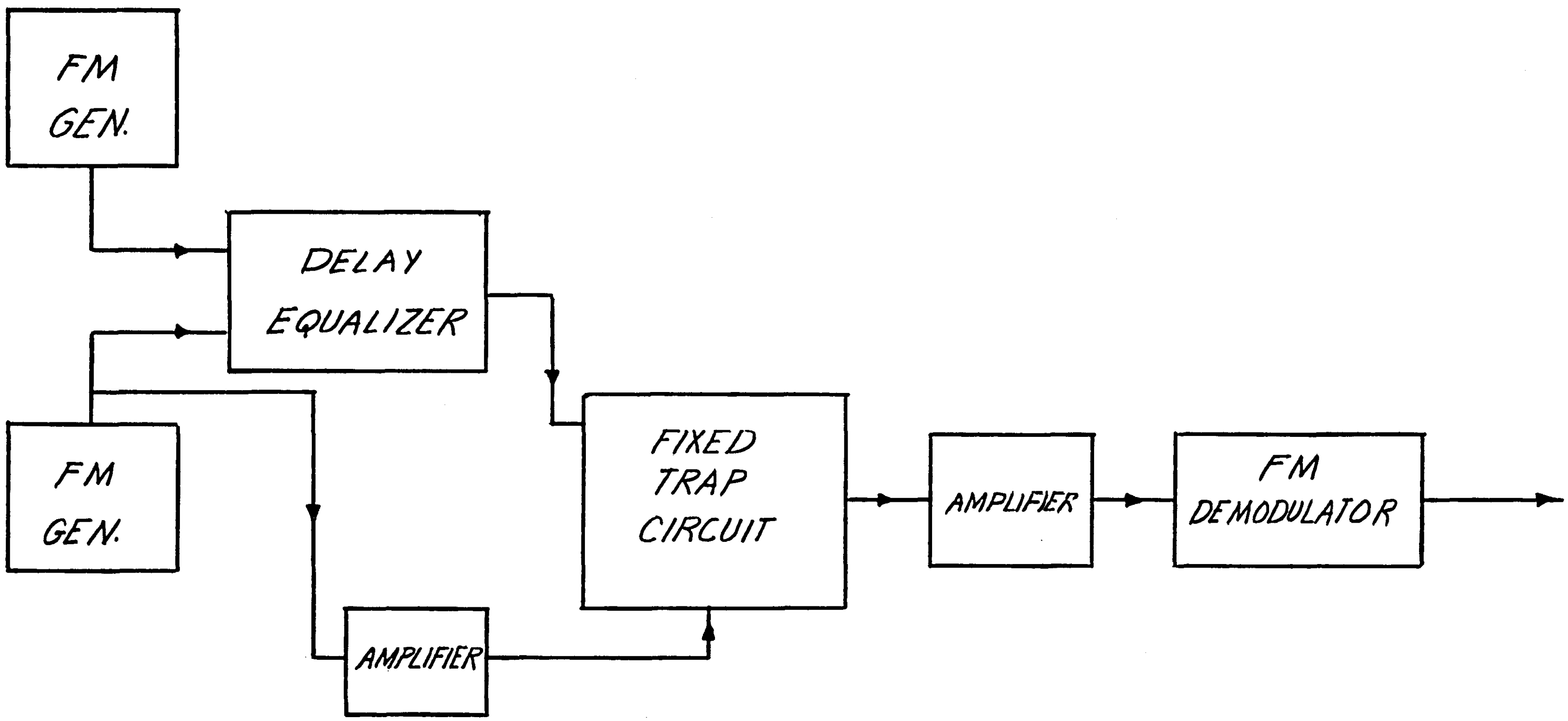


Fig 8.1, Fixed Trap System Set Up For Simulation of Predetection of Stronger Signal

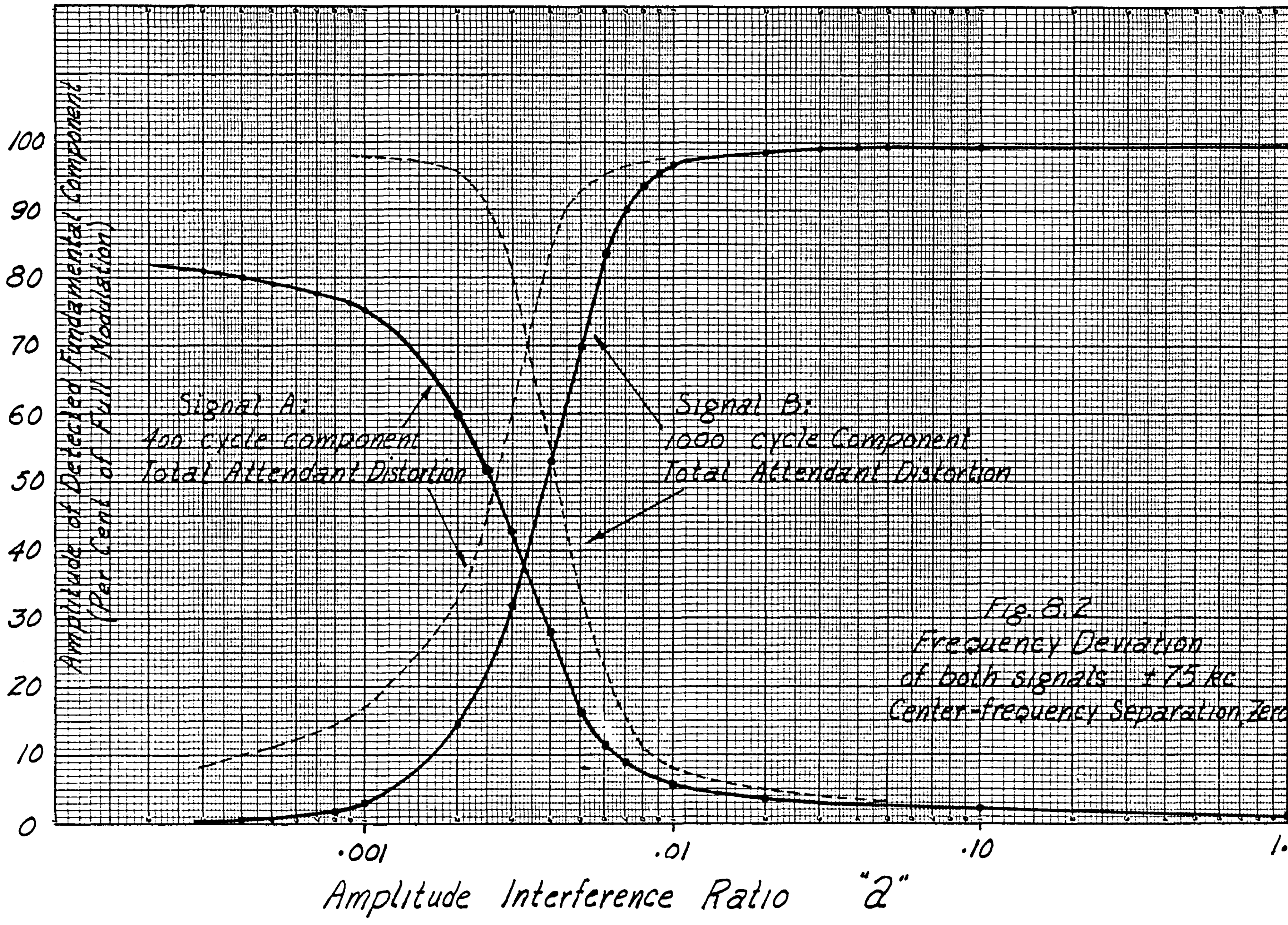


Fig. 8.2
Frequency Deviation
of both signals ±75 kc
Center-frequency Separation Zero

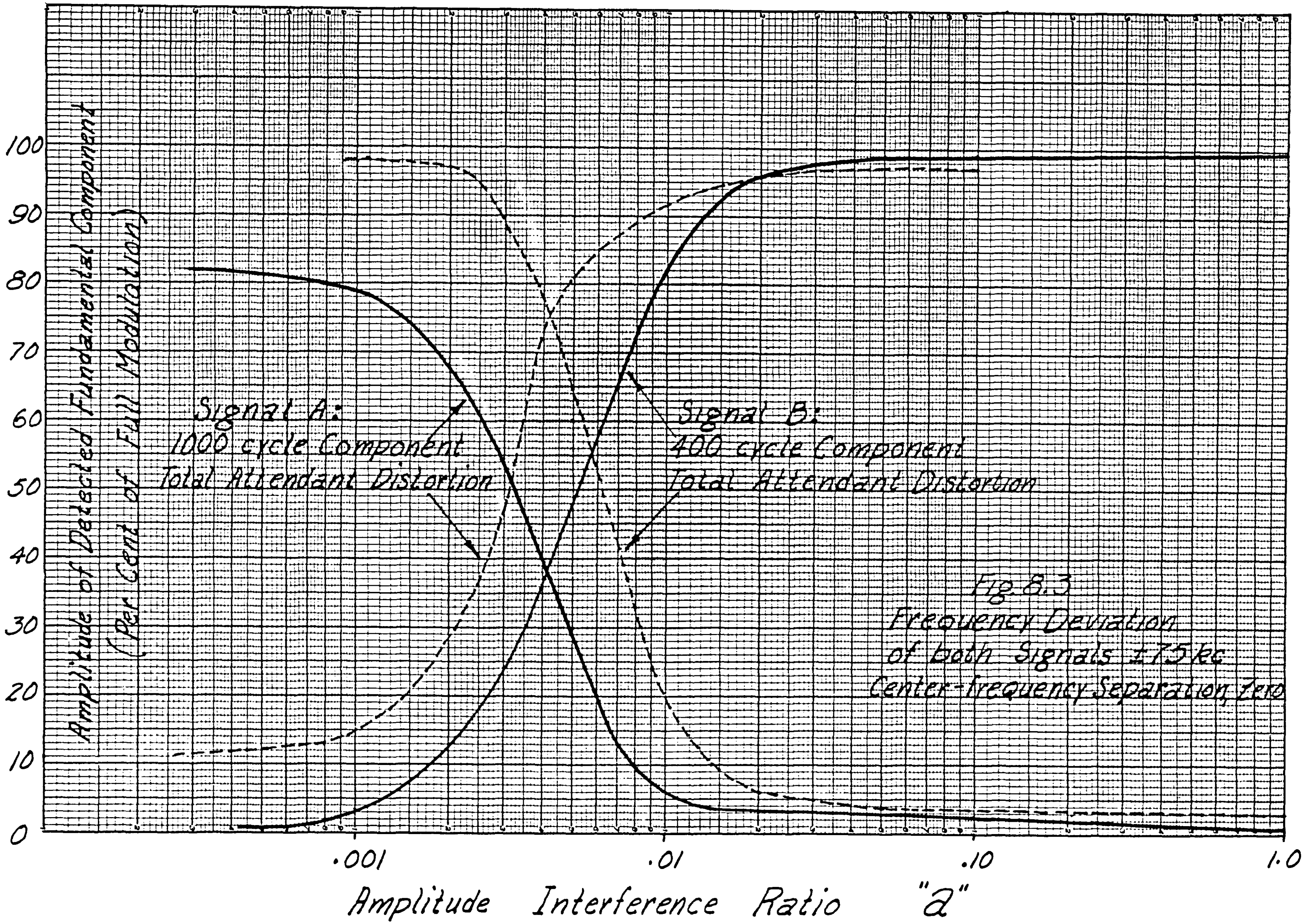
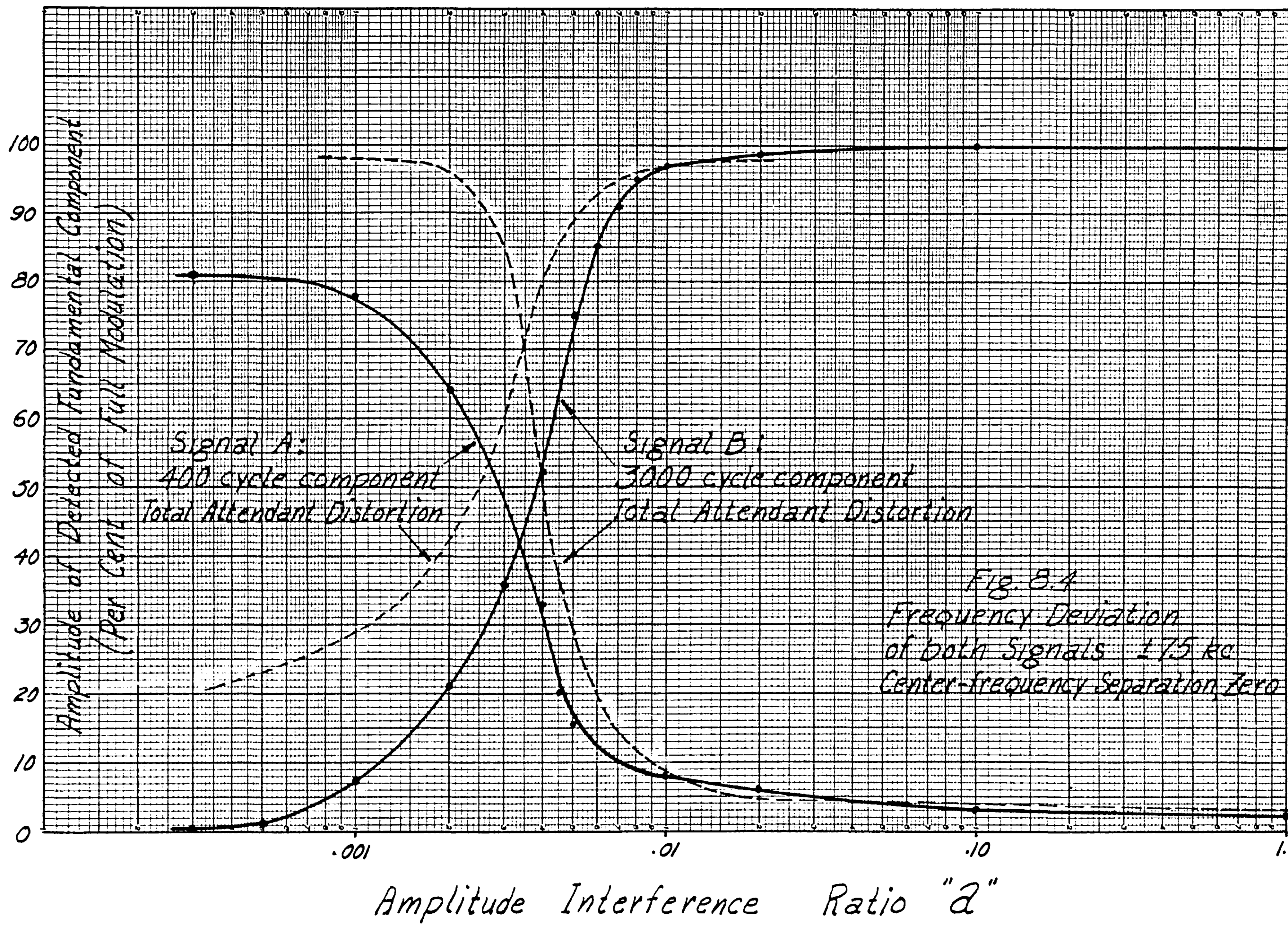
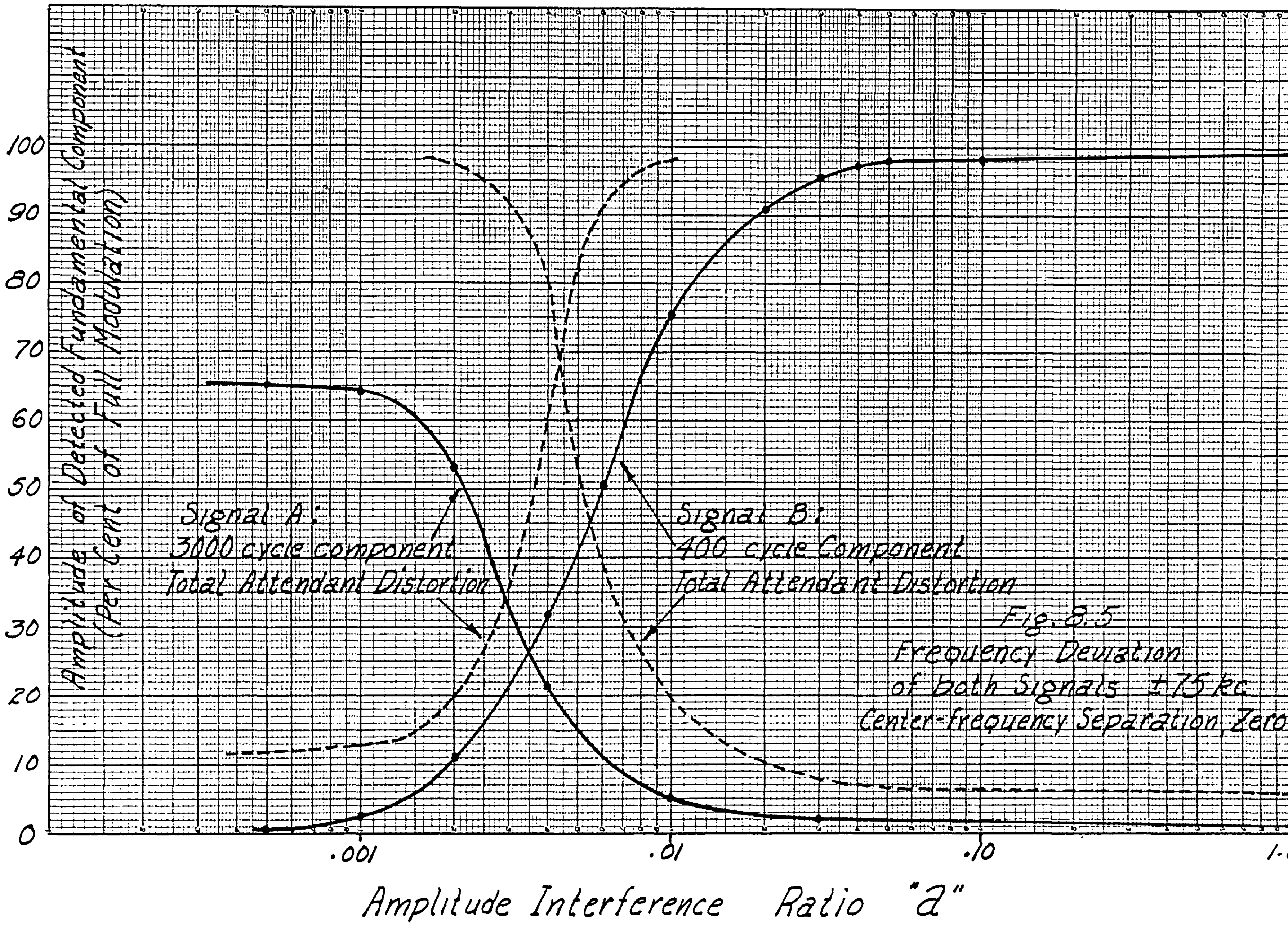


Fig 8.3
Frequency Deviation
of both Signals ±75 kc
Center-frequency Separation, Zero



Amplitude of Detected Fundamental Component
(Per Cent of Full Modulation)
100
90
80
70
60
50
40
30
20
10
0
.001
.01
.10
1.0
Amplitude Interference Ratio "a"
Signal A:
400 cycle component
Total Attendant Distortion
Signal B:
3000 cycle component
Total Attendant Distortion
Fig. 8.4
Frequency Deviation
of both Signals ±75 kc
Center-frequency Separation Zero


Amplitude of Detected Fundamental Component (Per Cent of Full Modulation)
100
90
80
70
60
50
40
30
20
10
0
.001
.01
.10
1.0
Amplitude Interference Ratio "a"
Signal A:
3000 cycle component
Total Attendant Distortion
Signal B:
400 cycle Component
Total Attendant Distortion
Fig. 8.5
Frequency Deviation
of both Signals ±75 kc
Center-frequency Separation Zero

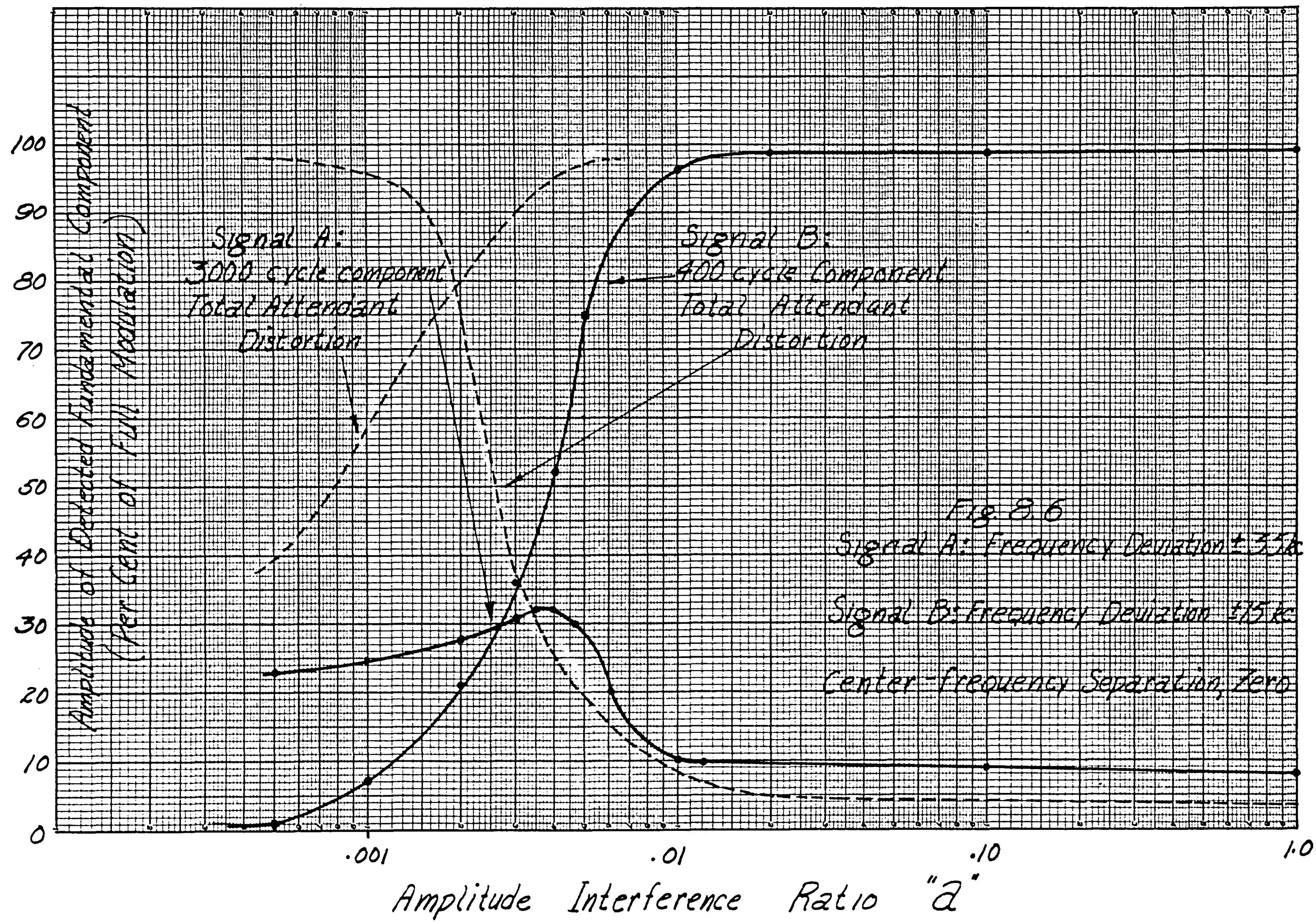
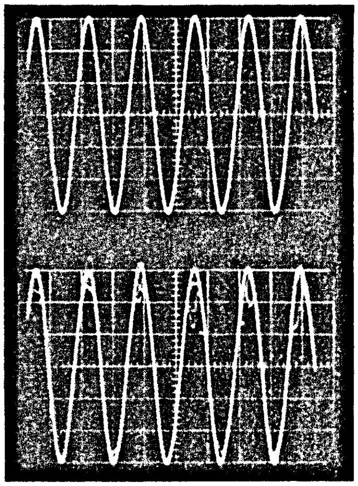


Fig. 8.6
Signal A: Frequency Deviation ±3.5kc
Signal B: Frequency Deviation ±7.5kc
Center-Frequency Separation, Zero

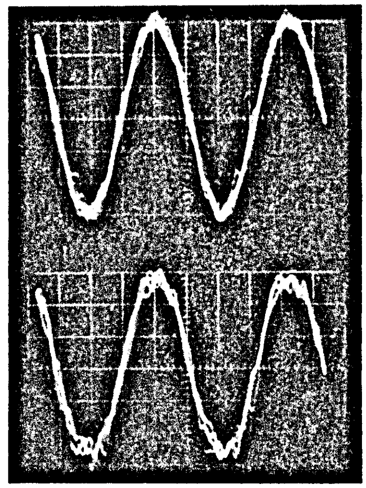
a = 0.10
Fund. = 99%
Dist. = 2.8%

a = 0.01
Fund. = 97%
Dist. = 8.0%



Weak Sig. Mod.
75 kc dev.
1000 cps mod.freq.

