INTERFERENCE AND NOISE REJECTION IN F. M. RECEIVERS

by

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Accepted by

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Submitted to the Department of Electrical Engineering on May 20, 1957 in partial fulfillment of the requirements for the degree of Master of Science.

ABSTRACT

Recent theories on the effective use of positive, narrow band feedback around the limiter in f.m. receivers to reduce interference and noise, are tested experimentally. Quantitative data is shown on just how much improvement in capture ratio can be expected with simple circuits, and how much quieting is obtained due to squelch action.

A discussion is made on the application of these techniques to the commercial f.m. industry.

Thesis Supervisor: Elie J. Baghdady
Title: Assistant Professor of Electrical Engineering
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INTRODUCTION

The problem of interference rejection in frequency modulation receivers is rapidly becoming more and more important in their design. As has been shown by Paananen\(^1\), there are numerous cases in the more populated regions of the country where co-channel and adjacent-channel stations overlap in their fringe areas. The work done by Arguimbau and Granlund\(^2\) reveals how multipath signals can greatly hamper the use of f.m. for communications over long distances via the ionosphere. This certainly holds true not only at the short wave frequencies used in their tests, but also at very-high and ultra-high frequencies where multipaths can occur by reflections from buildings, mountains, airplanes, and other surfaces.

It is well known that an f.m. receiver will exhibit a capture effect; that is, it will tend to lock on to the stronger of two interfering signals in such a way as to reduce or totally eliminate the amount of distortion caused by the weaker signal that one might normally expect from a.m. experience. The measure of this effect, known as the capture ratio, was shown by Arguimbau and Granlund to be a function of receiver design.

It behooves the f.m. industry to design the standard f.m. broadcast receivers to take full advantage of this capture effect. This would allow more f.m. stations to operate in the present band, and would reduce annoying reception due to multipaths. The use of f.m. in mobile communications installations would become infinitely more practical, for the consequences of multipath interference with the poor
capture ratio equipment that is now on the market usually turns
people to the use of amplitude modulation. Arguimbau and Gran-
lund postulated the use of receivers having ultra-wide limiter and
discriminator bandwidths for most effective capture. These specifications
are impossible to obtain in a competitively-priced receiver, so
another way to achieve improved performance remained to be found.

In 1952, Wilmotte\textsuperscript{3} presented a theory, whereby negative, band-
limited feedback around the amplitude limiter in an f.m. receiver
will act in such a way as to increase the capture ratio. This paper,
however, is not at all quantitative, no circuit is shown, and the
vector diagram that was used to prove the point is incorrect and very
misleading. In 1953, Onder\textsuperscript{4} discredited Wilmotte's theory, bas-
ing his point on experimental evidence that this type of feedback led
to no improvement in capture ratio at all.

Baghdady,\textsuperscript{5} in 1956 went into a thorough discussion and analysis
of interference in f.m. receivers, and from this was able to pre-
scribe conditions whereby narrow banding of cascaded limiters in
the receiver would significantly improve the capture ratio for a given
discriminator bandwidth. It was also concluded that band-limited
positive feedback around an amplitude limiter will effectively accompl-
ish the same mission. This latter theory was tested by Emmons,\textsuperscript{6}
and qualitative results showed it to be valid.

As a by-product of this positive feedback, Baghdady pointed out
that a squelch effect would be realized under the no-signal condition.
This, in itself, would be a boon to the f.m. radio industry, for an objection that listeners have to f.m. is the disturbing interstation noise that accompanies tuning.

This thesis is an application of Baghdady's theories on limiter feedback to an f.m. receiver of inexpensive design, in an attempt to get quantitative data on just how much improvement in capture ratio can be expected with simple circuits, how much quieting is obtained due to squelch action, and how applicable the circuit would be to a competitive radio market.
THEORETICAL BACKGROUND

In order to gain insight into the particular problem at hand it is important to carefully study the nature of the capture effect so as to appreciate what must be done in the receiver, in order to make this effect most pronounced. The obvious first question that comes into our minds is, "Why should an f.m. receiver exhibit a capture effect in the first place?"

Let us consider two f.m. signals arriving at an f.m. receiver that are close enough in frequency to be passed through the r.f. and i.f. linear amplifiers with no distortion. If the modulation of each signal is sufficiently slow compared with the i.f. bandwidth, we can regard each signal as the variable frequency resultant of a spectrum, and describe a dynamic situation in quasi-static terms. Thus, we can say that the two signals are described by \( \cos pt \) and \( a \cos(p+r)t \), where \( a<1 \) and \( r \ll p \).