Strong Sig. Mod.
75 kc dev.
400 cps mod.freq.



a = 0.10
Fund. = 98%
Dist. = 6%

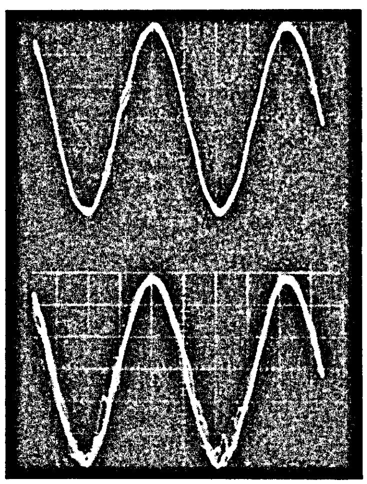
a = 0.01
Fund. = 76%
Dist. = 19%

Weak Sig. Mod.
75 kc dev.
400 cps mod.freq.

Strong Sig. Mod.
75 kc dev.
3000 cps mod.freq.

a = 0.10
Fund. = 99%
Dist. = 4%

a = 0.01
Fund. = 96%
Dist. = 8%



Weak Sig. Mod.
75 kc dev.
400 cps mod.freq.

Strong Sig. Mod.
35 kc dev.
3000 cps mod.freq.

Fig. 8.7 Demodulated Weaker Signal Message Waveforms Predetection of Stronger Signal Simulated regions centered about higher interference-ratios which varied over a wider range of a, $0.0135 \leq a \leq 0.069$. The present fixed trap system indicates three improvements in response; a capture transition region centered at a lower a and a variation of the capture transition region that is restricted to a much smaller range of a. Secondly, we note that the interference ratio for which the distortion rises to 10% is below $a = 0.02$ for all the capture plots. Lastly, we note that although there is still some broadening of the capture transition region, this broadening effect is restricted to a smaller range of interference ratios.

Thus, we see that the response of the improved fixed-trap system does not vary excessively with the stronger signal modulation. However, the same trends of variation in response as those exhibited by Sheftman's system still exist. These trends may be observed by comparing the following pairs of capture plots: Fig. 8.2 with Fig. 8.3, Fig. 8.4 with Fig. 8.5, and Fig. 8.3 with Fig. 8.5. In each pair of capture plots, the modulating frequency of the undesired signal was increased. In comparing each figure of a pair we note a slight shift of the capture transition region to larger a and a slight broadening of the capture transition region accompanied by a premature increase in the capture of the undesired signal.

The reason for this trend is that apparently there is still some error in the compensation provided by the delay equalizer. The slope or the non-linearties of the phase characteristic for the delay equalizer are probably not matched exactly to those of the converter channel filters.

If comparison is made of the capture plots of Fig. 8.5 and Fig. 8.6 we note a second trend similar to that experienced by Sheftman. In Fig. 8.5 the stronger signal modulation deviation is $\pm$ 75 kc while in Fig. 8.6 it was reduced to $\pm$ 35 kc. We note in Fig. 8.6 the improvement of weaker signal capture over that in Fig. 8.5. For the conditions of smaller deviations of the stronger signal, the capture ratio of the weaker signal receiver shifted to a lower interference ratio. There are two possible reasons for this improvement. First, for smaller deviations of the stronger signal modulation, the stronger signal is confined to the more linear portions of the phase characteristics of the equalizer and the converter channel filters. Thus the correlation between the stronger signal modulation appearing at both inputs of the first mixer is improved providing for more complete cancellation of the frequency modulation of the stronger signal. Second, the stronger signal sweeps over a narrower, flatter portion of the amplitude characteristic of the equalizer filter and of the converter channel filters. This results in considerable reduction of incidental amplitude modulation of the envelope of the stronger FM signal. The percentage AM modulation of the carrier frequency component is reduced resulting in much smaller AM sidebands. According to the discussion presented in Chapter III, a reduction in the incidental AM prior to trapping effectively gives the appearance of a greater trap attenuation because of the reduced significance of the AM sidebands. As a result, the fixed trap system showed an improvement in response when the deviation of the undesired stronger signal modulation was decreased.

Certain anomaies regarding the capture of the stronger signal below the capture transition (for $\underline{a} < \delta$) are evident from these capture plots. The first of these is that in all of these tests, the percent fundamental of the stronger signal never exceeded 85% of its fully modulated value even for very low $\underline{a}$ ($\underline{a} < 0.001$). For $\underline{a}$ in this range the percent capture of the weaker signal is below 5% in all of the plots. Since the weaker signal is far below the level of the stronger signal, we may assume that there is only one signal present in the channel. If we reflect upon the basic theory of the fixed trap system, we see that when only one signal is present in the channel, this signal is attenuated by the trap. The trap attenuation may be sufficient to suppress the signal below the demodulator threshold resulting in a noise distorted captured signal. The effectiveness of the trap in accomplishing this one signal cancellation is further manifested when the stronger signal is only frequency deviated to $\pm$ 35 kc. For reasons given regarding the reduction in the AM developed when smaller deviated signals are used the trap attenuation seems to be more effective, resulting in further suppression of the one FM signal present in the channel.

This last phenomenon is a basic limitation of all of the trapping systems since removal of the undesired signal would also result in a loss of the desired signal.

Performance of Fixed-Trap System for Unsimulated Conditions

Weaker-Signal Suppression Fixed, Modulations Varied

The next series of capture characteristics are presented in Fig. 8.9 through Fig. 8.18 with photographs of typical output waveforms following each set of curves. All these capture characteristics were obtained utilizing four cascaded narrow-band limiters as weaker-signal suppressors. The block diagram of the system is shown in Fig. 7.1. The narrow band limiters were designed to handle a maximum amplitude interference ratio of $\underline{a} = 0.867$. The bandwidths of each of the filters following each limiter was set to one IF bandwidth (3 db BW $\approx$ 200 kc).

All these capture characteristics were obtained for a fixed degree of weaker signal suppression; however, the characteristics were obtained for various magnitudes of frequency deviation of both the desired and undesired signal modulations. The modulating frequencies of the FM signals in all the capture characteristics are fixed at 1000 cps and 400 cps.

In all the capture characteristics, a set of curves are plotted describing the capture characteristics of the system for the modulation conditions cited in the upper right hand corner of each plot. The curves plotted are as follows:

1 - Capture of the 1000 cycle fundamental component in percent of full modulation.

2 - Total attendant distortion of 1000 cycle component in percent of resultant audio signal level (after 3 kc of low-pass filtering).

3 - Capture of the 400 cycle fundamental component in percent of full modulation.

4 - Total attendant distortion of 400 cycle component in percent of <u>resultant</u> audio signal level (after 3 kc of low-pass filtering).

Each plot is described by an approximately symetrical two sided-capture characteristic centered about $\underline{a} = 1$. The reason for this two-sided capture characteristic is that the role of desired and undesired signal interchanges when the amplitude interference ratio is varied through unity. For instance, for $\underline{a} < 1$, the 1000 cycle component is the desired signal while the 400 cycle component is the undesired signal. For $\underline{a} > 1$, the 1000 cycle component now becomes the undesired signal and the 400 cycle component becomes the desired signal. The amplitude interference ratio for <u>these</u> tests is <u>always</u> determined by the ratio of the 1000 cycle modulated FM signal to the 400 cycle modulated FM signal. Thus for $\underline{a} > 1$, the amplitude interference ratio for the weaker 400 cycle signal is actually $1/\underline{a}$; however, $\underline{a}$ is the ratio plotted on the logarithmic abscissa. In addition to the two-sidedness feature, each capture characteristic contains three capture transition regions centered about $\underline{a} \approx 0.01$, $\underline{a} = 1.0$, $\underline{a} \approx 100$ respectively. The capture transition regions at the extremities of the characteristics ($\underline{a} = 0.01$, $a = 100$) are affected by the trap attenuation characteristics, δ, as described for the response of the fixed trap system under simulated conditions. From the results we will see that the capture transition region centered about $\underline{a} = 1$ is affected by the weaker-signal suppressor the fixed-trap circuit, and the demodulator.

In this system, the weak signal suppressor aids capture only of the signal that is the weaker of the two co-channel FM signals. Whenever the amplitude ratio $\underline{a}$ approaches unity from either side, the suppression properties of the limiters begin to fail, and the response of the entire system is effected. At an $\underline{a} = 1$, the weaker-signal suppressor cannot decide which of the two co-channel signals is the desired signal and suppression of neither signal is achieved. Under these conditions, the output of the fixed trap can aid neither signal, thus delivering to the input of the demodulator two signals whose RF amplitudes are still equal. As we move away from $\underline{a} = 1$ to $\underline{a} = 0.9$ or $\underline{a} = 1.1$, the suppression properties of the limiters improve enabling the trap to suppress the amplitude of the stronger of the two co-channel signals below the level of the weaker signal. The stronger signal suppression of the fixed trap is sufficient to permit capture of the desired weaker signal by the demodulator although significant distortion prevails.

Thus we see qualitatively that the capture transition region centered about $\underline{a} = 1$ is effected by the following:

1 - The ability of the weak signal suppressors to provide significant suppression of the weaker signal.

2 - The ability of the fixed-trap performing under conditions of partial weaker signal suppression to provide significant suppression of the stronger signal.

3 - The stronger-signal capture characteristics of the de-

modulator.

Before actually considering the capture characteristics for the various modulation conditions, we can make some predictions concerning the performance based on heuristic reasoning deduced from the discussion presented in Chapter II of the response that can be expected from a fixed trap system used in conjunction with narrow band limiters that provide for the initial weaker signal suppression.

First of all, the effectiveness of the trap in suppressing the stronger signal depends upon the ability of the narrow band limiters to suppress the effects of the weaker signal with respect to the stronger signal. The degree of weaker signal suppression that is required by the fixed trap will be considered later. For the moment, we are concerned with the suppression effectiveness of the limiters for various modulation conditions of the desired and undesired signal. From the previous discussion, the two parameters of the input signals that have a direct bearing on the effectiveness of the limiters in suppressing interference are the amplitude ratio, $\underline{a}$, and the difference in instantaneous frequency, r, between the two FM input signals. Let us consider a hypothetical situation where r is fixed but $\underline{a}$ is allowed to vary. In particular, we would like to consider situations where $\underline{a}$ lies in the range $0.5 \leq \underline{a} \leq 1$. From Baghdady's theory of the limiter (6) we note that as $\underline{a}$ approaches unity, the interference suppression properties of the limiters diminish (see Fig. 20 Ref. No. 6). Thus, for $\underline{a}$ approaching unity the weaker signal appearing in the output of the limiters is still significant and we can expect poor response. As a diminishes ($a \leq 0.5$) the interference suppression properties of the limiters improve, and we can expect an improvement in the response.

Now let us review the numerical analysis presented in Chapter II where we fixed a but let r vary. When the instantaneous frequency of the two co-channel FM signals is such that r is a small fraction of the total limiter BW, then a greater number of spectral components of the limited sprectrum of the resultant signal will still exist in the output of the limiter after filtering and we can expect that the response of the system will be poor. As r is increased so that it becomes a larger fraction of the limiter BW, more components of the limited spectrum are filterd and the response of the system should improve.

In order to show this dependence of the response of the receiver upon a and r, the receiver was tested for conditions of interference where the stronger interfering signal was unmodulated. Three tests were conducted in an effort to show this dependence. For each test the unmodulated stronger interfering signal was located at a different position in the IF pass band. These positions in frequency are as follows:

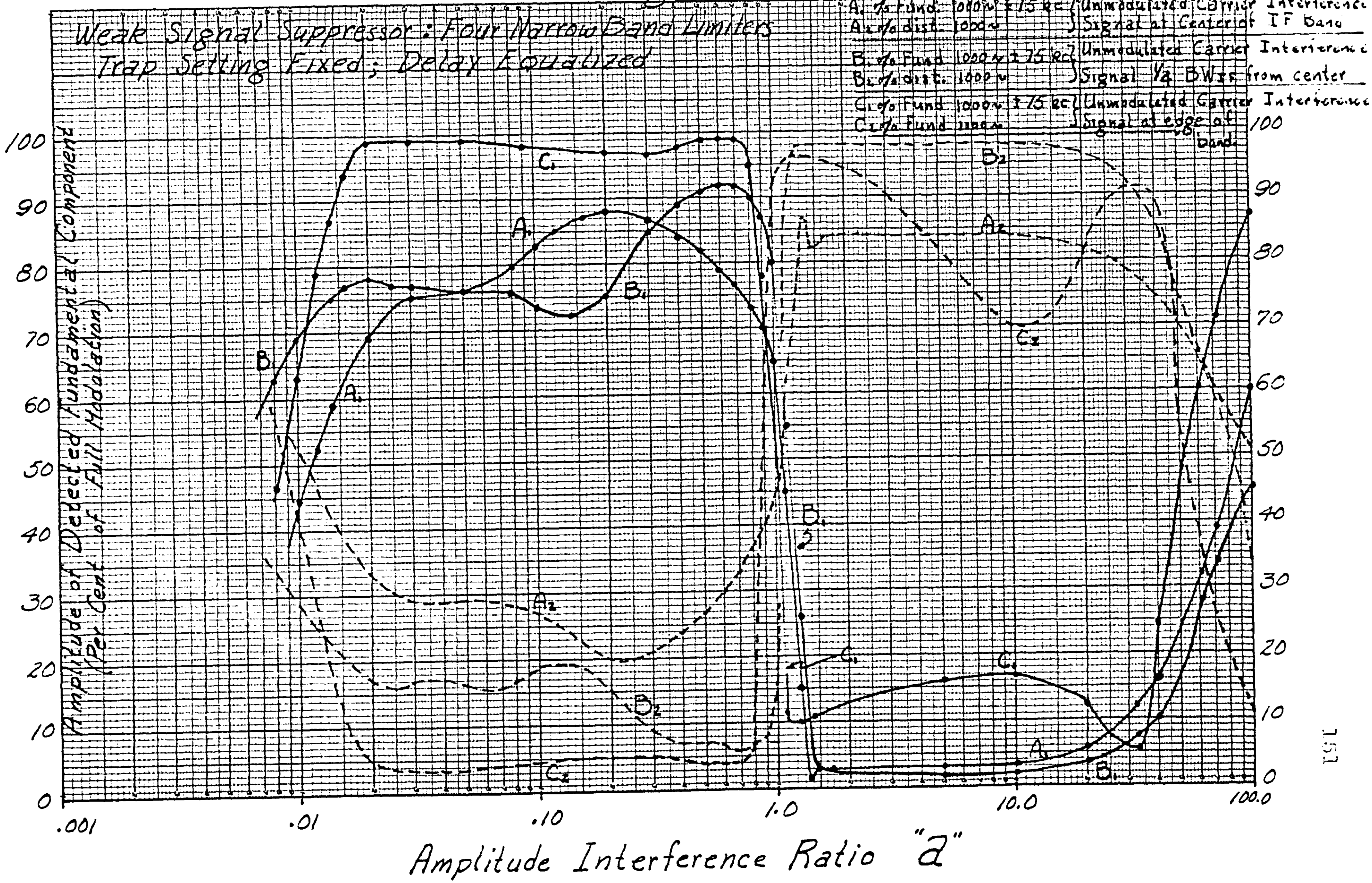
1 - Center of IF pass band.

2 - Halfway between center and edge of IF pass band (35 kc from center)

3 - Extreme edge of IF pass band (75 kc from center)

In all these tests the weaker signal was modulated by an audio frequency of 1000 cycles to a frequency deviation of $\pm$ 75 kc. Fig. 8.9 shows the capture characteristics and Fig. 8.10 shows some

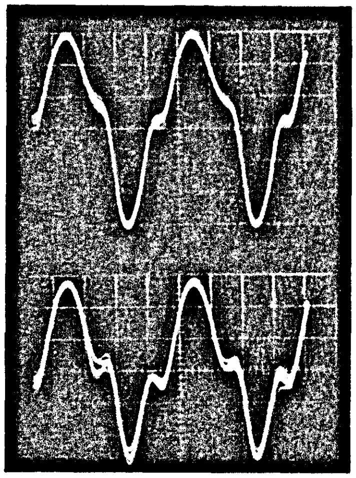
Fig. 8.9


Weak Signal Suppressor: Four Narrow Band Limiters
Trap Setting Fixed; Delay Equalized
A1 % Fund. 1000~ ± 75 kc } Unmodulated Carrier Interference
A2 % dist. 1000~ } Signal at Center of IF band
B1 % Fund 1000~ ± 75 kc } Unmodulated Carrier Interference
B2 % dist. 1000~ } Signal 1/4 BW IF from center
C1 % Fund 1000~ ± 75 kc } Unmodulated Carrier Interference
C2 % Fund 1000~ } Signal at edge of band.
Amplitude of Detected Fundamental Component (Per Cent of Full Modulation)
Amplitude Interference Ratio "a"
.001
.01
.10
1.0
10.0
100.0
0
10
20
30
40
50
60
70
80
90
100
A1
A2
B1
B2
C1
C2

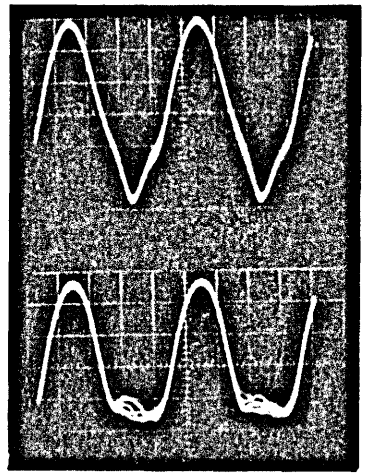
151

a = 0.2
Fund. = 88%
Dist. = 20%

a = 0.03
Fund. = 75%
Dist. = 26%



Carrier at
Center of IF



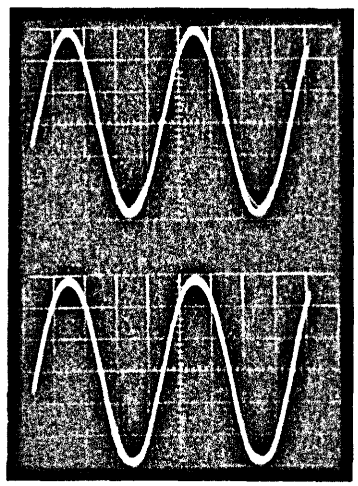
a = 0.5
Fund. = 91%
Dist. = 7%

a = 0.014
Fund. = 75%
Dist. = 18%

Carrier 36 kc
from Center of IF

a = 0.1
Fund. = 98%
Dist. = 4%

a = 0.02
Fund. = 99%
Dist. = 4%



Carrier at
Center of IF

Weaker Signal Modulation
75 kc. deviation
1000 cps. mod. freq.

Stronger Signal Modulation
Unmodulated Carrier

Fig. 8.10 DEMODULATED WEAKER SIGNAL MESSAGE WAVEFORMS

photographs taken of the demodulated weaker signal.

We note from the capture characteristics and the photographs that the poorest response occured when the unmodulated interfering stronger signal was centered in the IF pass band. The best response occured when the unmodulated interfering signal was at the edge of the IF pass band. When the unmodulated interfering signal was halfway between the center and the edge of the IF pass band we note some improvement in the capture characteristic and in the distortion characteristic as compared to the case where the interfering signal was centered in the IF band. The reason for this improvement in the response as the unmodulated interfering signal was moved towards the edge of the band, is that the instantaneous frequency separation, r, was larger with respect to the limiter bandwidth for a larger fraction of the modulation cycle of the desired signal. Since r was larger with respect to the limiter bandwidth, the suppression afforded by the limiters was more effective over a larger fraction of the modulation cycle. By virtue of the more effective suppression of the weaker signal in the limiter stages, the ability of the fixed-trap to suppress the the stronger signal was enhanced resulting in improved capture of the weaker signal with lower attendant distortion (below 5% for curve C).

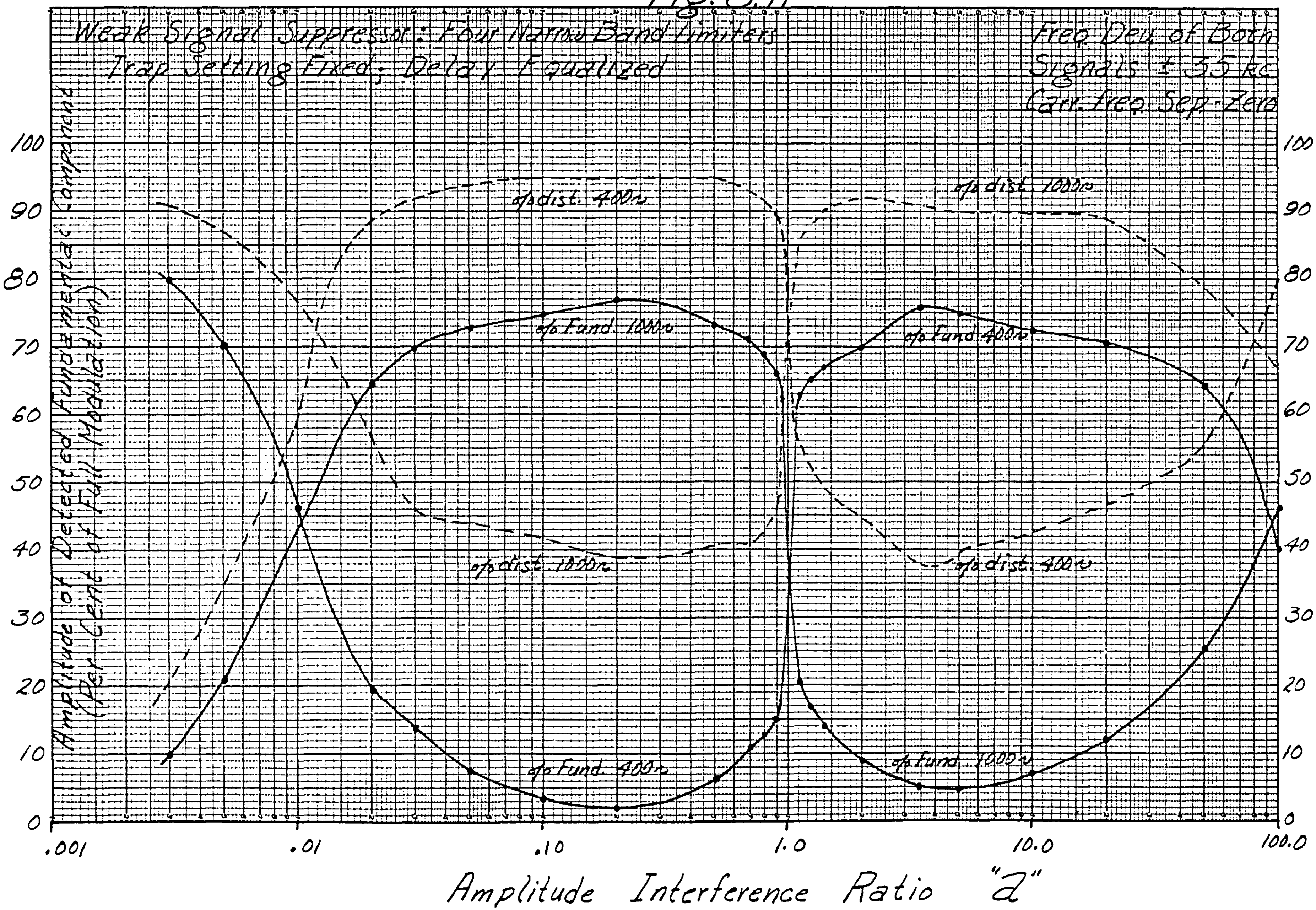
In comparing curve A_1 with curve B_1 some anomalies are evident. We note first that curve B_1 peaks sooner than curve A_1 but there it drops below the capture curve A_1 by 15% at $\underline{a} = 0.15$ where the capture curve A_1 is nearest its peak. This was not expected since according to the heuristic reasoning given earlier regarding the instantaneous frequency separation r, we should expect an improvement in the response for all a when the unmodulated interfering signal was displaced from the center of the IF pass band. Apparently, the suppression of the weaker signal by the limiters may not have improved significantly when the interfering signal was shifted by only 35 kc. This possible lack of suppression by the limiters handicaped the trap and the interference ratio at the demodulator was the same or in some instances, lower than that which resulted when the unmodulated interfering signal was centered in the IF pass band. In addition, from a comparison of the oscillograms, it appears that when the interfering carrier was displaced slightly from center, harmonics resulted that were more detrimental to the magnitude of the detected fundamental. The oscillograms for the case where the interfering carrier is centered, indicates a predominant third harmonic component which appears to augment the peaks of the resultant waveform. For the case where the interfering signal was displaced by 35 kc from the center of the IF pass band the harmonic content is such that the resultant waveform is clipped on the lower peaks, (although the percent total distortion was lower). This later phenomenom could have more of an effect on the fundamental which could account for the reduction of detected fundamental component for this case.

Near the capture transition region centered at $a = 1$, the response in general is what it should be. We note the gradual increase in the weaker signal capture for the case where the unmodulated interfering signal is at the center of the band. The capture transition region for the other two cases where the unmodulated interfering signal is displaced from the center of the IF pass band is much more abrupt. The reason for this is that for larger a, the suppression properties of the limiters were significantly better than when the interfering signal was at the center of the IF pass band.

Some peculiar anomalies exist for $a > 1$ especially for the case where the unmodulated interfering signal is at the edge of the IF pass band. However, this region of the curves is not of particular interest since we are unaware of what happens to the weaker unmodulated component. Of some interest though, is the effectivenenss of suppression of the modulated component when the interfering carrier is at the center of the IF pass band or displaced from the center by 35 kc. This was not expected because of the relatively poorer response when it was desired to capture these signals for $a < 1$.

The next capture characteristic is shown in Fig. 8.11 accompanied by photographs of typical output waveforms in Fig. 8.12. The frequency deviation of each signal is now $\pm$ 35 kc while the modulating frequencies are 1000 cycles and 400 cycles. The carrier frequency separation is zero. This capture characteristic is indicative of the poorest response of the fixed-trap weaker-signal receiver. Detection of the fundamental components amounted to only 75% of full modulation accompanied by 40% total attendant distortion after 3 kc of low-pass filtering. The reason for this poor

Fig. 8.11

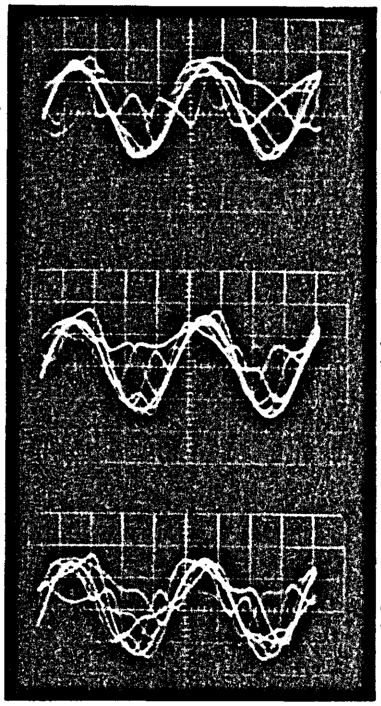


Weak Signal Suppressor: Four Narrow Band Limiters
Trap Setting Fixed; Delay Equalized
Freq. Dev. of Both Signals ± 35 kc
Carr. freq. Sep. Zero
Amplitude of Detected Fundamental Component (Per Cent of Full Modulation)
% dist. 400~
% dist. 1000~
% Fund. 1000~
% Fund. 400~
% dist. 1000~
% dist. 400~
% Fund. 400~
% Fund. 1000~
0
10
20
30
40
50
60
70
80
90
100
.001
.01
.10
1.0
10.0
100.0
Amplitude Interference Ratio "a"

Pic. No. 1

Pic. No. 2

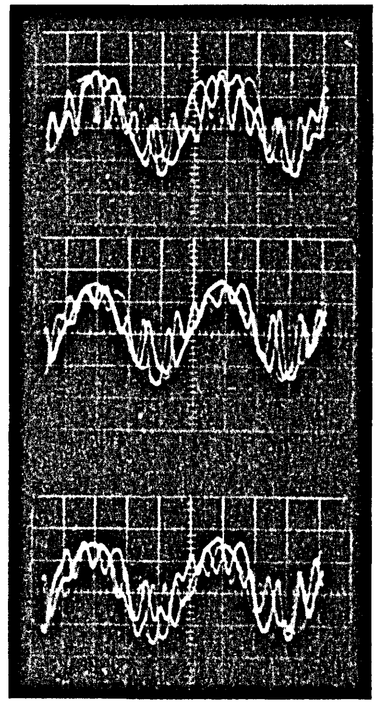
Pic. No. 3



Pic. No. 4

Pic. No. 5

Pic. No. 6



Pic. No. 1	$\underline{a} = 0.8$ % Fund. = 69% % Dist. = 43%	Pic. No. 4	$\underline{a} = 0.8$ % Fund. = 65% % Dist. = 52%
Pic. No. 2	$\underline{a} = 0.1$ % Fund. = 75% % Dist. = 42%	Pic. No. 5	$\underline{a} = 0.1$ % Fund. = 73% % Dist. = 43%
Pic. No. 3	$\underline{a} = 0.03$ % Fund. = 70% % Dist. = 46%	Pic. No. 6	$\underline{a} = 0.05$ % Fund. = 71% % Dist. = 47%

Weaker Signal Modulation
±35 kc. deviation
1000 cps. mod. freq.

Stronger Signal Modulation
±35 kc. deviation
400 cps. mod. freq.

Weaker Signal Modulation
±35 kc. deviation
400 cps. mod. freq.

Stronger Signal Modulation
±35 kc. deviation
1000 cps. mod. freq.

Fig. 8.12 DEMODULATED WEAKER SIGNAL MESSAGE WAVEFORMS

response is that the separation frequency difference, r, between the instantaneous frequencies of both signals is at maximum, half a limiter bandwidth. Effective suppression of the weaker signal in the limiters is not realized and as a result the trap is handicapped in suppressing the effect of the stronger signal. However, we do note that the fundamental component of the stronger signal modulation is suppressed below 10% but yet we still have an unaccountable excess of distortion. In order to see why we have this distortion let us re-examine the analysis of the trap for the condition of incomplete weaker signal suppression.

Let the input to the first mixer and limiters be denoted by

$$e_{in}(t) = \cos[\omega_1 t + \theta_s(t)] + \underline{a}\cos[\omega_1 t + \theta_w(t)] \qquad 8.1$$

Let the output of the weaker signal suppressor when excited by this signal be denoted by

$$e_{Lim}(t) = \cos[\omega_1 t + \theta_s(t)] + \underline{a}'\cos[\omega_1 t + \theta_w(t)] \qquad 8.2$$

The presence of the weaker signal at the new interference ratio $\underline{a}'$ ($\underline{a}' < \underline{a}$) indicates the incomplete suppression of the weaker signal by the limiters. For these two input signals given by Eqs. 8.1 and 8.2, the output of the fixed-trap circuit is given by

$$\begin{aligned} e_{OUT}(t) = {} & [\underline{a} + \delta\underline{a}'(1+\underline{a}\underline{a}')]\cos[\omega_1 t + \theta_w(t)] \\ & + [\underline{a}'^2 + \delta(1+\underline{a}\underline{a}')]\cos[\omega_1 t + \theta_s(t)] \\ & + \underline{a}'\cos[\omega_1 t + 2\theta_s(t) - \theta_w(t)] \\ & \underline{a}\,\underline{a}'\cos[\omega_1 t + 2\theta_w(t) - \theta_s(t)] \end{aligned} \qquad 8.3$$

If we assume that $\underline{a} \gg \delta$ where $\delta \approx .01$ and that $\underline{a}' \approx 1/2\ \underline{a}$ then we can express each amplitude coefficient in terms of a and put Eq. 8.3 into simpler form.

$$\begin{aligned} e_{OUT}(t) &= \underline{a}\, Cos[\omega_1 t + \theta_w(t)] \\ &+ \frac{1}{4\underline{a}^2} Cos[\omega_1 t + \theta_s(t)] \\ &+ \frac{1}{2\underline{a}} Cos[\omega_1 t + 2\theta_s(t) - \theta_w(t)] \\ &+ \frac{1}{2\underline{a}^2} Cos[\omega_1 t + 2\theta_w(t) - \theta_s(t)] \end{aligned} \qquad 8.4$$

Appearing in the output, then, we have four co-channel signals as a result of the incomplete suppression and the double mixing operation in the trap system. The largest component is still the desired signal $Cos[\omega_1 t+\theta_w(t)]$; however, there are components very near in magnitude having the modulations $[2\theta_s(t)-\theta_w(t)]$ and $[2\theta_w(t)-\theta_s(t)]$. The originally undesired signal has been significantly suppressed since the ratio of its magnitude to the desired signal magnitude is $\frac{1}{4\underline{a}}$. We note also that in the range of a $\delta \ll \underline{a} < 1$ the parameter, δ, is of little significance since by approximation, it doesn't even appear in any of the signals present in the output.

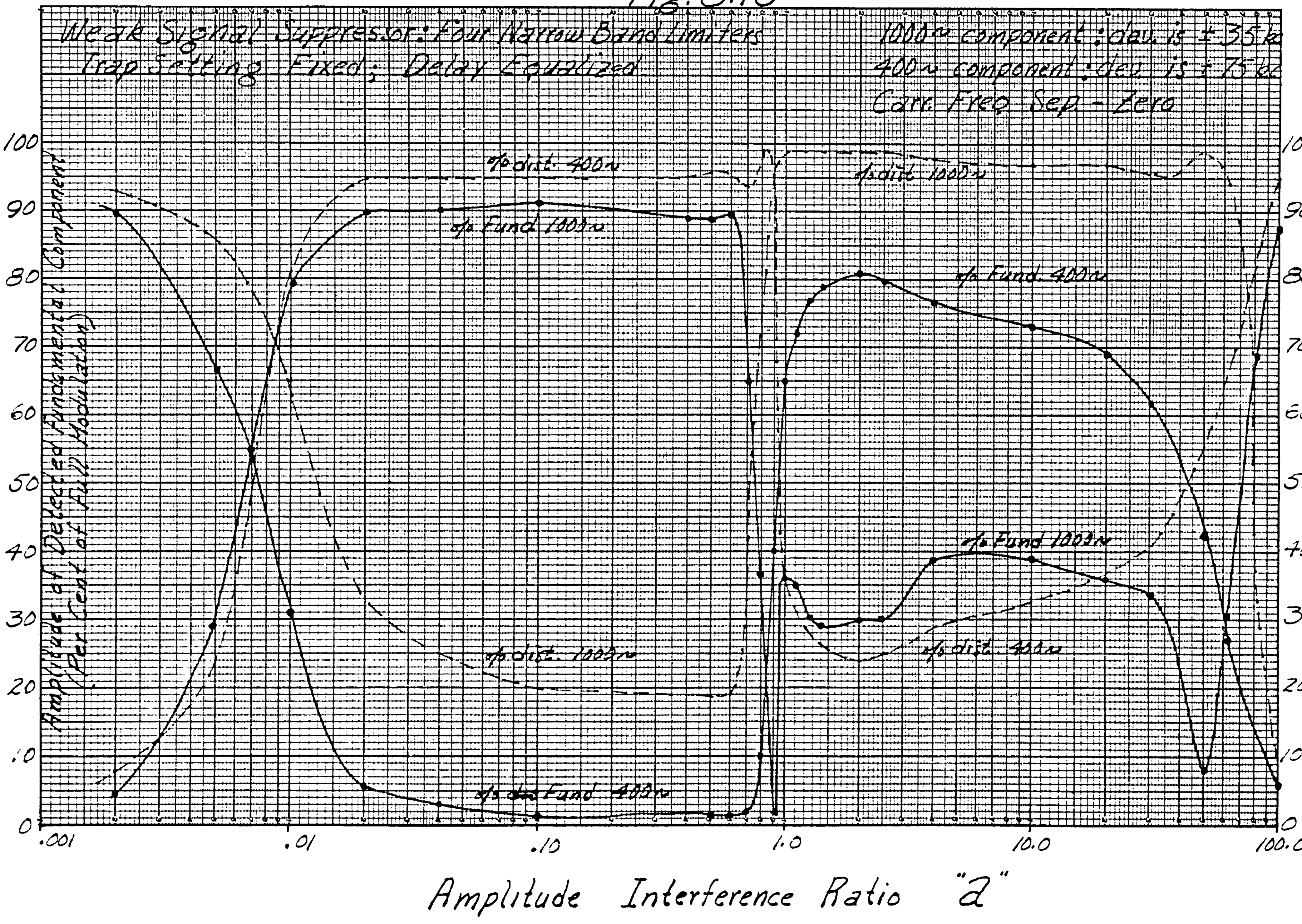
The problem is complicated more if there are any nonlinearities present in either the phase or the amplitude characteristic of the filters directly preceding the demodulator. The presence of these nonlinearities causes harmonic distortion and intermodulation distortion among all the signals present.

Even though the fundamental component of the undesired signal is suppressed, the 45% distortion of the desired signal could probably be due to the harmonics and intermodulation products arising because of the unavoidable phase nonlinearities in the IF amplifier immediately preceding demodulator and because of the inability of the limiters to significantly suppress the weaker signal.

The next two sets of capture characteristics and photographs appear in Fig. 8.13 through 8.16. The modulation conditions are described in the upper right hand corner of each characteristic. Both of these characteristics contain a multitude of anomalies. First to be observed is that the FM signals which were frequency deviated to $\pm$ 35 kc are effectively captured with a relatively low percentage of distortion (12 to 20%). However, the capture of the FM signals which were frequency deviated to $\pm$ 75 kc for either the 400 cycle or the 1000 cycle modulating frequency is very poor. The capture of these signals reach a maximum of 80% with a minimum distortion of 25% in Fig. 8.13, and a maximum of 76% with a minimum distortion of 25% in Fig. 8.15. This poor capture is also accompanied by an incomplete suppression of the interfering signal in these capture regions, although the distortion present in the interfering signal demodulated component is above 95% for both cases.

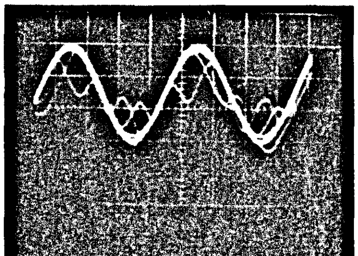
The phenomenon occuring near the capture transition region of each characteristic is one that occured only for these conditions of modulation of interference and desired signal. Furthermore, only the FM signals that were frequency deviated to $\pm$ 35 kc experienced this anomaly. In this capture transition region we expect a smooth sharp rise in the capture of the weaker signal towards

Fig. 8.13

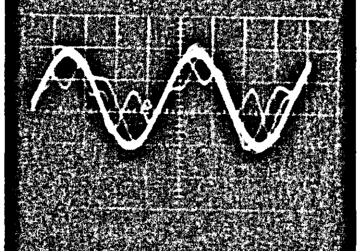


Weak Signal Suppressor: Four Narrow Band Limiters
Trap Setting Fixed; Delay Equalized
1000~ component: dev. is ± 35 kc
400~ component: dev. is ± 75 kc
Carr. Freq. Sep. - Zero
% dist. 400~
% Fund. 1000~
% dist. 1000~
% Fund. 400~
% Fund. 1000~
% dist. 1000~
% dist. 400~
% Fund. 400~
Amplitude of Detected Fundamental Component (Per Cent of Full Modulation)
0
10
20
30
40
50
60
70
80
90
100
.001
.01
.10
1.0
10.0
100.0
Amplitude Interference Ratio "a"

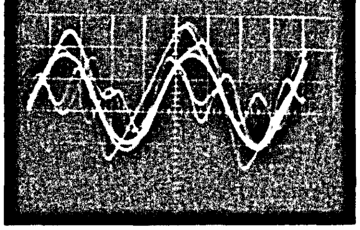
Pic. No. 1



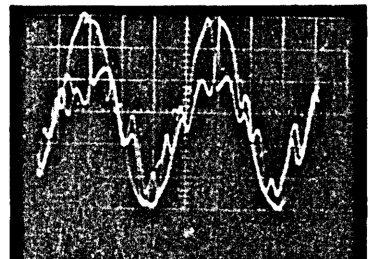
Pic. No. 2



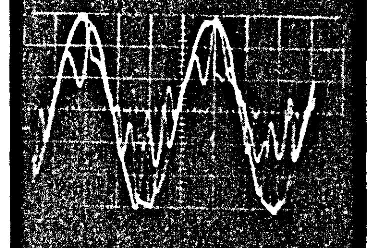
Pic. No. 3



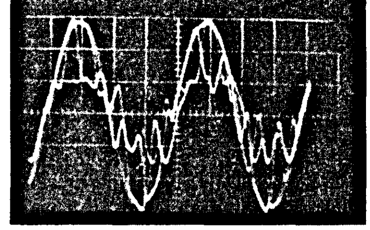
Pic. No. 4



Pic. No. 5



Pic. No. 6



Pic. No. 1 — $\underline{a} = 0.6$
% Fund. = 90%
% Dist. = 19%

Pic. No. 2 — $\underline{a} = 0.1$
% Fund. = 91%
% Dist. = 20%

Pic. No. 3 — $\underline{a} = 0.02$
% Fund. = 90%
% Dist. = 33%

Weaker Signal Modulation
35 kc. deviation
1000 cps. mod. freq.