The signal appearing at the input to the limiter is \( f(t)=\cos pt+a \cos(p+r)t \). This signal can be studied by the superposition of rotating vectors as shown in Fig. 1. This approach is based on the relationship

\[
f(t) = \text{Re} \left[ e^{ipt} + ae^{i(p+r)t} \right].
\]

The action of the limiter will produce a resultant signal that is of constant amplitude and whose instantaneous frequency is the time rate of change of its phase angle. Since the amplitude limiting does not effect phase angle, the instantaneous frequency will be just \( \frac{dp}{dt} \) or \( p + \frac{da}{dt} \).
FIG. 1

Phasor diagram showing the relationships between two signals of \( r \) radians/sec frequency difference, the stronger of amplitude 1 and the other of amplitude \( a \). It is readily seen that the resultant, \( R \), will have an instantaneous frequency of \( \frac{d\beta}{dt} \) whose average value will be just \( p \) radians/sec.
\[ \frac{da}{dt} = \frac{d}{dt} \arg \left[ 1 + ae^{jrt} \right] \]

\[ = \frac{d}{dt} \Im \left[ \ln (1 + ae^{jrt}) \right] \]

\[ = \Im \left[ \frac{iar e^{jrt}}{1 + ae^{jrt}} \right] \]

\[ = \Re \left[ \frac{r e^{jrt}}{1 + ae^{jrt}} \right] \]

\[ = r \left[ \frac{a \cos rt + a^2}{1 + 2a \cos rt + a^2} \right] \]

thus:

\[ \frac{d\beta}{dt} = p + r \left[ \frac{a \cos rt + a^2}{1 + 2a \cos rt + a^2} \right] \]

It is evident from the phasor diagram that, as long as \( a \) is less than 1, the average value of \( \frac{da}{dt} \) will be zero, and the resultant signal frequency will be the same, on the average, as that of the stronger signal at the input to the limiter.

A plot of the function \( p + \frac{da}{dt} \) is shown in Fig. 2. It is seen that this function will have its maximum frequency excursion at the times when \( \cos rt \) is equal to -1. Thus the maximum change in frequency is of the amount \( -r \frac{a}{1-a} \).

It is standard practice to frequency-modulate a transmitter with a peak deviation of ± 75 kc. Thus, \( r \), the frequency difference between the two signals will be supersonic most of the time. Since the average of the excursions about the frequency of the stronger signal is zero, and the spikes recur at the usually supersonic difference-
A plot of the function $p + \frac{da}{dt}$. Notice that the area below the curve is equal to the area above the curve for each cycle. Because of this, the frequency $p$, which is the desired information, can be maintained under interference conditions.
frequency rate, the interference in an f.m. system will be inaudible most of the time. Fortunately, it is also standard practice to pre-emphasize the higher frequency audio components in an f.m. transmitter with a 75 msec. differentiating network. To equalize the system receivers are equipped with 75 msec. integrating networks which have upper half-power points at 2.12 kc. Therefore, as far as interference is concerned, the audio bandwidth is 2.12 kc. This makes the fraction of time we might expect audible interference proportionally less than if the full 15 kc. audio bandwidth was equally utilized.

The foregoing conclusions sound remarkable until we realize that they were based on the assumption that an ideal discriminator was being used. If we had a discriminator that would be able to follow the frequency spikes with no distortion, then the average frequency, \( p \), would be maintained, and all would be well. Let us see how good this "ideal" discriminator would have to be. If we assume the condition of maximum spike height, which arises when \( r \) is equal to one i.f. bandwidth, we can construct Table 1 to show the discriminator bandwidths necessary for the reproduction of spike patterns caused by undesired signals of various relative amplitudes.
\[
\text{Bandwidth necessary} = 2r \frac{a}{1-a} + r
\]

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<tr>
<td>.95</td>
<td>5850.0</td>
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</table>

**Table 1**

Maximum Bandwidth Necessary for Spike Reproduction

\[ r = 150\text{kc} = \text{one i.f. bandwidth} \]

In the computation of these bandwidths, we recognize that the particular case so far discussed deals with an interfering signal which is \(r\) radians/sec higher in frequency than the desired signal. This causes the spikes produced to point downward. A similar argument shows that an interfering signal that is lower in frequency will produce spikes which point upward. Consequently, the bandwidths required are equal to twice the maximum spike excursion plus the i.f. bandwidth.