Stronger Signal Modulation
75 kc. deviation
400 cps. mod. freq.

Pic. No. 4 — $\underline{a} = 0.8$
% Fund. = 17%
% Dist. = 28%

Pic. No. 5 — $\underline{a} = 0.1$
% Fund. = 73%
% Dist. = 33%

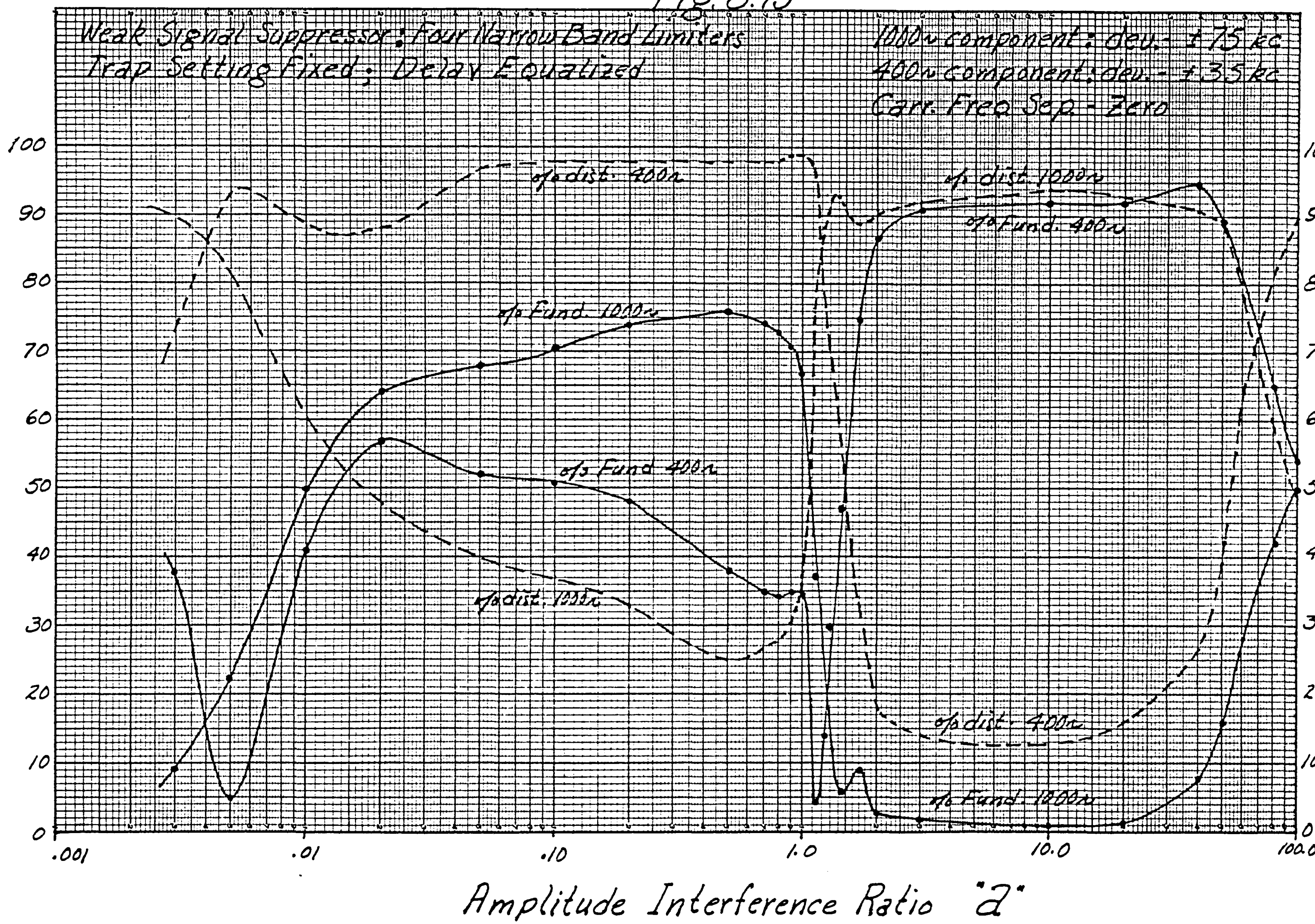
Pic. No. 6 — $\underline{a} = 0.05$
% Fund. = 69%
% Dist. = 37%

Weaker Signal Modulation
75 kc. deviation
400 cps. mod. freq.

Stronger Signal Modulation
35 kc. deviation
1000 cps. mod. freq.

Fig. 8.14 DEMODULATED WEAKER SIGNAL MESSAGE WAVEFORMS

Fig. 8.15

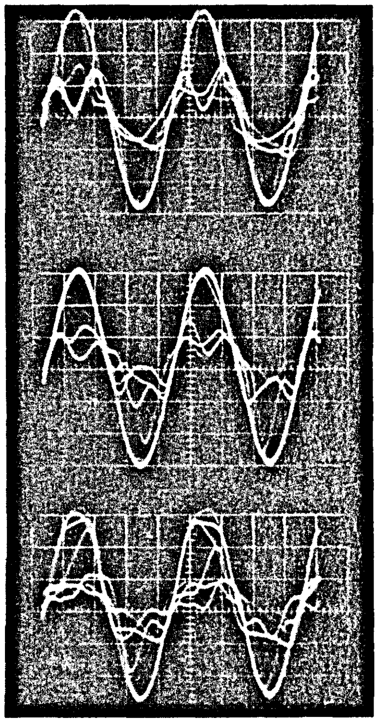


Weak Signal Suppressor: Four Narrow Band Limiters
Trap Setting Fixed; Delay Equalized
1000~ component: dev. ± 75 kc
400~ component: dev. ± 35 kc
Carr. Freq Sep. - Zero
% dist. 400~
% dist. 1000~
% Fund. 400~
% Fund. 1000~
% Fund 400~
% dist. 1000~
% dist. 400~
% Fund. 1000~
100
90
80
70
60
50
40
30
20
10
0
.001
.01
.10
1.0
10.0
100.0
Amplitude Interference Ratio "a"

Pic. No. 1

Pic. No. 2

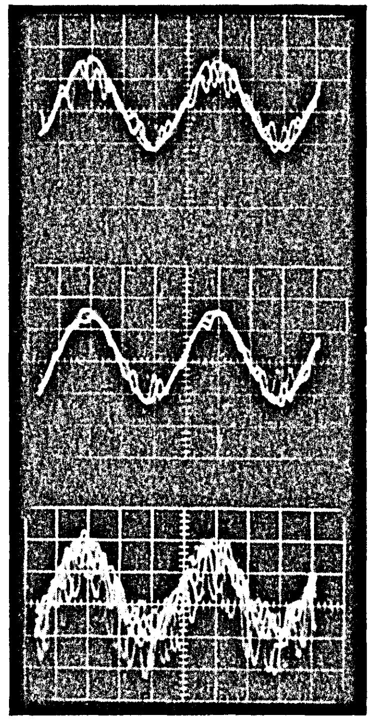
Pic. No. 3



Pic. No. 4

Pic. No. 5

Pic. No. 6



Pic. No. 1 $\underline{a}$ = 0.8
% Fund. = 73%
% Dist. = 28%

Pic. No. 2 $\underline{a}$ = 0.1
% Fund. = 71%
% Dist. = 37%

Pic. No. 3 $\underline{a}$ = 0.02
% Fund. = 64%
% Dist. = 43%

Weaker Signal Modulation
75 kc. deviation
1000 cps. mod. freq.

Stronger Signal Modulation
35 kc. deviation
400 cps. mod. freq.

Pic. No. 4 $\underline{a}$ = 0.5
% Fund. = 87%
% Dist. = 18%

Pic. No. 5 $\underline{a}$ = 0.1
% Fund. = 92%
% Dist. = 13%

Pic. No. 6 $\underline{a}$ = 0.02
% Fund. = 90%
% Dist. = 43%

Weaker Signal Modulation
35 kc. deviation
400 cps. mod. freq.

Stronger Signal Modulation
75 kc. deviation
1000 cps. mod. freq.

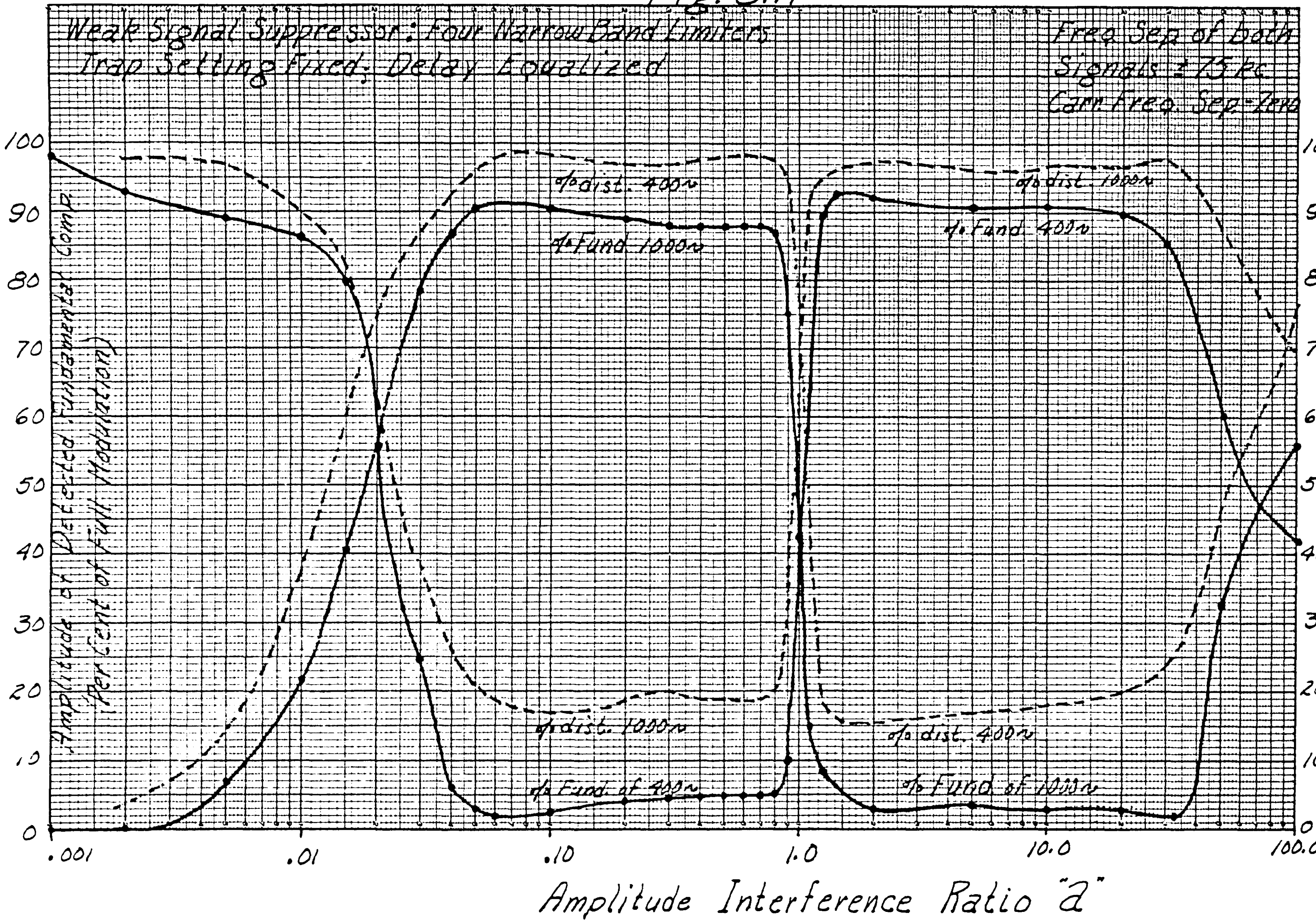
Fig. 8.16 DEMODULATED WEAKER SIGNAL MESSAGE WAVEFORMS

maximum capture; however, the capture curve takes a sharp dip down to below 5% and then rises very abruptly to its maximum when the interference ratio was decreased slightly. In both cases this occured for an $\underline{a} = 0.9$ and $\underline{a} = 1.1$ ($\underline{a} = 0.9$). What probably happened is that because of the incomplete suppression in the limiters, the weaker signal which was deviated to only $\pm$ 35 kc was also thrown into the trap resulting in a loss of capture.

Another anomaly evident, was that for both characteristics, the suppression of the undesired FM signal that was frequency deviated to $\pm$ 35 kc experienced a dip for very small $\underline{a}$ just prior to the capture transition region at the extermities. ($\underline{a} = 0.01$, $\underline{a} = 100$) This happened only when the interfering signal was deviated to $\pm$ 35 kc. This is similar to the peculiar response of the fixed-trap system for simulated conditions when the interfering FM signal was deviated to $\pm$ 35 kc (see Fig. 8.6). However, most of these peculiar responses, including those in the capture transition region centered at $\underline{a} = 1$, occur at values of $\underline{a}$ for which we knew the system would not be reliable.

The last capture characteristic for this series shown in Fig. 8.17 is for the conditions of modulations where both the interference and the desired signals were deviated to $\pm$ 75 kc. Fig. 8.18 shows some photographs of some of the typical waveforms of the demodulated signal. This characteristic represents the best response of the system for all conditions of modulation of both signals. Although the conditions of modulation of the interfering stronger signal are more severe for the trap, they are

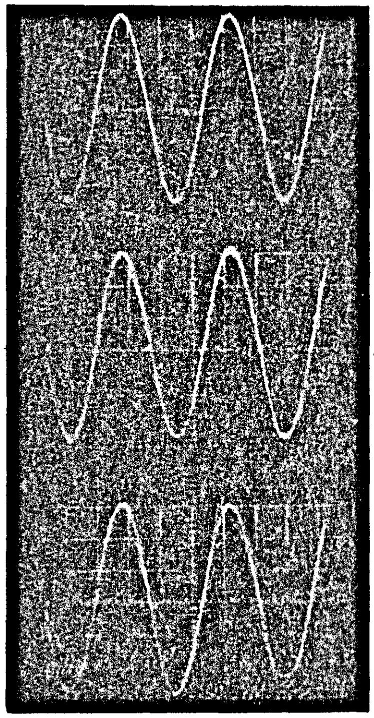
Fig. 8.17


Weak Signal Suppressor: Four Narrow Band Limiters
Trap Setting Fixed; Delay Equalized
Freq. Sep. of Both Signals ±75 kc
Carr. Freq. Sep. = Zero
Amplitude of Detected Fundamental Comp. (Per Cent of Full Modulation)
% dist. 400~
% Fund. 1000~
% dist. 1000~
% Fund. of 400~
% dist. 1000~
% Fund. 400~
% dist. 400~
% Fund. of 1000~
0
10
20
30
40
50
60
70
80
90
100
.001
.01
.10
1.0
10.0
100.0
Amplitude Interference Ratio "a"

Pic. No. 1

Pic. No. 2

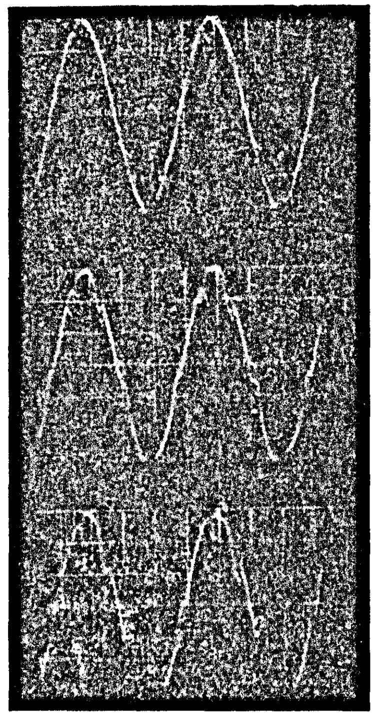
Pic. No. 3



Pic. No. 4

Pic. No. 5

Pic. No. 6



Pic. No. 1   a = 0.8
% Fund. = 87%
% Dist. = 20%

Pic. No. 2   a = 0.1
% Fund. = 90%
% Dist. = 17%

Pic. No. 3   a = 0.05
% Fund. = 91%
% Dist. = 91%

Weaker Signal Modulation
75 kc. deviation
1000 cps. mod. freq.

Stronger Signal Modulation
75 kc. deviation
400 cps. mod. freq.

Pic. No. 4   a = 0.8
% Fund. = 91%
% Dist. = 18%

Pic. No. 5   a = 0.1
% Fund. = 91%
% Dist. = 18%

Pic. No. 6   a = 0.03
% Fund. = 86%
% Dist. = 25%

Weaker Signal Modulation
75 kc. deviation
400 cps. mod. freq.

Stronger Signal Modulation
75 kc. deviation
1000 cps. mod. freq.

Fig. 8.18 DEMODULATED WEAKER SIGNAL MESSAGE WAVEFORMS

more favorable for the limiters, which resulted in improved capture of the desired signal. The capture of the fundamental component was 90% with about 17% total distortion. With both signals deviated to $\pm$ 75 kc, the suppression of the weaker signal in the limiters was more complete thus facilitating suppression of the stronger signal by the fixed-trap circuit resulting in improved weaker signal capture.

We note also that the capture transition region centered about $\underline{a} = 1$ is uniform and quite abrupt. Furthermore, because of the delay equalization both capture regions of the 1000 cycle and the 400 cycle extend over the same range of $\underline{a}$. (This was true of all the capture curves presented so far since in all cases the delay was equalized). The system also does not seem to favor capturing of either component for these conditions of modulation. Both components are captured equally well when they are the weaker signals; and both are suppressed equally well when they are the stronger signals.

Performance of Fixed-Trap System for Unsimulated Conditions

Modulations Fixed, Weaker-Signal Suppression Varied

The next series of capture characteristics are presented in Fig. 8.19 through Fig. 8.33 with photographs of typical output waveforms accompanying each capture characteristic. What we intend to show from these curves is the following:

1 - The degree of weaker signal suppression required for satisfactory capture of the desired weaker signal.

2 - The effect upon the response of an unbalance in the delay

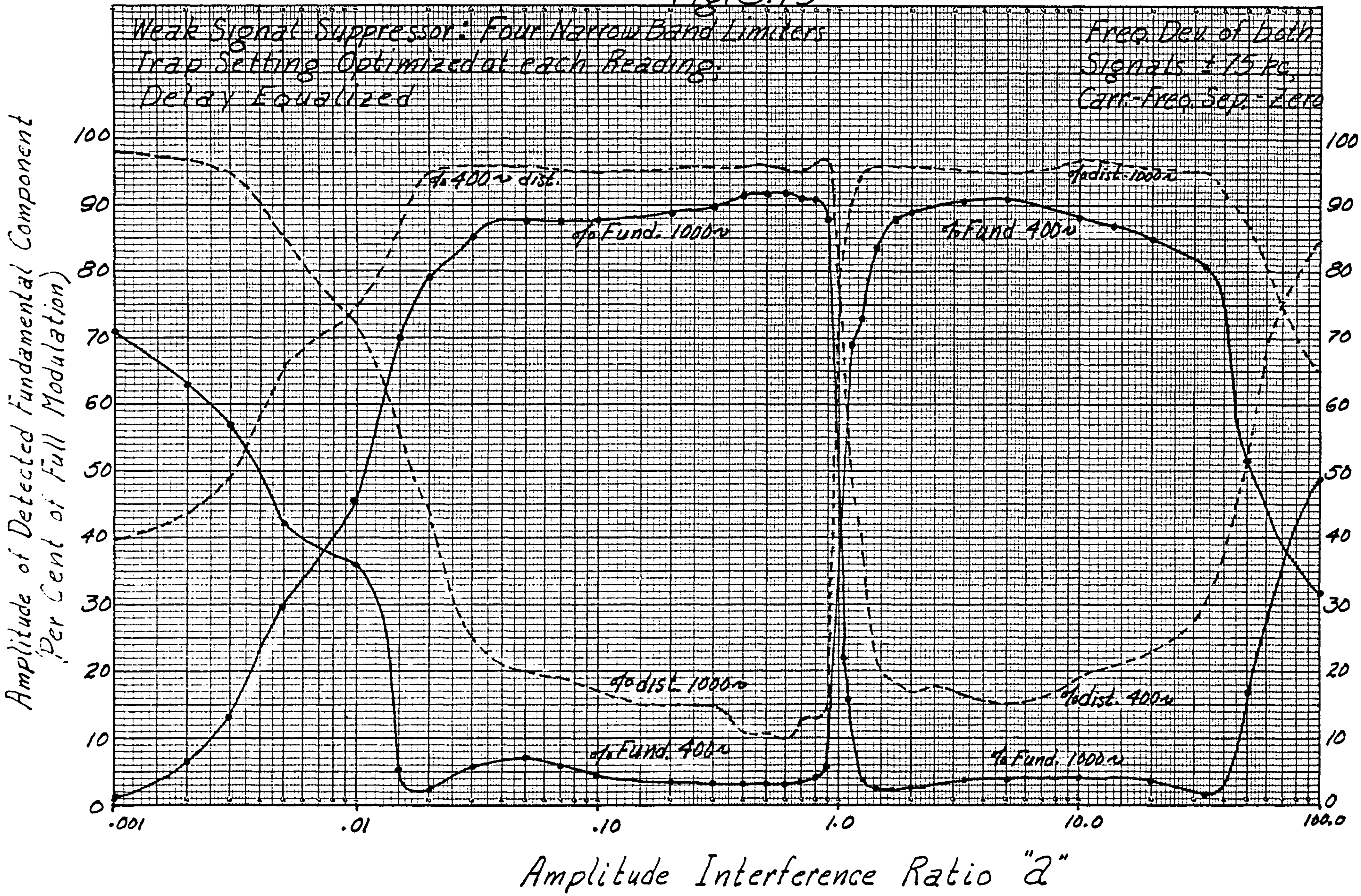
between the two channels feeding the first mixer.

For the remainder of the capture characteristics that will be presented, the frequency deviation of both signals is fixed at $\pm$ 75 kc. The modulating frequencies are still 1000 cycles and 400 cycles respectively.

The first capture characteristic of the series shown in Fig. 8.19 and the accompanying photographs shown in Fig. 8.20 was obtained for the equalized delay case using four narrow band limiters as the weak signal suppressor. The only difference between this capture characteristic and that discussed previously (Fig. 8.17) is that the trap setting was optimized at each setting of a for this test. This accounts for the peculiar dips in all the curves at various points. Apparently, all that was accomplished by optimizing the trap for each setting of a was an extension of the lower and upper capture transition regions to lower and higher values of a, respectively. As a result of optimizing the trap response (position and attenuation of the trap null was adjusted by the fine tuning controls provided) the lower capture transition region was shifted from an $a = 0.02$ for the previous characteristic to an $a = 0.007$ for the present characteristic. This represents some improvement; however, in lieu of the difficulty of having to optimize the trap at each reading (which was very sensitive for low values of a) the improvement attainable was not worth all the trouble. Significant results were still attainable for one trap setting at an $a \approx 0.05$.

In performing this test, it was learned that the trap setting

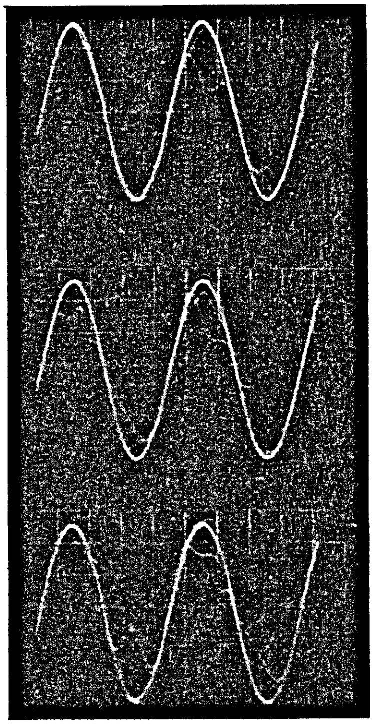
Fig. 8.19


Weak Signal Suppressor: Four Narrow Band Limiters
Trap Setting Optimized at each Reading;
Delay Equalized
Freq. Dev. of both
Signals ±75 kc,
Carr-Freq. Sep.-Zero
Amplitude of Detected Fundamental Component
(Per Cent of Full Modulation)
% 400~ dist.
% Fund. 1000~
% dist. 1000~
% Fund. 400~
% dist. 1000~
% Fund. 400~
% dist. 400~
% Fund. 1000~
100
90
80
70
60
50
40
30
20
10
0
.001
.01
.10
1.0
10.0
100.0
Amplitude Interference Ratio "a"

Pic. No. 1

Pic. No. 2

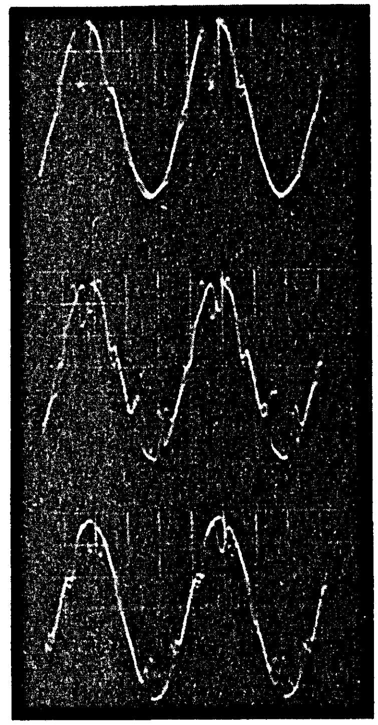
Pic. No. 3



Pic. No. 4

Pic. No. 5

Pic. No. 6



Pic. No. 1 — a = 0.9
% Fund. = 88%
% Dist. = 15%

Pic. No. 2 — a = 0.1
% Fund. = 88%
% Dist. = 17%

Pic. No. 3 — a = 0.05
% Fund. = 88%
% Dist. = 20%

Weaker Signal Modulation
75 kc. deviation
1000 cps. mod. freq.

Stronger Signal Modulation
75 kc. deviation
400 cps. mod. freq.

Pic. No. 4 — a = 0.5
% Fund. = 2.7%
% Dist. = 96%

Pic. No. 5 — a = 0.1
% Fund. = 4.2%
% Dist. = 97%

Pic. No. 6 — a = 0.05
% Fund. = 4.0%
% Dist. = 96%

Weaker Signal Modulation
75 kc. deviation
400 cps. mod. freq.

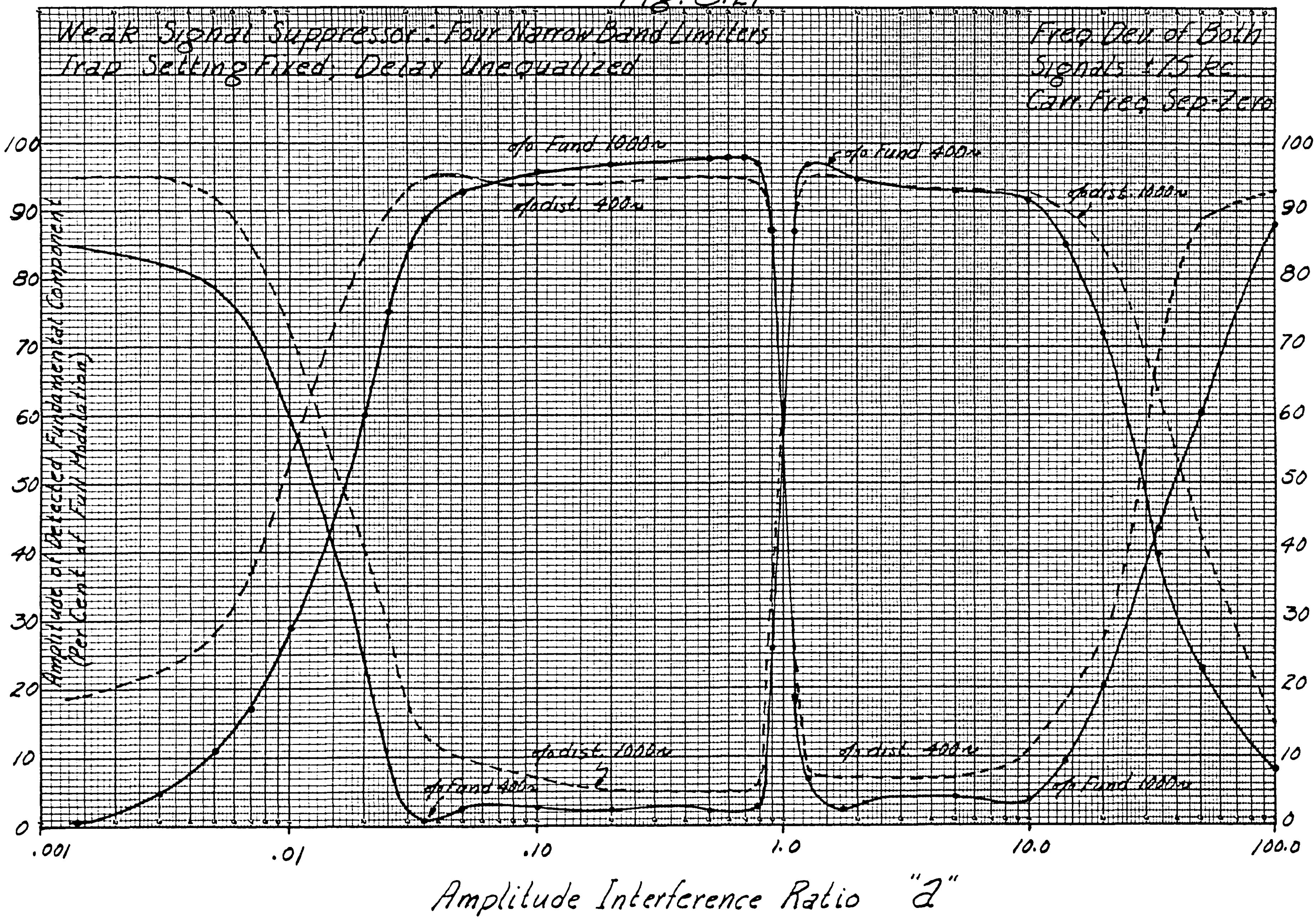
Stronger Signal Modulation
75 kc. deviation
1000 cps. mod. freq.

Fig. 8.20 DEMODULATED WEAKER SIGNAL MESSAGE WAVEFORMS

was not critical for high values of a. This confirms the discussion pertinent to Eq. 8.4 where we deduced that for $a \gg \delta$, the trap attenuation factor δ was of little significance to the capture of the weaker signal. The tuning of the trap for values of $a > 0.5$ was rather arbitrary since capture was achieved over a broad range of trap attenuations. The maximum value for the trap attenuation that still resulted in capture of the weaker signal for values of $a > 0.5$ was approximately 6 db. For values of attenuation less than this, capture of the stronger signal resulted. However, although the trap attenuation setting was somewhat arbitrary for values of $a > 0.1$, it was very critical for values of $a < 0.1$. Hence in all of the tests that follow, the trap setting was optimized for an $a = 0.05$. This was quite simple and quick to do and the trap setting although critical, did not require further adjustment for the duration of each run.

The second characteristic for this series shown in Fig. 8.21 with photographs in Fig. 8.22 was also obtained for a weaker-signal suppressor consisting of four narrow-band limiters. However, for this test the delay between the converter channel and the equalizer channel was unbalanced by about 5 μsec. The unequalization in delay resulted in two significant changes in the capture characteristics for these test conditions. The first change was the familiar broadening and shifting of the upper capture transition region. Thus capture of the desired weaker signal 400 cycle component was restricted to a smaller range of a than the range of a for which the 1000 cycle component was captured. The reason for

Fig. 8.21



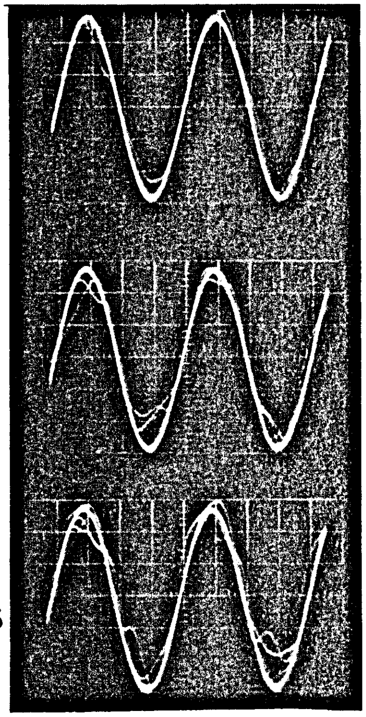
Weak Signal Suppressor: Four Narrow Band Limiters
Trap Setting Fixed, Delay Unequalized
Freq Dev of Both Signals ±7.5 kc
Carr. Freq. Sep-Zero
% Fund 1000~
% Fund 400~
% dist. 400~
% dist. 1000~
% dist. 1000~
% Fund 400~
% dist. 400~
% Fund 1000~
Amplitude of Detected Fundamental Component (Per Cent of Full Modulation)
100
90
80
70
60
50
40
30
20
10
0
.001
.01
.10
1.0
10.0
100.0
Amplitude Interference Ratio "a"

173

Pic. No. 1

Pic. No. 2

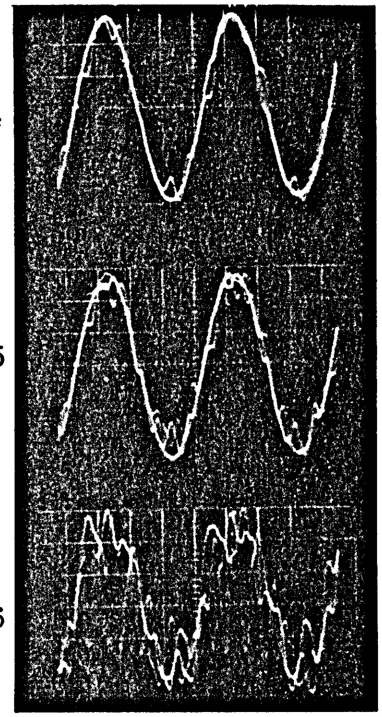
Pic. No. 3



Pic. No. 4

Pic. No. 5

Pic. No. 6



Pic. No. 1 $\underline{a} = 0.8$
% Fund. = 97%
% Dist. = 6%

Pic. No. 2 $\underline{a} = 0.1$
% Fund. = 96%
% Dist. = 7%

Pic. No. 3 $\underline{a} = 0.05$
% Fund. = 94%
% Dist. = 10%

Weaker Signal Modulation
75 kc. deviation
1000 cps. mod. freq.

Stronger Signal Modulation
75 kc. deviation
400 cps. mod. freq.

Pic. No. 4 $\underline{a} = 0.8$
% Fund. = 97%
% Dist. = 8%

Pic. No. 5 $\underline{a} = 0.1$
% Fund. = 92%
% Dist. = 11%

Pic. No. 6 $\underline{a} = 0.05$
% Fund. = 72%
% Dist. = 37%

Weaker Signal Modulation
75 kc. deviation
400 cps. mod. freq.

Stronger Signal Modulation
75 kc. deviation
1000 cps. mod. freq.

Fig. 8.22 DEMODULATED WEAKER SIGNAL MESSAGE WAVEFORMS

this, as explained previously, is that because of the delay differential of 5μsec between the two channels, a larger phase differential results for the 1000 cycle message when this is the modulating frequency of the stronger signal than for the situation where the modulating frequency of the stronger signal is only 400 cycles. The correlation between the modulation of the stronger signal appearing at both inputs to the first mixer diminishes more when the modulating frequency is 1000 cycles, resulting in an FM signal appearing in the output of the mixer in place of the required constant difference frequency signal.

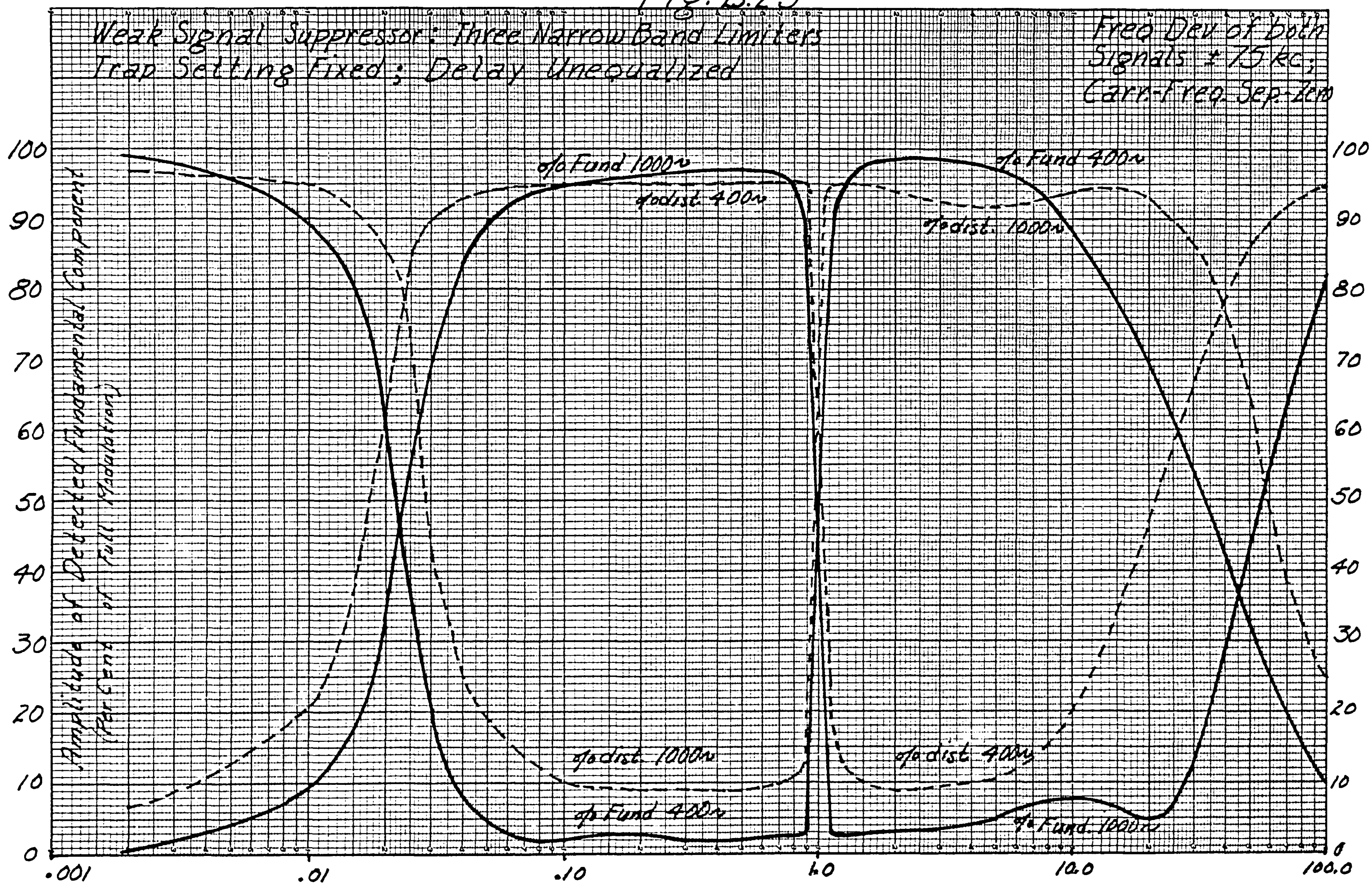
The second effect of unbalancing the delay resulted in a significant improvement of the capture of the desired signal accompanied by a noticeable decrease in the percentage of total attendant distortion. The percentage capture of the fundamental rose about 95% accompanied by a total distortion below 10%. The reason for this improvement is again contingent upon the suppression properties of the limiters in conjunction with the effectiveness of the fixed-trap. Since we always have some residual weaker signal present in the output of the limiters, under conditions where the delay is equalized, this residual weaker signal when mixed with the weaker signal in the delay channel tends to compress the weaker signal modulation to the confines of the trap. However, when there is a slight delay differential between the two channels the cross correlation between the two weaker signals appearing at both inputs of the first mixer is diminished considerable. As a result, the weaker signal modulation is not compressed by the mixing operation thus preventing the weaker signal from being trapped.

The same argument could also be applied to the stronger interfering signal thus incapacitating the system. However, from the results obtained for this system for the simulated conditions we see that the system is still able to capture the weaker of the two signals even though the delay differential exists between the two stronger signals appearing at the input to the first mixer. The only detrimental effect of this delay differential is to reduce the range over which capture of the weaker signal is achieved when the modulating frequency of the stronger interfering signal is 1000 cycles or greater.

The next capture characteristic shown in Fig. 8.23 with photographs in Fig. 8.24 was obtained for a reduced degree of weaker-signal suppression. Only three narrow-band limiters were cascaded in the suppressor unit. The delay for this test was unequalized and the trap setting was optimized for an $\underline{a} = 0.05$. The delay differential between the two channels is now only 2.5 μsecs.

From this characteristic we note that in the capture regions, the capture of the desired signal is as effective as for the test when four limiters were used to suppress the weaker signal (compare to Fig. 8.21). However, the range of $\underline{a}$ over which this capture is achieved has decreased slightly accompanied by a slight shift of the center of the capture transition region to larger $\underline{a}$. The shift of the center was from $\underline{a} = 0.015$ for the four limiter test to an

Fig. 8.23

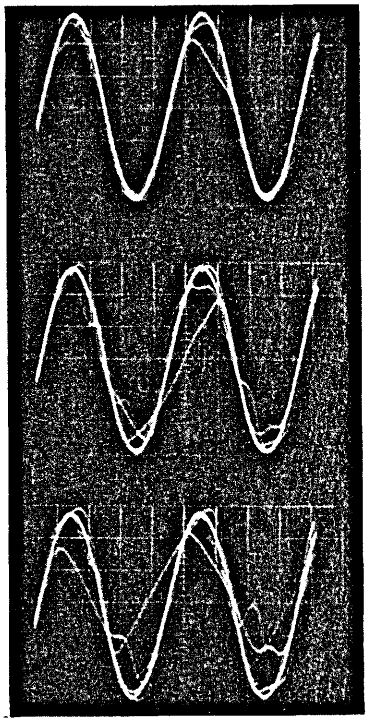


Amplitude Interference Ratio "a"

Pic. No. 1

Pic. No. 2

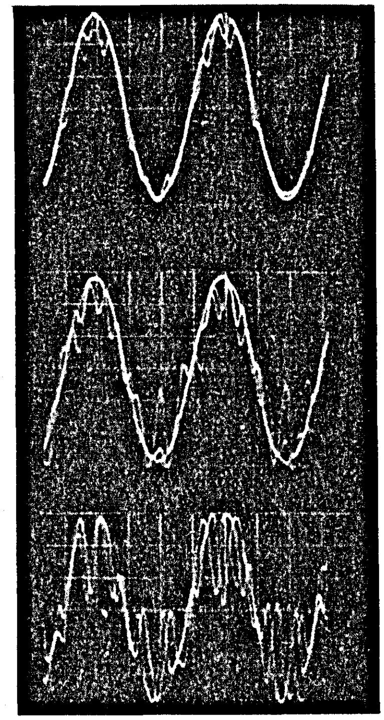
Pic. No. 3



Pic. No. 4

Pic. No. 5

Pic. No. 6



Pic. No. 1 $\underline{a} = 0.8$
% Fund. = 95%
% Dist. = 11%

Pic. No. 2 $\underline{a} = 0.1$
% Fund. = 95%
% Dist. = 10%

Pic. No. 3 $\underline{a} = 0.05$
% Fund. = 90%
% Dist. = 19%

Weaker Signal Modulation
75 kc. deviation
1000 cps. mod. freq.

Stronger Signal Modulation
75 kc. deviation
400 cps. mod. freq.

Pic. No. 4 $\underline{a} = 0.8$
% Fund. = 95%
% Dist. = 13%

Pic. No. 5 $\underline{a} = 0.1$
% Fund. = 89%
% Dist. = 21%

Pic. No. 6 $\underline{a} = 0.05$
% Fund. = 70%
% Dist. = 47%

Weaker Signal Modulation
75 kc. deviation
400 cps. mod. freq.

Stronger Signal Modulation
75 kc. deviation
1000 cps. mod. freq.