Arguimbau and Granlund built a receiver with limiter and discriminator bandwidths of 6 mc. to test the theory on which the
tabulated figures are based. They were able to successfully separate signals which differed in amplitude by only 5%. This was all well and good, but producing receivers with these bandwidths would be an impractical undertaking. Many problems appear when broadbanding is attempted, a major one being the gain that must be sacrificed in return for bandwidth in both the limiter and discriminator circuits.

In a comprehensive work on interference rejection in f.m. receivers, Baghdady showed that wide band limiters and discriminators were not necessary for high capture ratio. He showed that feeding the limiter output into a narrow band filter and then re-limiting did nothing more than narrow down the spike excursion. The same average frequency was still maintained. Not only could this be done once, but many times, and each time it was done, the spikes became less and less prominent. If a sufficient number of narrow-band limiting stages are cascaded, the discriminator bandwidth necessary could be reduced to one i.f. bandwidth. He established criteria for achieving optimum capture ratio in a given number of stages by controlling filter bandwidths. Needless to say, this was a great boon to the art, for widebanding problems could be done away with.

However, a number of cascaded, narrow-band limiters prove an expensive proposition to receiver manufacturers, and Baghdady realized this. He, therefore, suggested the use of narrow band positive feedback around the limiter stage in an f.m. receiver as a practical method of achieving the same results.
The band-limited feedback loop in conjunction with the limiter can be looked at as a high-$Q$ limiter which achieves the spike modification in one stage that would only be equalled by many stages of conventional limiting and filtering.

Due to the fact that positive feedback leads to self-oscillation in the absence of an input signal, an important by-product of this system emerges. If the loop is allowed to oscillate under the no-signal condition, a squelch effect will be produced that will do away with the annoying interstation noise prevalent in f.m. receivers.
EQUIPMENT

A. The Receiver

In order to carry out the experimental work, either a commercial f.m. receiver or a reasonable equivalent seemed desirable. Fortunately, similar studies had been done by Emmons, and his equipment was available and quite suitable for preliminary work.

The circuit diagram of the modified basic unit is shown in Fig. 3. Two stages of i.f. amplification at 10.7 mcs. were employed to insure limiter saturation with the available signal source. Two limiters were necessary to provide full action over the wide variations in amplitude caused by conditions of interference. The limiter and discriminator coils and transformers were hand-wound. Provisions were made for varying coupling between primary and secondary windings. Not shown on the schematic are the various shields, heater wire bypasses, and high voltage bypasses that were found necessary to stabilize the unit.

The design of the feedback loop is of fundamental importance to this study. It is therefore appropriate to devote a separate section to a detailed discussion of it.

B. The Adder

Since this work deals with interference problems, a method must be devised to control the amounts of signal and interference, and their ratios in a satisfactory way. The f.m. signals are obtained from two signal generators. The interaction between the two is
CIRCUIT DIAGRAM OF THE
FIG. 3

THE FIRST EXPERIMENTAL UNIT
minimized by connecting their outputs to the adder circuit shown in Fig. 4.

For the proper matching, the adder was designed to work from and into a 50 ohm coaxial line.

C. Laboratory Setup

A block diagram of the experimental setup is shown in Fig. 5.

It might be pointed out that the use of one univerton is sufficient, if both signal generators are fed into it. However, fine frequency adjustment of the Boonton is very difficult because of poor vernier calibration and bandspread. It proved much easier to use the two univerters which afforded excellent fine frequency control. The audio amplifier with speaker was used as a constant aural monitor of the output signal. It proved quite helpful in determining capture conditions.

The scope was used as an output waveform monitor, and in connection with the f.m. generators as a dynamic i.f., and discriminator curve plotter.

Two multirange VTVMs were used in the measurements. One of these was a special ac and r.f. VTVM with a scale that was calibrated as low as .05 volts. This was necessary to measure low 10.7 mc levels.

The RLE power supply was used and the plate voltage applied to the receiver was set at 220 volts. This supply had a variable 0-150 volt negative supply that proved of use in controlling the feedback.
FIG. 5

BLOCK DIAGRAM OF EXPERIMENTAL SETUP
A General Radio 805C signal generator was used as a frequency standard and as a marker generator for dynamic measurements, and a General Radio Wave Analyzer was used for capture ratio measurements.

An r.f. and converter unit that tunes from 88 to 108 mcs. was used for tuning in local broadcast stations in order to provide realistic signals for listening tests.
THE FEEDBACK LOOP

In designing the feedback loop, certain specifications had to be met. They are listed below.

(1) Zero phase shift at center frequency around loop.

(2) Controllable bandwidth.

(3) Controllable gain.