Fig. 8.24 DEMODULATED WEAKER SIGNAL MESSAGE WAVEFORMS

$\underline{a} = 0.023$ for the three limiter test. This shift in the center of the lower and upper capture transition to regions could also be due to a slight difference in the setting of the trap attenuation factor δ, prior to making the test run for each case.

Because of the slight delay differential between the two channels, the capture region where the 400 cycle component is the desired signal and the 1000 cycle component is the interfering signal does not extend over as large a range of $\underline{a}$ as for the opposite side where the role of desired and interfering signal modulations are interchanged. As we shall see in the remainder of the capture characteristics, the only detrimental effect of this delay differential was to reduce the range of $\underline{a}$ for which capture was achieved when the modulating frequency of the interfering signal was equal to or exceeded 1000 cycles.

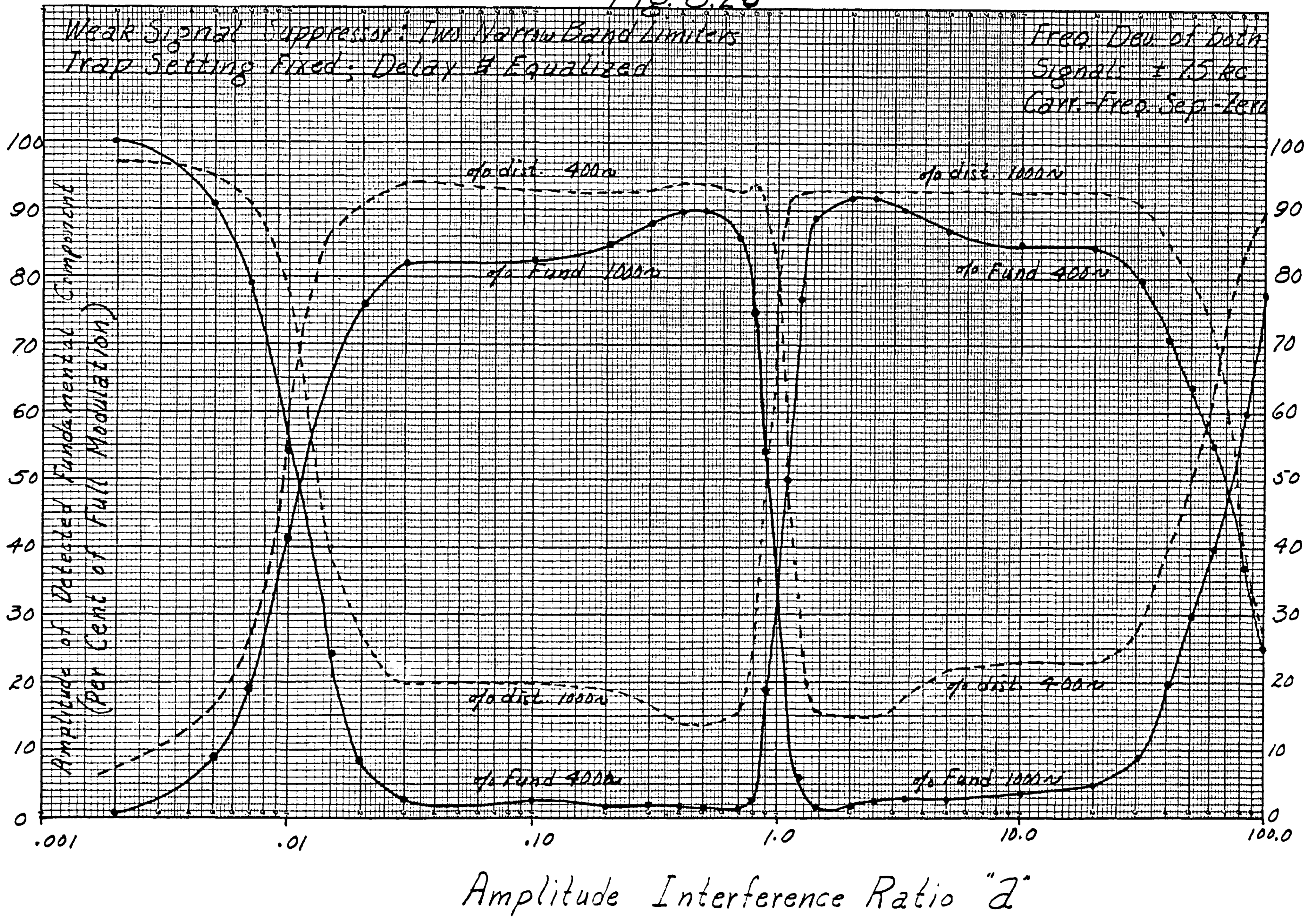
We note from this characteristic that although capture of the desired 400 cycle component in the upper capture region is restricted to a smaller range of $\underline{a}$, the suppression of the 1000 cycle component extends over a broader range of $\underline{a}$. This seems to indicate again, the predominance of harmonics and intermod products in this upper capture region. These distortion products are generated because of the failure of the limiters to suppress significantly the weaker signal. This incomplete weaker suppression of the limiters results in the appearance of the four signals described by Eq. 8.4. The slight nonlinearities of phase and amplitude characteristics of the filters that immediately precede the demodulator manifests all these spurious components in the form of harmonic and intermodulation distortion.

The next capture characteristic is shown in Fig. 8.25 with photographs of typical waveforms in Fig. 8.26. This characteristic was obtained using only two narrow band limiters with the delay equalized. In comparison with the characteristic obtained for the balanced delay, four narrow band-limiter test, Fig. 8.21, capture of the desired signal in the capture regions is not as effective. A peak of 90% in the capture of the desired signal was obtained for an $\underline{a} = 0.5$, after which, the capture of the desired signal receded to a plateau of 82%. For the four narrow-band limiter test, the capture of the desired signal in the capture region did not vary significantly with $\underline{a}$. The distortion for the two narrow-band limiter test was about 20% in both capture regions. Because of the delay equalization, both capture regions extend over approximately the same range of $\underline{a}$.

We note also in comparing the two limiter characteristic Fig. 8.25 with the four limiter characteristic Fig. 8.21 (delay for both equalized) that there are no significant differences in the lower and upper capture transition regions between each characteristic. Apparently then, the trap attenuation characteristic and the demodulator stronger signal capture characteristics are the controlling factors for these lower and upper capture transition regions.

In comparing these same characteristics, we note a broadening of the capture transition region centered about $\underline{a} = 1$. The reason for this is that the higher degree of weaker signal suppression

Fig. 8.25

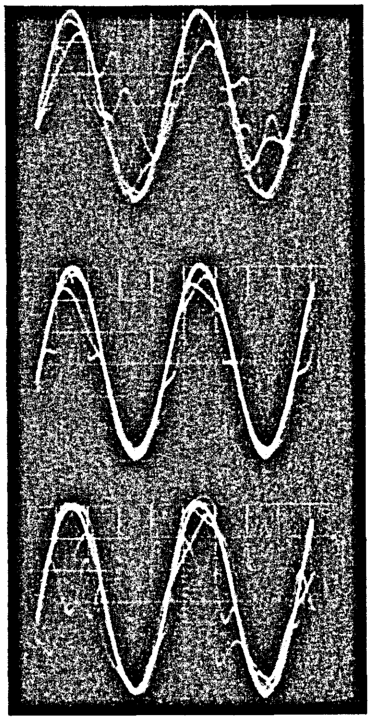


Weak Signal Suppressor: Two Narrow Band Limiters
Trap Setting Fixed; Delay & Equalized
Freq. Dev. of both Signals ± 7.5 kc
Carr.-Freq. Sep.-Zero
% dist. 400 ~
% dist. 1000 ~
% Fund 1000 ~
% Fund 400 ~
% dist. 1000 ~
% dist. 400 ~
% Fund 400 ~
% Fund 1000 ~
Amplitude of Detected Fundamental Component (Per Cent of Full Modulation)
0
10
20
30
40
50
60
70
80
90
100
.001
.01
.10
1.0
10.0
100.0
Amplitude Interference Ratio "a"

Pic. No. 1

Pic. No. 2

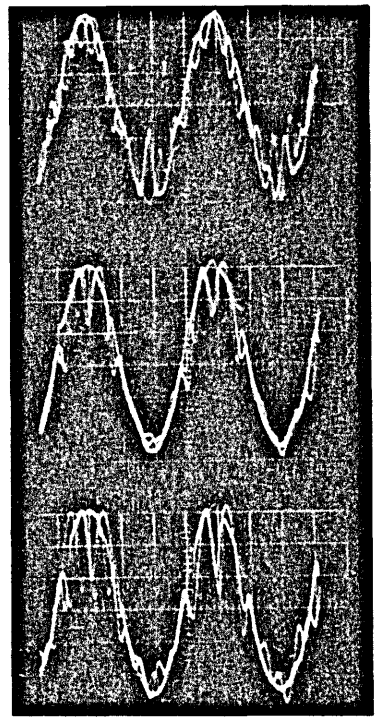
Pic. No. 3



Pic. No. 4

Pic. No. 5

Pic. No. 6

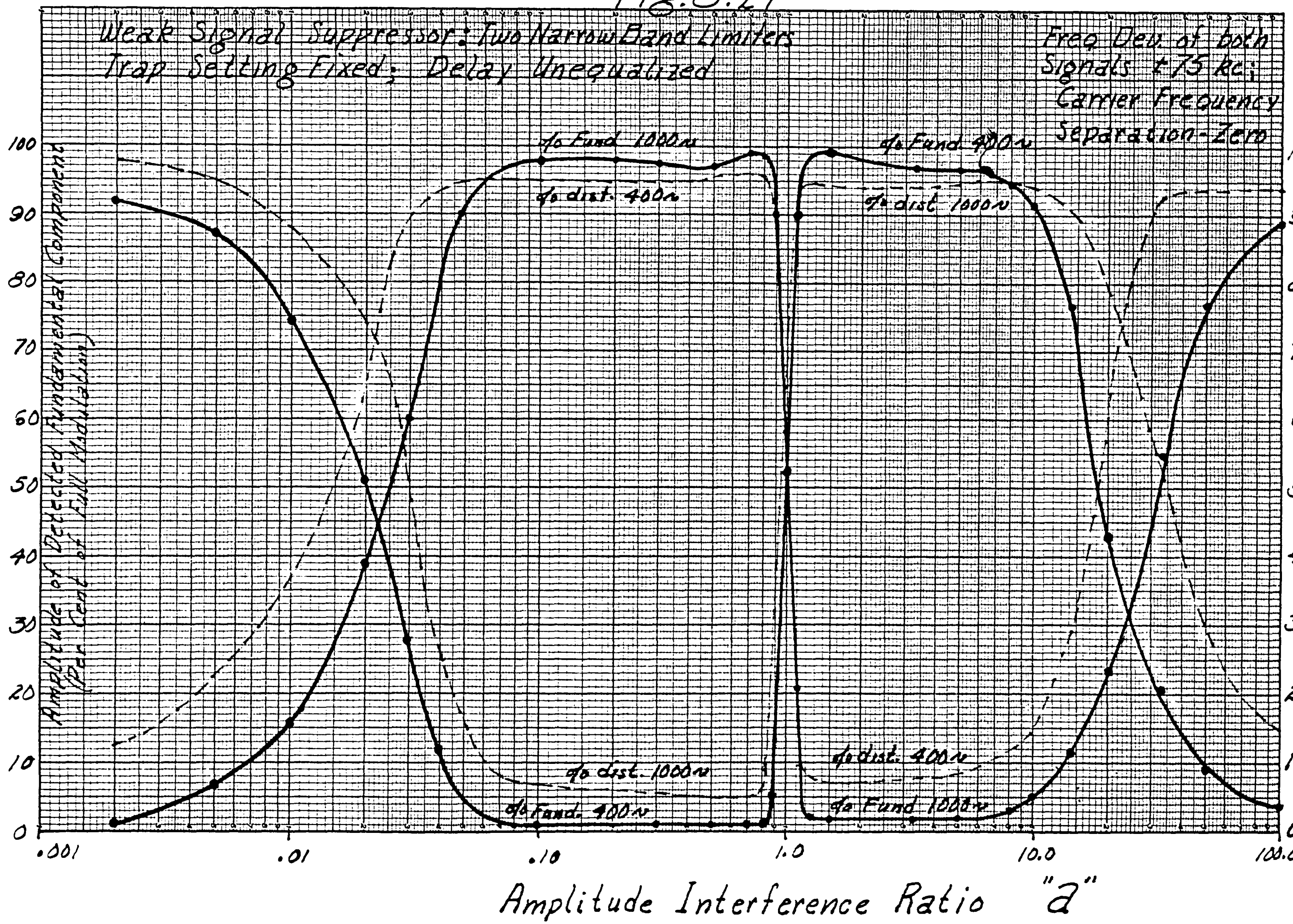


Pic. No. 1	<u>a</u> = 0.8 % Fund. = 78% % Dist. = 26%	Pic. No. 4	<u>a</u> = 0.8 % Fund. = 77% % Dist. = 29%
Pic. No. 2	<u>a</u> = 0.4 % Fund. = 90% % Dist. = 14%	Pic. No. 5	<u>a</u> = 0.4 % Fund. = 92% % Dist. = 15%
Pic. No. 3	<u>a</u> = 0.1 % Fund. = 83% % Dist. = 20%	Pic. No. 6	<u>a</u> = 0.1 % Fund. = 85% % Dist. = 23%

Weaker Signal Modulation 75 kc. deviation 1000 cps. mod. freq.	Weaker Signal Modulation 75 kc. deviation 400 cps. mod. freq.
Stronger Signal Modulation 75 kc. deviation 400 cps. mod. freq.	Stronger Signal Modulation 75 kc. deviation 1000 cps. mod. freq.

Fig. 8.26 DEMODULATED WEAKER SIGNAL MESSAGE WAVEFORMS

Fig. 8.27

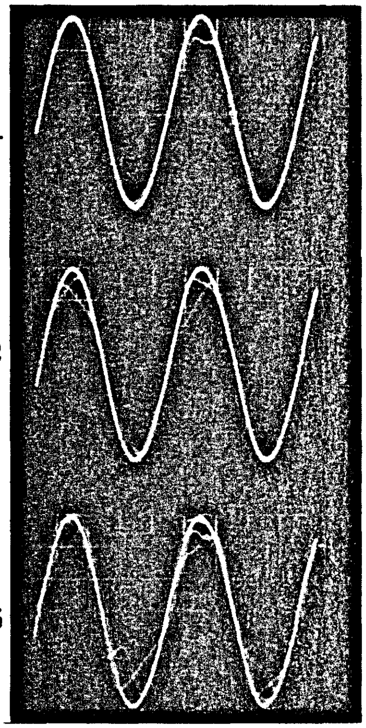


Weak Signal Suppressor: Two Narrow Band Limiters
Trap Setting Fixed; Delay Unequalized
Freq. Dev. of both Signals ±75 kc; Carrier Frequency Separation - Zero
% Fund. 1000~
% dist. 400~
% Fund. 400~
% dist. 1000~
% dist. 1000~
% Fund. 400~
% dist. 400~
% Fund 1000~
Amplitude of Detected Fundamental Component (Per Cent of Full Modulation)
0
10
20
30
40
50
60
70
80
90
100
.001
.01
.10
1.0
10.0
100.0
Amplitude Interference Ratio "a"

Pic. No. 1

Pic. No. 2

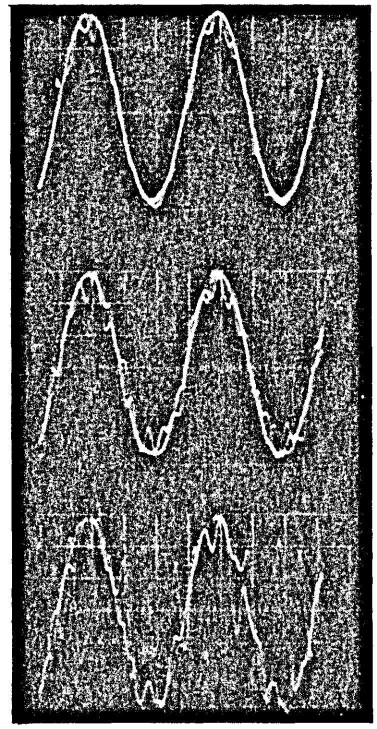
Pic. No. 3



Pic. No. 4

Pic. No. 5

Pic. No. 6



Pic. No. 1 — a = 0.8
% Fund. = 99%
% Dist. = 6%

Pic. No. 2 — a = 0.1
% Fund. = 98%
% Dist. = 7%

Pic. No. 3 — a = 0.05
% Fund. = 91%
% Dist. = 17%

Weaker Signal Modulation
75 kc. deviation
1000 cps. mod. freq.

Stronger Signal Modulation
75 kc. deviation
400 cps. mod. freq.

Pic. No. 4 — a = 0.8
% Fund. = 99%
% Dist. = 8%

Pic. No. 5 — a = 0.1
% Fund. = 92%
% Dist. = 15%

Pic. No. 6 — a = 0.05
% Fund. = 59%
% Dist. = 57%

Weaker Signal Modulation
75 kc. deviation
400 cps. mod. freq.

Stronger Signal Modulation
75 kc. deviation
1000 cps. mod. freq.

Fig. 8.28 DEMODULATED WEAKER SIGNAL MESSAGE WAVEFORMS

achieved by the four narrow-band limiters for values of a that approach unity facilitates the suppression of the stronger signal by the fixed trap circuit thus improving the capture of the weaker signal.

A second set of characteristics were also obtained utilizing two narrow band limiter; however in this test the delay was unbalanced by about 5μsecs. These characteristics and accompanying photographs are shown in Fig. 8.27 and Fig. 8.28. We note again the improved capture of the desired signal in both capture regions. Also evident is the reduced range of a for which capture of the desired signal is achieved in the upper capture region because of the unbalance in the delay. In addition, we also note an improvement in the capture transition region centered about a = 1 when the delay is unbalanced.

The next capture characteristic is shown in Fig. 8.29 with accompanying photographs in Fig. 8.30. This characteristic was obtained for a weaker signal suppressor consisting of two narrow band limiters (BW$>$300 kc) plus a feedforward circuit. One of the limiters was incorporated directly into the design of the feedforward circuit shown in block diagram form in Fig. 8.31

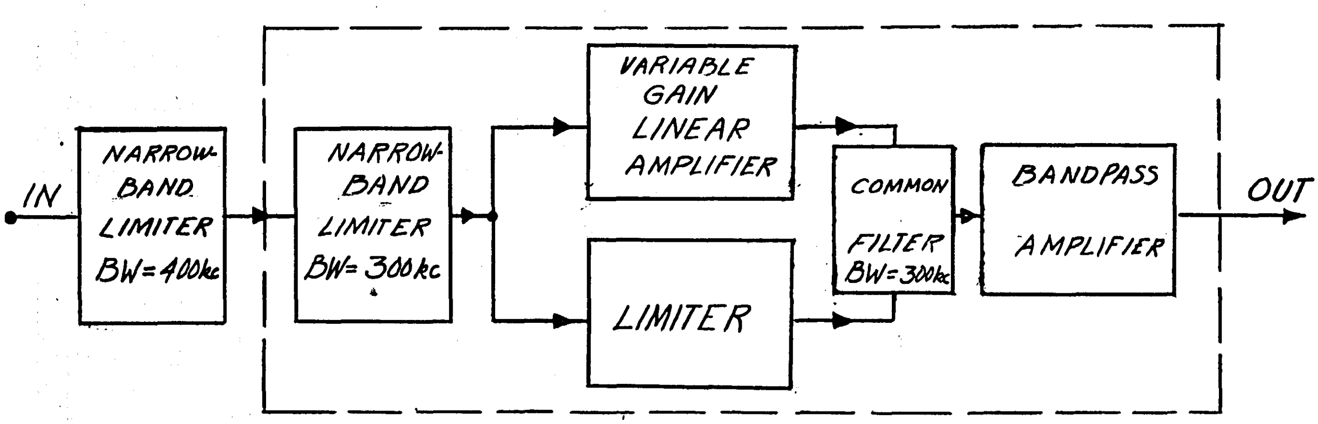
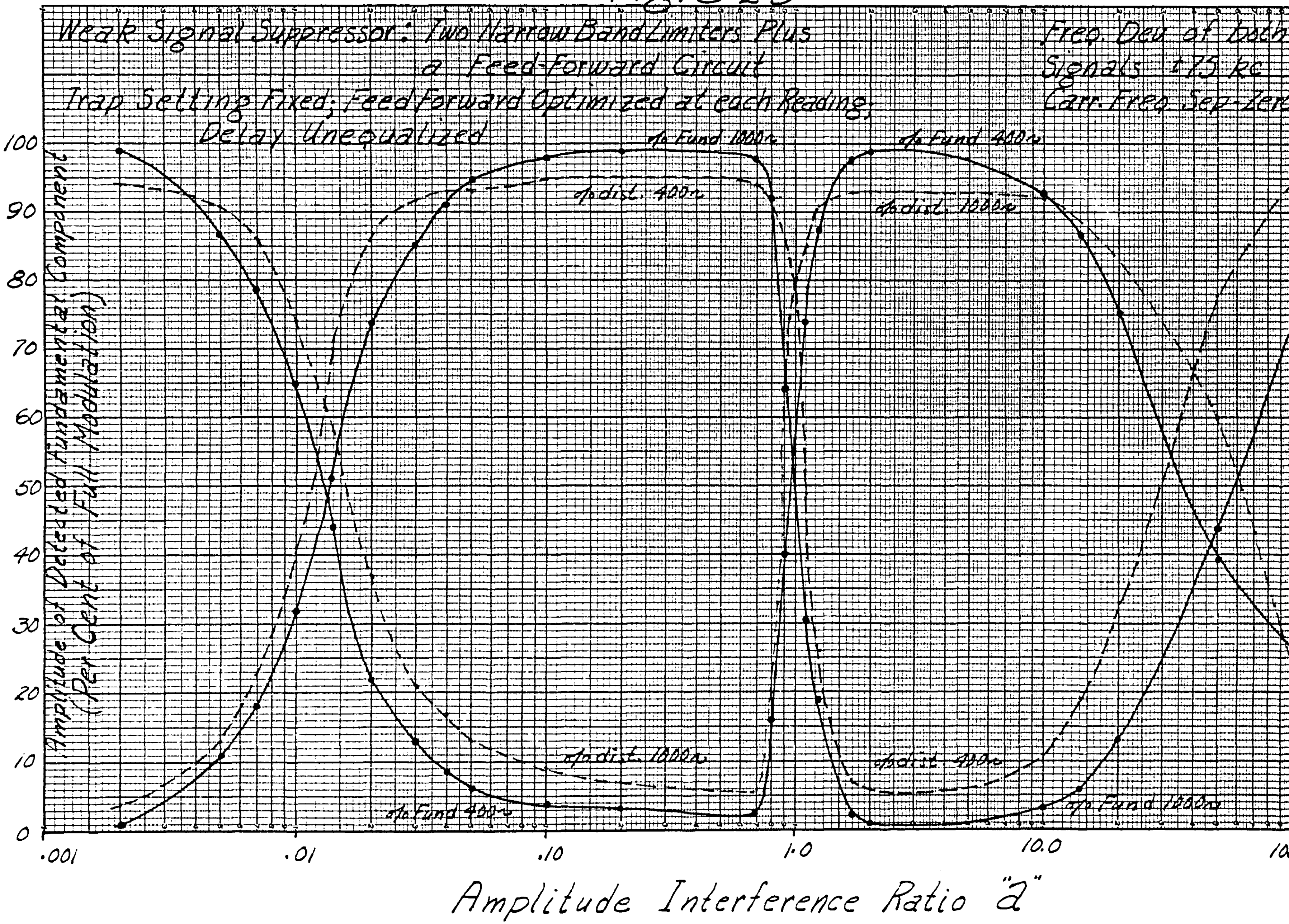


Fig. 8.31 Block Diagram of Weak Signal Suppressor Using Feedforward.

Fig. 8.29

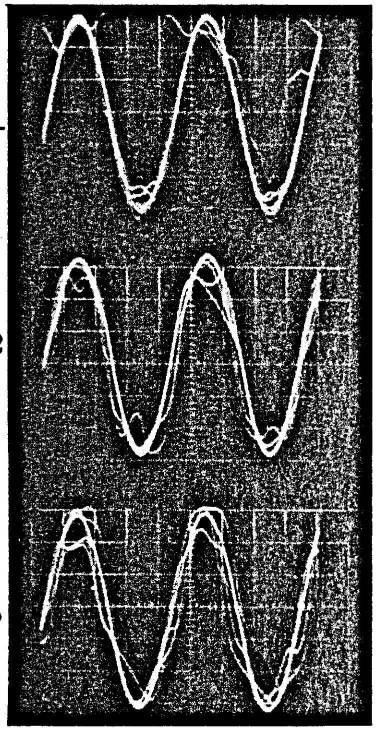


Weak Signal Suppressor: Two Narrow Band Limiters Plus a Feed-Forward Circuit
Trap Setting Fixed; Feed Forward Optimized at each Reading; Delay Unequalized
Freq. Dev. of both Signals ±75 kc
Carr. Freq. Sep. Zero
% Fund 1000~
% Fund 400~
% dist. 400~
% dist. 1000~
% Fund 400~
% dist. 1000~
% dist. 400~
% Fund 1000~
Amplitude of Detected Fundamental Component (Per Cent of Full Modulation)
100
90
80
70
60
50
40
30
20
10
0
.001
.01
.10
1.0
10.0
100.0
Amplitude Interference Ratio "a"

Pic. No. 1

Pic. No. 2

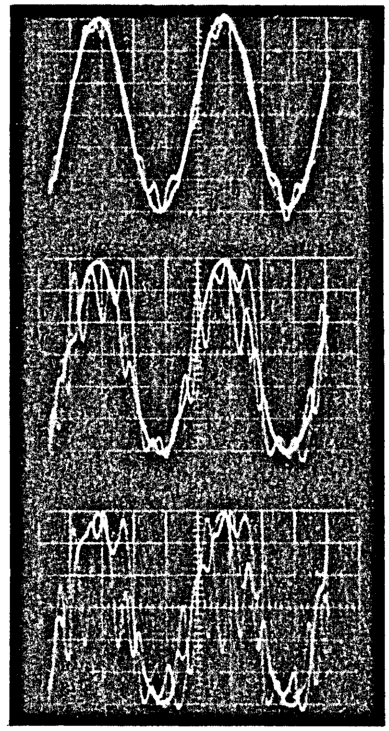
Pic. No. 3



Pic. No. 4

Pic. No. 5

Pic. No. 6



Pic. No. 1 $\underline{a} = 0.8$
% Fund. = 92%
% Dist. = 20%

Pic. No. 2 $\underline{a} = 0.1$
% Fund. = 98%
% Dist. = 9%

Pic. No. 3 $\underline{a} = 0.05$
% Fund. = 95%
% Dist. = 13%

Weaker Signal Modulation
75 kc. deviation
1000 cps. mod. freq.

Stronger Signal Modulation
75 kc. deviation
400 cps. mod. freq.

Pic. No. 4 $\underline{a} = 0.8$
% Fund. = 88%
% Dist. = 26%

Pic. No. 5 $\underline{a} = 0.1$
% Fund. = 93%
% Dist. = 11%

Pic. No. 6 $\underline{a} = 0.05$
% Fund. = 75%
% Dist. = 33%

Weaker Signal Modulation
75 kc. deviation
400 cps. mod. freq.

Stronger Signal Modulation
75 kc. deviation
1000 cps. mod. freq.

Fig. 8.30 DEMODULATED WEAKER SIGNAL MESSAGE WAVEFORMS

When the circuits were rearranged to accomodate the feed-forward circuit, the first limiter of the original cascaded section of four limiters was left in place since this limiter was always the limiter driven by the IF amplifier. This was done in order not to upset the tuning of the IF amplifier. However in an effort to manifest the weaker signal suppression effected by the feedforward only, the bandwidth of the first limiter was widened to 400 kc, or double the IF bandwidth. Although this is still classified as a narrow band limiter, its bandwidth is large enough so that its weaker signal suppression properties are minimized.

In addition, the narrow band limiter incorporated in the feedforward circuit had a bandwidth in excess of 300 kc, hence its suppression properties are also minimized. In essence then, the weak signal suppression is accomplished mainly by the feedforward circuit; however, the sole effect of the feedforward circuit in its enhancement of the weak signal response is to some extent masked by the weaker signal suppression afforded by the two broadly tuned limiters.

The limiters employed in this feedforward circuit were of the grid-charging, self-biasing pentode type which were designed to handle an $\underline{a} = 0.75$ for a maximum instantaneous frequency separation, r, of 150 kc.

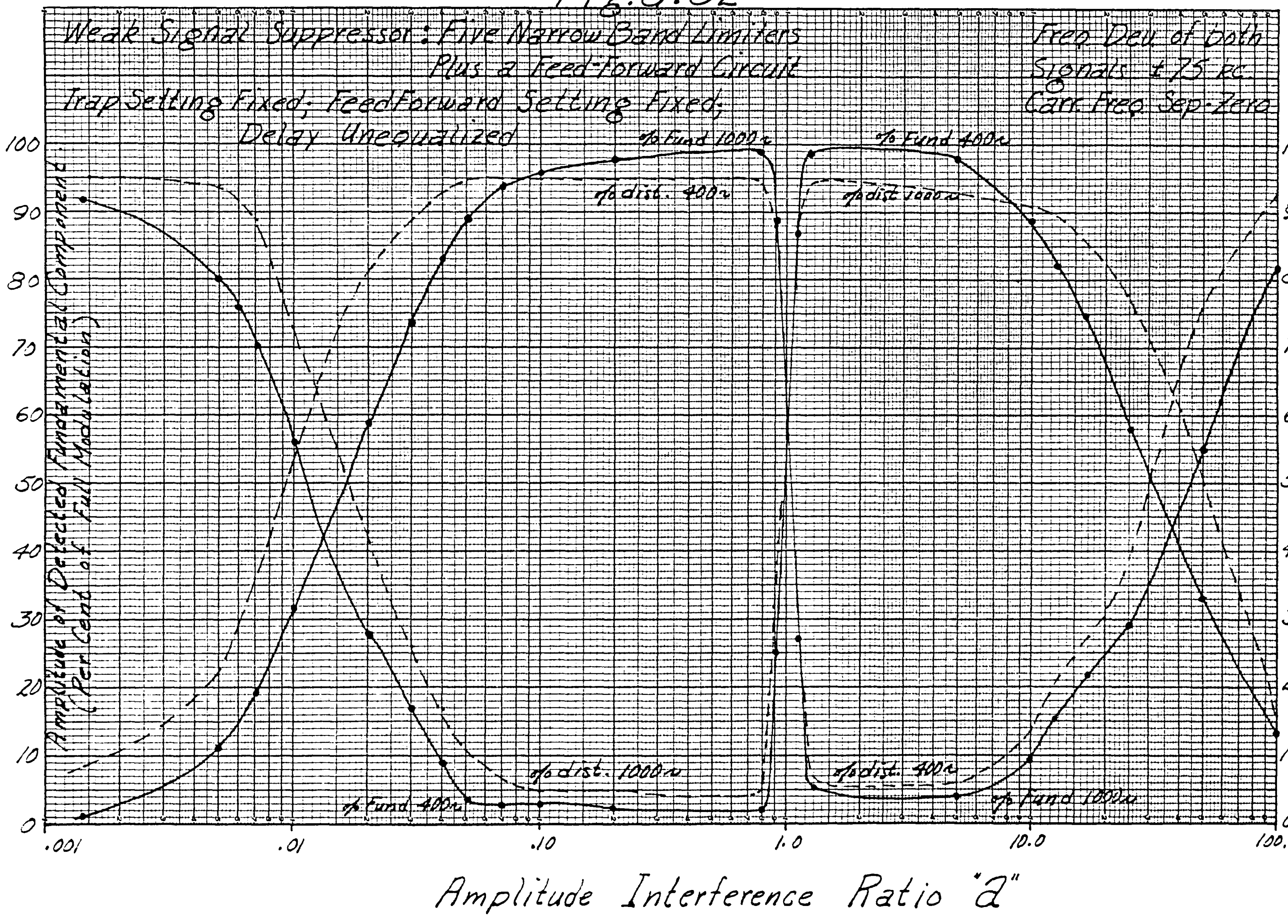
The remaining conditions for which the test was performed were as follows: trap setting fixed; delay unequalized; feedforward circuit optimized at each setting of $\underline{a}$.

We note from the characteristic (Fig. 8.29) that because of the delay unbalance the upper capture region begins to decline prematurely as encountered in the previous tests. However, capture of the desired signal in the capture regions reaches 90% with less than 10% distortion.

The capture transition region centered about $\underline{a} = 1$ is also broadened when compared to the characteristics obtained for the previous tests. The reason for this is that the feedforward circuit begins to fail for $\underline{a}$ in excess of 0.75 and the suppression of the weaker signal afforded by the feedforward circuit diminishes. This in turn, prevents the trap from accomplishing effective suppression of the stronger signal and the resulting capture of the weaker signal by the demodulator begins to decline.

The last capture characteristic for the system is shown in Fig. 8.32 with the accompanying photographs in Fig. 8.33. The weaker signal suppressor for this characteristic consists of the suppressor used for the last test plus three other narrow band limiters ($BW \approx 200$ kc). For this test, the feedforward setting was optimized for maximum weaker signal suppression at one setting of $\underline{a} = 0.5$. The trap was optimized for an $\underline{a} = 0.05$ and the delay for this test was unbalanced. We note that for this characteristic the lower capture region for the 1000 cycle component gradually diminishes from its peak of 99% at $\underline{a} = 0.7$. In comparing this to the previous characteristic Fig. 8.29 we note that the response of the former is more uniform over the same range of $\underline{a}$. Apparently, the optimization of the feedforward at each setting of $\underline{a}$ in the former characteristic minimized the variation over the capture

Fig. 8.32

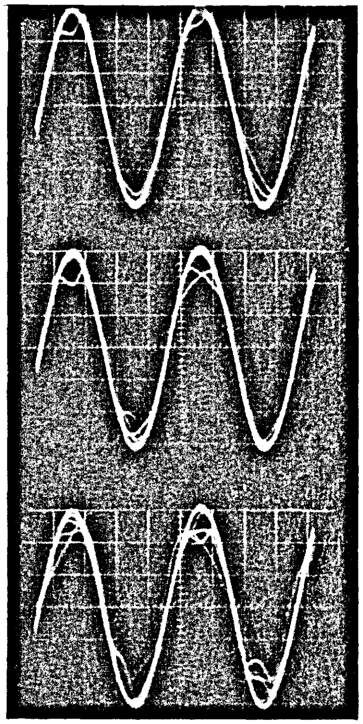


Weak Signal Suppressor: Five Narrow Band Limiters
Plus a Feed-Forward Circuit
Trap Setting Fixed; FeedForward Setting Fixed;
Delay Unequalized
Freq. Dev. of both Signals ±75 kc.
Carr. Freq. Sep.-Zero
% Fund 1000~
% Fund 400~
% dist. 400~
% dist 1000~
% dist. 1000~
% Fund 400~
% dist. 400~
% Fund 1000~
Amplitude of Detected Fundamental Component (Per Cent of Full Modulation)
0
10
20
30
40
50
60
70
80
90
100
.001
.01
.10
1.0
10.0
100.0
Amplitude Interference Ratio "a"

Pic. No. 1

Pic. No. 2

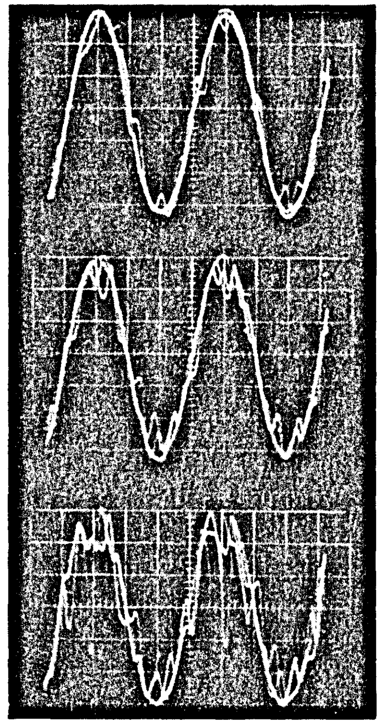
Pic. No. 3



Pic. No. 4

Pic. No. 5

Pic. No. 6



Pic. No. 1	$\underline{a} = 0.8$ % Fund. = 99% % Dist. = 5%	Pic. No. 4	$\underline{a} = 0.8$ % Fund. = 99% % Dist. = 7%
Pic. No. 2	$\underline{a} = 0.1$ % Fund. = 96% % Dist. = 5%	Pic. No. 5	$\underline{a} = 0.1$ % Fund. = 89% % Dist. = 14%
Pic. No. 3	$\underline{a} = 0.05$ % Fund. = 89% % Dist. = 11%	Pic. No. 6	$\underline{a} = 0.06$ % Fund. = 75% % Dist. = 30%
Weaker Signal Modulation 75 kc. deviation 1000 cps. mod. freq.		Weaker Signal Modulation 75 kc. deviation 400 cps. mod. freq.	
Stronger Signal Modulation 75 kc. deviation 400 cps. mod. freq.		Stronger Signal Modulation 75 kc. deviation 1000 cps. mod. freq.	

Fig. 8.33 DEMODULATED WEAKER SIGNAL MESSAGE WAVEFORMS

region.

We note that the addition of the other limiters sharpened the capture transition region centered near $\underline{a} = 1$ because of the more effective suppression of the weaker signal afforded by these limiters.

Other than those mentioned, no significant improvements in the capture of the weaker signal were achieved then the full degree of weaker signal suppression was used.

The effect of the delay unbalance for these characteristics is similar to that experienced for the previous tests. It might be added before concluding this discussion that the reason for not balancing the delay for all the tests was twofold. First, we wanted to see what effect an unbalance in the delay would produce. Secondly, because of the delay equalizer used, it was very difficult to equalize the delay for all the tests performed. Providing the exact amount of equalization would have required deletion of some of the equalizer filters, readjustment of the bandwidths of the remaining filters, and realignment of the entire equalizer. This was done only for the test involving four limiters and for the test involving two limiters. Enough was learned from the study of these two sets of characteristics to warrant the omission of the tests for the equalized delay case involving three limiters or the feedforward circuits.

<u>Summary of Results Deduced from Capture Characteristics</u>

1 - For the simulated conditions of stronger signal predetection (or simulation of weaker signal suppression) the inclusion

of a delay equalizer to compensate for the delay accumulated by the stronger signal message in the interstage filters between the converter and first mixer resulted in improved capture of the weaker signal. Capture of the weaker signal was achieved for an $\underline{a} = 0.02$ with less than 10% distortion for all conditions of modulation of the stronger interfering FM signal and the weaker desired FM signal. The performance improved even more when the deviation or the modulating frequency of the stronger interfering signal was reduced. This indicated that the delay was not completely equalized and that the incidental AM problem still existed although the effects of both were minimized.

2 - The weak-signal capture performance of the system using four narrow band limiters to achieve the weaker-signal suppression varied in accordance with the frequency modulation of both co-channel FM signals. The worst capture performance occurred when both signals were centered in the IF band and frequency deviated to only $\pm$ 35 kc, or one-fourth of a limiter bandwidth. Fig. 8.11 depicts the characteristic for these conditions. For these deviations, the amplitude of the detected weaker fundamental component amounted to only 70% of the full modulation accompanied by 45% total attendant distortion.

The best capture performance occurred when both signals were fully deviated to $\pm$ 75 kc or one-half of a limiter bandwidth. Fig. 8.17 depicts the characteristic for these conditions. For these deviations, the amplitude of the detected weaker signal fundamental component was 90% of the full modulation accompanied by only 17% total attendant distortion.

3 - The weak signal capture performance of the system was <u>not</u> significantly affected by a reduction in the degree of weaker signal suppression (compare Fig. 8.21 with Fig. 8.25) for situations where the delay was equalized. In reducing the number of weaker-signal-suppressor limiters from four to two, the percent capture of the weaker signal dropped from 90% to 80% accompanied by a slight rise in distortion from 17% to 20%. As evidenced from all the capture characteristics, the capture transition region centered near $\underline{a} = 1$ is quite sharp. Near this region the suppression properties of the limiters are very poor but yet we still achieved weaker-signal capture. These results indicate that there is not necessarily any great need for having a large degree of weaker-signal suppression. The improvement in the response was insignificant for the situation where five limiters plus a feedforward circuit were used as a weaker-signal suppressor as compared to the response obtained when just two limiters were used (compare Fig. 8.27 and Fig. 8.32).

4 - Optimizing the trap at each setting of $\underline{a}$ did not improve the capture of the weaker signal in the capture regions (where $\underline{a} \gg \delta$). The only effect in optimizing the trap was to shift the capture transition regions to lower $\underline{a}$.

5 - The effect of an unbalance in the delay equalization resulted in an expected reduction in the range of $\underline{a}$ for which the weaker signal was satisfactorily captured. The reduction in the range of $\underline{a}$ for which capture was achieved was more noticeable when

the modulating frequency of the interfering signal was the higher audio frequency component. However, even with this reduction in capture range, satisfactory capture of the weaker signal was still achieved with less than 10% distortion over a range of a extending from a = 0.1 to a = 0.9.

6 - The effect of an unbalance in the delay equalizer resulted in an unexpected improvement in the capture of the weaker signal (compare Fig. 8.18 with Fig. 8.21, and Fig. 8.25 with Fig. 8.27). For all situations investigated for situations where the delay was unbalanced, the capture of the desired weaker signal was in excess of 95% with a total attendant distortion below 10%.

7 - The use of feedforward circuits as weaker-signal suppressors showed an improvement in the response. With the delay unequalized 99% capture was achieved with 6% distortion. However, the full significance of the advantages of using just a feedforward circuit in place of the entire weaker-signal suppressor unit of limiters was not realized. The reason for this was that an extra limiter was used between the feedforward circuit and the IF amplifier in an effort to preserve the IF filter response without having to realign.

8 - The capture transition region centered at a = 1 is affected only slightly by the inclusion of more weak signal suppressor devices. An improvement in this capture transition region was also improved when the delay was unbalanced by 5 μsecs.

9 - Satisfactory performance can be achieved using just two narrow band limiters as weaker-signal suppressors with a small

unbalance between the delay in each channel of 5μsec. Capture of the fundamental component exceeded 90% of full modulation by a total attendant distortion below 10% (after 3 kc low-pass filtering). This capture was achieved for interference ratio in the range $0.1 \leq \underline{a} \leq 0.9$. (See Fig. 8.27)

Correlation of a Listening Test to Results Deduced from Capture Tests

In order to evaluate the system for a practical situation, the system was connected for optimum response utilizing four narrow-band limiters with the delay equalized. One of the FM generators supplying the weaker signal was externally modulated by a voice signal. The second FM generator supplying the interfering stronger signal was internally modulated by an audio frequency tone with the frequency deviation initially set to $\pm$ 75 kc. The output of the demodulator was amplified and then connected to a loudspeaker. A listening test was performed for various interference conditions; a summary of which is as follows:

1 - With the average level of deviation of the voice modulated weaker FM signal set to about $\pm$ 60 kc and with the deviation of the stronger signal at $\pm$ 75 kc, the received voice signal was intelligible over a wide range of $\underline{a}$ given by $0.05 \leq \underline{a} \leq 0.8$. However, at all times there was background cross-talk and noise.

2 - The average level of deviation of the desired signal was left at $\pm$ 60 kc and the stronger signal deviation was varied over a range from $\pm$ 20 kc to $\pm$ 75 kc. The interference ratio was set to an $\underline{a} = 0.2$. The optimum conditions for reception occurred when

the stronger signal was fully deviated to $\pm$ 75 kc or slightly half a limiter bandwidth. As the deviation of the stronger signal was slowly reduced the quality of the received voice signal declined. At the lowest deviation of $\pm$ 20 kc (one-tenth of a limiter bandwidth) for the stronger signal, the response was quite poor. In addition to the reduction in quality of the voice signal, the background tones were more pronounced. No significant improvement nor reduction in quality was noticed when a was increased.

3 - The stronger-signal deviation was fixed at $\pm$ 75 kc with the amplitude interference ratio set to $a = 0.2$. The deviation of the voice modulated generator was now varied over a range from $\pm$ 20 kc to $\pm$ 60 kc (or roughly one-tenth to one-third of a limiter bandwidth). The response was better than for case 2 as long as the deviation of the desired signal was in excess of $\pm$ 35 kc, (one-fourth of a limiter bandwidth). Below $\pm$ 35 kc the response was poor and of the same quality as that experienced in case 2.

This listening test confirms the results deduced from the capture characteristics. Whenever the deviation of the stronger or the weaker signal is reduced the capture of the weaker signal is affected.

Quality Comparison of the Fixed Trap System Response to the Response of Other Weaker Signal Receivers

In chapter I, we pointed out that two other schemes were tried and have been successful in capturing the weaker of two co-channel FM signals. These are the feedforward receiver and the variable-tuned-trap receiver. The results of both of these systems have been published recently. (1) A summary of the published results is given in order to make comparison.

(A) Feedforward Receiver Summary

The feedforward system was tested for the co-channel case where both signals were deviated to only $\pm$ 20 kc. The modulating frequencies were 1000 cycles and 400 cycles respectively. The percent capture of the weaker signal amounted to 80% with a total attendant distortion of about 20%. The stronger signal message was suppressed to about 10% of full modulation. The range of $\underline{a}$ over which weaker-signal capture was achieved extended from $\underline{a} = 0.1$ to $\underline{a} = 0.7$. For increased deviations of either signal, the response worsened.

It might be mentioned here that these results taken from Ref. No. 1 are from tests conducted of the feedforward in 1958. At the writing of this thesis, the feedforward design is in stages of being perfected by B. Hutchinson at RLE. Hutchinson, in the preliminary tests of his perfected design shows that capture exceeding 90% is attainable with the feedforward over a range of $\underline{a}$ given by $0.1 < \underline{a} < 0.9$. He has also succeeded in lowering the capture transition region to lower interference ratio.

(B) Variable-Tuned-Trap Receiver Summary

The variable-tuned-trap receiver was tested simulating predetection of the stronger signal by the same method used in this thesis.

The effectiveness of weaker signal capture for the dynamic trap system was displayed for various settings of the variable trap bandwidth and trap attenuation. The best response was achieved for a trap attenuation setting of $\delta = 0.04$ with a trap BW = 15 kc. With a trap setting of $\delta = 0.04$, the weaker signal capture transition region was centered near $\underline{a} = 0.04$. The tests were carried out for deviations of $\pm$ 30 kc for both co-channel FM signals centered in the IF pass band. The modulating frequency of the stronger signal was 400 cycles and the modulating frequency of the weaker signal 1000 cycles.

For these conditions the weaker signal was successfully captured to 100% of full modulation for all $\underline{a} > 0.05$. No distortion curves were presented. For increased deviations, the capture performance was not as effective.

(C) Comparison to Thesis Receiver

In comparing the fixed-trap weaker signal capture capabilities to those of the feedforward we note first that the range over which capture of the weaker signal was achieved for the fixed trap was almost double the range for which weaker signal capture was achieved by the feedforward. Furthermore, the worst capture response of the fixed trap system for the most adverse conditions of modulation (both co-channel signals deviated to $\pm$ 35 kc or one-fourth of BW_{lim}) was better than the best capture response of the feedforward for its most favorable conditions of modulation (both co-channel signals deviated to $\pm$ 20 kc (or one-tenth of BW_{lim}).

In comparing the fixed trap performance to the dynamic trap performance, a comparison will first be made for the response of the fixed trap when predetection of the stronger signal was simulated. On this basis, and for the same modulating frequencies of the stronger and weaker signal as used in the dynamic trap tests, the fixed trap response surpasses the best response attainable from the dynamic trap. The capture transition region for the optimized fixed trap was centered at $\underline{a} = 0.0033$ whereas for the dynamic trap it was centered at $\underline{a} = 0.04$. At an $\underline{a} = 0.015$, the distortion was below 10% for the fixed trap. Furthermore these results were obtained for the most adverse modulation deviations of the stronger signal ($\pm$ 75 kc) relative to the trap bandwidth whereas the dynamic trap was tested for its most favorable deviation $\pm$ 30 kc (IF bandwidth = 120 kc).

Comparing the dynamic trap response to the fixed trap response when weaker signal suppressors are used in place of the demodulator-modulator unit, we see that the range over which cpature is satisfactorily achieved for both systems is approximately the same. The capture transition regions for the fixed trap is centered at lower interference ratio in the range $\underline{a} = 0.01$ to $\underline{a} = 0.02$; but, the capture transition region for the fixed trap is not as abrupt as for the dynamic trap. The reason for this is that a wider trap bandwidth for the fixed trap was used and also, because the stronger signal capture characteristics for the demodulator used in conjunction with the fixed trap system, were not as good as the demodulator used with the dynamic trap system.

In situations where the delay was equalized in the tests carried out for the fixed trap, capture of the weaker signal was not as effective as that achieved by the dynamic trap. However, these capture capabilities for the fixed trap system were significantly improved almost to the extent of the capture capabilities of the dynamic trap when the delay was unequalized slightly in the fixed trap receiver. The reduction in the capture region that was suffered by unbalancing the delay for the fixed trap still provided for a region of weaker signal capture that was comparable to that achieved by the dynamic trap.

Chapter IX

Conclusion

In this thesis, we have investigated the feasibility of one form of the static-trap technique for capturing the weaker of two co-channel FM signals. The technique studied, did not require prior demodulation of the stronger signal message (or that which is equivalent for the static-trap technique-complete pre-suppression of the desired weaker signal). Instead, practical weaker-signal suppressors (such as narrow-band limiters, feedforward, etc.) were employed in order to pre-suppress the desired weaker signal relative to the undesired stronger signal amplitude. Fig. 2.4 illustrates the block diagram of the thesis weaker-signal receiver.

Of principal interest in this thesis was to investigate the dependence of the system performance upon the following:

(a) the conditions of frequency modulation of the desired and interfering signals.

(b) the degree of weaker signal pre-suppression required to effect capture of the weaker signal.

However, before an investigation could be made of the feasibility of the proposed system, other problems affiliated with the fixed-trap circuit had to be solved. Fortunately, we were able to rely upon some of the results of a thesis presented previously based on a similar form of the static-trap technique but one which relied upon predetection of the stronger signal message. (4)

As a result of an analysis of the instantaneous frequency response of linear filters and from some observations made of the fixed-trap circuit, it was deduced that the most detrimental spurious effect upon the capture response of the fixed-trap system was the time delay differential existing between the two channels feeding the first mixer stage. Closely related to the deteriorative effect of the time delay was the affect of the phase nonlinearties of the filters on the modulation of the stronger signal in each channel feeding the first mixer stage. Both of these spurious effects tended to reduce the degree of correlation between the stronger signal frequency modulation appearing in each channel. Consequently, removal of the modulation from the stronger signal by the first mixing process was impaired thus preventing the system from functioning properly.

In the analysis of the fixed-trap system the effect of the incidental amplitude modulation of the stronger signal envelope was also considered. This analysis was guided by the conclusions reached by Sheftman in his study of the fixed-trap weaker signal receiver (4). The effect of this incidental AM resulted in AM sidebands straddling the constant frequency signal appearing at the input to the trap circuit. Higher modulating frequencies of the stronger FM signal resulted in further displacement of the sidebands from the center of the trap where the attenuation was greatest. Larger frequency deviations of the stronger FM signals resulted in a greater percentage of incidental AM since the FM signal swept over a larger portion of the filter amplitude characteristics. As a result the AM sidebands were not sufficiently attenuated. Hence, when the stronger signal and weaker signal modulations were restored by the second mixing process, the relative suppression of the stronger signal was insufficient to allow capture of the weaker signal over the full potential range of interference ratio theoretically capable of the system.

Special features had to be incorporated into the design of the fixed trap circuit as a result of this time delay effect and the AM problem. A delay equalizer was improvised to compensate for the delay acquired by the stronger signal message in the suppressor-converter channel filters. The AM problem was partially cleared up by broadening and tuning the filters for ultra-flatness and by providing a wide-band limiter in the converter-first mixer interstage circuits..

Among other problems, it was desireable and necessary to design a stable bandpass trap filter whose center frequency attenuation was achieved passively but whose overall attenuation characteristics (such as trap bandwidth, trap location, and trap attenuation) could be controlled to some degree. The filter in this thesis met these specifications; however, it still had a tendency to drift.

It was also necessary to devote considerable attention to the design of narrow-band limiters whose response could be faithfully depended upon for pre-suppression of the weaker signal over a wide range of interference ratio and for a wide variety of modulation conditions of the interfering and desired signal. In order that the system could be tested for various degrees of weaker signal pre-suppression it was also necessary to incorporate flexibility into the design of these limiters. This feature permitted interchanging of the limiters and quick and easy adjustment of the bandwidths of the narrow-band filters following each limiter.

The last consideration given to the design of the system was related to the optimization of the demodulator stronger signal capture characteristics. This made it possible to confine our attention to the local disorders caused by the fixed-trap circuit and the weaker-signal suppressors.

In order to confirm the considerations given to the improvement of design of the fixed trap circuits, performance tests were conducted for the simulated condition of predetection of the stronger signal (or simulation of the process of complete suppression of the weaker signal in the converter channel of the fixed-trap circuit). The response of the system for these conditions was highly favorable and in comparing to Sheftman's results the response indicated the necessity of providing for delay equalization.

Only after the preceding considerations could we make a meaningful study of the feasibility of the fixed trap system ulitizing limiters to pre-suppress the weaker signal. Our first finding in this feasibility study was that the response did vary, as expected, with the modulation conditions of the desired and interfering signals. It appeared that when the frequency deviation of either signal was reduced relative to the limiter bandwidth, the quality of reception diminished. Although a narrower frequency deviation of either signal resulted in less FM transient distortion introduced by the sluggishness of the trap filter, a narrow frequency deviation also prevented effective pre-suppression of the weaker signal by the narrow-band limiters. Whenever the frequency deviation of either signal was small, the frequency separation, r, between the instantaneous frequencies of either signal was a small fraction of the limiter bandwidth for a larger portion of the modulation cycle of the stronger signal. The result was that the limiters could not provide adequate suppression of the weaker signal and this prevented the fixed trap circuits from switching the roles of weaker and stronger signal effectively. Thus for small frequency deviations with respect to the limiter bandwidth the percent capture of the desired signal was poor accompanied by a considerable level of distortion.

It was also noted that under the conditions where a delay unbalance existed between the two channels feeding the first mixer, there was further deterioration in response when the modulating frequency of the stronger interfering signal was high (greater than 400 cps). For these conditions, capture was not achieved over as wide a range of interference ratio as for similar tests where the delay was balanced.

However, one other important feature was noted concerning this unbalance of delay (unbalance of approximately 5μsec). Although the weaker signal capture region did not extend over as wide a range of interference ratio, in the region for which capture was achieved, the response was notably <u>improved</u> in comparison to the same test where the delay was balanced.

In the feasibility study conducted of the system it was also found that there really was no great need in providing a high degree of pre-suppression of the weaker signal. This became evident after a series of tests were performed for various degrees of weaker signal suppression with the modulation conditions of both interference and desired signals fixed. When only two narrow-band limiters were used as weaker signal suppressors, satisfactory capture of the weaker signal was still achieved. Although there was some improvement, the inclusion of more limiters or even the feed-forward circuit did not result in a significant degree of improvement of the response.

This conclusion was also confirmed by the fact that even in the vicinity of $\underline{a} = 1$, where we knew the limiter suppression would be poor, capture of the weaker signal was still achieved, although considerable distortion prevailed.

One other conclusion regarding the performance was made after a cursory test was performed by varying the trap attenuation factor while the interference ratio $\underline{a}$ was in the range $0.1 \leq \underline{a} \leq 0.9$. In this range of $\underline{a}$, a high degree of trap attenuation was not required for capture of the desired weaker signal. In fact, the attenuation was reduced to 6 db and capture of the weaker signal was still achieved. Thus it is apparent that for $\underline{a}$ in the range quoted, the trap attenuation is not as critical a parameter as was originally believed.

Relative Simplicity of the Fixed-Trap Receiver

In lieu of its concept, the form of the fixed-trap system as studied in this thesis lends itself to simplications of design much more readily than does the dynamic-trap system studied by Rubissow (3). First of all, the dynamic-trap system requires two complete demodulators each containing their own amplifiers, limiters, discriminators, audio filtering and audio amplifiers. The fixed trap system studied in this thesis requires only one demodulator unit. Secondly, the circuitry involved in the dynamic-trap unit is much more complex than that involved in the fixed-trap unit. The dynamic trap requires a highly stable parametric device in a time variant filter to produce the required tracking of the stronger signal. It also requires the provision of a stable variable tuned-trap filter whose attenuation remains substantially constant over the entire range for which the trap is deviated. However, all we require of the fixed-trap filter is the provision of a null at one frequency only. No compromises had to be made in achieving sufficient attenuation with a narrow bandwidth. Since this system does not rely on predetection of the stronger signal message, the need for critical circuits such as a variable frequency local oscillator or time varying filters is obviated.

In comparing the simplicity of the fixed trap system to the feedforward circuit we find that the feedforward circuit is much simpler. In theory the feedforward circuit requires a minimum of three tubes to accomplish weaker signal capture. Although extremely simple in concept, satisfactory results that compare to either trapping technique have not until recently been realized mostly because of some of the practical problems involved with the feedforward circuit. In an effort to improve the performance of the feedforward the present studies mentioned previously in Chapter VIII are leaning towards a more complicated design of the basic feedforward scheme.

Another criterion of comparison in simplicity is that concerning the necessary adjustment of the trap that must be made to render the systems operative. Theoretically, both the dynamic trap and the fixed trap system once aligned and preset should never have to be readjusted. However, both trap circuits drifted considerably as a result of thermal drifts, tube ageing, vibrations, etc. By virtue of its simplicity in concept, the fixed trap system would normally require less readjustment. In fact, the only adjustment required prior to each test of the trap in this circuit was a slight retuning of the trap attenuation and location by the fine tuning controls provided. No other adjustments had to be made for the duration of each test. Actually this also was all that was required of the dynamic trap; however, this design was complicated by dynatron Q multipliers, differential amplifier gain controls, and reactance simulator controls all requiring interacting adjustments.

In comparing the required adjustments that must be made in the fixed trap system to those adjustments that must be made in the feedforward circuit, we note that there is greater ease in adjustment of the feedforward. However, one big disadvantage of the feedforward is that its optimum adjustment varies with each setting of a. This is also manifested in the theory of the feedforward where it is noted that the optimum signal combination ratio varies considerably with the input interference ratio. As we have mentioned, one setting of the trap for a low interference ratio is all the adjustment that is required in the fixed trap circuit. This setting does not vary with the input interference ratio. In addition, for capture of the weaker signal the feedforward circuit is highly critical of the filter alignment immediately preceding and following the two independent, unilaterial, feedforward channels. This is also true to some extent of the alignment of the filters in the two channels feeding the first mixer in the fixed trap circuit; however, the dependency upon alignment is not as great as in the feedforward filters.

Possible Simplifications of Design

Before discussing the simplifications that could be incorporated into the design it is first necessary to point out many of the redundancies that existed in the present design.

1 - The nine tube delay equalizer.

2 - The buffer amplifiers between each narrow-band limiter.

3 - The interstage circuitry between converter and both mixers.

4 - The four tube rejection amplifier providing the trap characteristic.

It was discussed previously that the delay equalizer was improvised from IF amplifiers. We have also seen in the performance of the system that it is not necessary to provide exact delay compensation. In fact, with a delay unbalance of 5 μsecs, improved capture resulted for all the situations investigated; however, the range over which captured was achieved was restricted somewhat.

In addition, satisfactory weaker signal capture was achieved utilizing a weak signal suppressor consisting of only two narrow band limiters. By simplifying the circuitry between the converter and first mixer, the narrow-band limiters themselves, could provide for the delay unbalance of 5μsec and the delay equalizer could be completely omitted.

By proper design of the interstage filters between each narrow-band limiters as McLaughlin did (11), the use of buffer amplifiers between the limiters could be dispensed with. However, because of their excellent limiting properties and higher input impedance, 6BN6 limiters are still recommended.

If a crystal notch filter were used it would eliminate having to achieve the trap filter characteristic utilizing active devices. Furthermore, stability and simplicity would automatically be incorporated into the circuit and once aligned the system would never require readjustments.

By substituting other devices to perform the mixing operations, the 6BE6 pentagrids could be eliminated except in the converter stage. A passive mixer such as a ring-modulator employing matched crystal diodes and transformers could be used in place of the pentagrids. We then would not have to provide amplification no buffering between the converter and the mixers thus eliminating all this interstage circuitry.

Conceivably, the fixed trap system, if designed at a lower IF frequency, could be reduced to a three tube device (not considering the demodulator or IF amplifier) employing passive elements in place of much of the active circuitry used in this thesis and in Sheftman's. The three tubes would consist of the two limiters and the converter. The trap could be produced passively, preferably with a crystal filter. As mentioned, the mixing could also be achieved passively using non-linear elements followed by passive bandpass filters. The use of passive mixers and a passive trap not only eliminates the active devices but it also eliminates any possible regenerative loops in the system. This avoids completely any tendency for the circuits to oscillate.

If in the event passive mixers are not acceptable, the pentagrid 6BE6 could still be used without any buffering provided between the converters and mixers. The mixers would be driven at lower levels and their mixing properties would suffer as a result; however, they could still suffice as acceptable mixers (Sheftman fed his mixers with relatively lower signal levels and his response was good). Using active mixers, the fixed trap system used in conjunction with narrow band-limiters would require only five tubes which would still represent considerable simplification.

These simplifications would put the fixed trap circuit on a par with the feedforward circuit in simplicity. However, we would expect an improved weaker signal capture over that of the feedforward. The reason for this is that unlike the feedforward, the fixed trap principle of operation for capturing the weaker signal does <u>not</u> <u>place</u> <u>sole</u> reliance on the weaker signal suppression properties of the narrow-band limiters.

Suggestions for Future Work

This thesis is a culmination of the basic research aspects of the fixed-trap system. Its feasibility has been proven; remaining, is the circuit development and engineering for simplifying the system and for making it more reliable.

The recommendations given in the previous section for utilizing passive components for the mixers and trap filter in conjunction with a three tube fixed trap circuit is a tempting engineering developmental problem and deserves further consideration.

The possibility of using this fixed trap technique for a half-channel duplex application was not investigated in this thesis. However, Sheftman investigated this under the simulated conditions for which he conducted his tests and the results were encouraging. From all indications of the performance of this system, half-channel signals could be easily suppressed by the limiters and the strong signal suppression properties of the trap could then switch the roles of weaker and stronger signals. This leads to the belief that the system would perform quite well for these conditions of interference. It would be desirable to actually see how the system could respond under these conditions.

Another test that was not performed but which would be of considerable interest, is the interference situation where we have a carrier signal that is frequency modulated by noise. Such a problem might arise in a channel jamming situation. It would be interesting to investigate whether or not the fixed trap system could capture the weaker signal for these conditions of interference.

Coordinated with this investigation would be to determine how much delay unbalance could be tolerated and to what degree must the limiters pre-suppress the weaker signal.

Lastly, it would be desirable to investigate the weaker signal capture capabilities of this receiver in the presence of a high level of incoherent channel noise. Many different problems would have to be considered relating to the preservation of an extremely high degree of correlation between the two channels feeding the mixers. In fact different components (passive, low noise devices) may have to be used to perform the mixing and converter functions because of the high level of additive noise in these active mixers. In addition, there is also a coherency problem associated with the noise as it is processed by a limiter and followed by a filter introducing delay. All these are interesting problems and are worth some further consideration.

(1) E. J. Baghdady, "New Developments in FM Reception and Their Application to the Realization of a System of "Power-Division" Multiplexing", IRE Transactions on Communications Systems (PGCS) pp. 147-161, September, 1959.

(2) R. H. Small, "A Single-Channel, Two-Carrier FM Multiplex System," S.M. Thesis, Department of Electrical Engineering, M.I.T., June 3, 1958.

(3) G. J. Rubissow, "Dynamic Trap for Capture of Weaker Signal in FM Interference," S.M. Thesis, Department of Electrical Engineering, M.I.T., January 31, 1958.

(4) F. I. Sheftman, "A Fixed-Trap System for Capturing the Weaker of Two Co-channel FM Signals," S.M. Thesis, Department of Electrical Engineering, M.I.T., September 5, 1958.

(5) E. J. Baghdady, "Theory of Low-Distortion Transmission of FM Signals Through Linear Systems," Technical Report 332, Research Laboratory of Electronics, M.I.T., July 30, 1957.

(6) E. J. Baghdady, "Interference Rejection in FM Receivers," Technical Report 252, Research Laboratory of Electronics, M.I.T. September 24, 1956.

(7) J. Granlund, "Interference in Frequency Modulation Reception," Technical Report 42, Research Laboratory of Electronics, M.I.T. January 20, 1949.

(8) L. C. Bahiana, "Nonlinear Properties of FM Limiters," Technical Report 350, Research Laboratory of Electronics, M.I.T. May 29, 1959.

(9) E. J. Baghdady, "Signal-Cancellation Techniques for Capturing the Weaker of Two Co-channel FM Signals," presented at the International Conference on Electro-Magnitic Wave Propagation, Liege, Belgium, Unpublished.

(10) R. M. Wilmotte, "Reception of an FM Signal in the Presence of a Stronger Signal in the Same Frequency Band, and Other Associated Results," Instn. Elect. Eng. Proc. Vol. 101, Pt. 3, 69-75, March 1954.

(11) R. J. McGlaughlin, "A Study of FM Capture Effects," S.M. Thesis, Department of Electrical Engineering, M.I.T., June, 1958.

(12) F. Langford-Smith, "Radiotron Designer's Handbook," 4th Edition, Reproduced and Distributed by Radio Corporation of America, 1953.

(13) G. E. Valley and H. Wallman, "Vacuum Tube Amplifiers, McGraw-Hill, 1948.

(14) "Reference Data for Radio Engineers," 4th Edition, International Telephone and Telegraph Corporation, 1956.