(4) Little reactive loading of circuits.

(5) Good isolation from the rest of the unit to cut out any leakage feedback paths.

A further specification is that it must fit in with the present limiter. Let us first determine the phase conditions necessary in the feedback device, then work from there. Reference to the schematic diagram (Fig. 3) shows that the total phase shift from the grid of the first limiter to the plate of the second is $360^\circ$, or $0^\circ$. Thus, our phase condition will be fulfilled if we have zero degrees in the feedback device. This immediately suggests a simple tuned parallel resonant circuit. However, as we glance down our specification list, we see that we would, among other things, want considerable control over feedback gain and bandwidth. The simple tuned circuit does not afford this control easily, and once more, the circuit is a bilateral device, which would tend to confuse the measurements.

For good control of bandwidth, it is worthwhile to use a conventional
double-tuned, inductively coupled circuit. If we provide this circuit with variable tuning and coupling, we can get a wide range of bandpass characteristics. This, however, does introduce additional phase shift, since the double-tuned circuit exhibits a shift of $90^\circ$ at its center frequency. Let us see how we can fit this into the loop.

The control of gain looks to be a formidable problem. It is very difficult with coupling circuits alone because with each change in gain there is a change in frequency and bandwidth. Since we would like to make these factors as independent of each other as possible, it seems that the use of a vacuum tube amplification stage would be very much in order, especially since it will also solve our unilateral transmission problem.

Now we have a total of $90^\circ$ plus $180^\circ$, or $270^\circ$ phase shift in our feedback device. An additional $90^\circ$ shift is necessary. A look at Fig. 3 shows that this condition will be fulfilled if we feed back to the primary of the first limiter transformer, instead of the secondary. The final circuit diagram of the feedback loop is shown in Fig. 6.
FIG. 6

CIRCUIT DIAGRAM OF THE FEEDBACK LOOP THAT WAS BUILT INTO THE BASIC UNIT
OPERATION AND MEASUREMENTS

A considerable amount of revision was necessary on the design of the equipment to get stable operation. Since the effect of feedback was the main consideration, no tendency toward feedback other than in the desired path and controllable by the desired control, could be tolerated. An extensive program of shielding, bypassing, and shortening of leads finally achieved this degree of stability.

Once the basic unit was operating smoothly, the feedback stage was added, and alignment of the system was attempted. This operation proved to be a major stumbling block for it quickly became evident that there was too much interaction between the variables to draw many reliable conclusions as to the cause of an effect. The feedback loop was composed of two double-tuned, variable coupling transformers, and the primary of a third. Variation of any one of the adjustments on these transformers affected feedback gain, feedback bandwidth, and phase shift around the loop. Furthermore, variation of gain of the feedback tube by control of its grid bias, varied the interelectrode capacitance of the tube, and thus the tuning of its associated tanks. Finally, variation of input and feedback signal strengths, changed the conduction angle of the limiters. This had the effect of loading down and detuning the tank circuits so that both bandwidth and frequency of zero phase shift (oscillation frequency of loop) was affected.

It is interesting to think of this circuit in terms of an oscillator which is controlled by a synchronizing signal. It is evident that
if the sync signal is at the center frequency of oscillation, its amplitude can be essentially zero, and as the sync signal begins to differ in frequency from this value, it must become progressively stronger in order to pull the oscillation along with it.

The controllable range of this type of system is smaller than that of a conventional oscillator because of the narrow band filter that is inserted in the feedback path. The filter, besides limiting the feedback range by its bandpass nature, also serves to steepen the slope of the phase characteristics of the circuit around the center frequency. This disrupts the conditions necessary for oscillation and causes the sync signal to lose control of the oscillation at much smaller values of deviation than would be normally expected. Consequently, a strong control signal is required to lock in the oscillator over the complete i.f. range. However, it was found that high intensity signals, even though they adequately control the oscillator, also swamp out the limiters to so great an extent, as to make the effects of the narrow band feedback negligible. The solution of this problem is to find the "happy medium" signal, and to design the circuit to apply only this level of input to the loop stages.

Even with all these complications, the circuit was able to exhibit a remarkable amount of interference rejection during listening tests, and of course the squelch effect was quite pronounced.

Figures 7-10 show various oscillograms taken of the units characteristics. The disconcerting factor about these results was
FIG. 7

The dynamic i.f. and limiter characteristics of the modified Emmons unit. The total frequency excursion shown is 480 kcs, and is centered at 10.7 mcs. Notice that positive bandlimited feedback increases i.f. gain and improves skirt selectivity. Horizontal and vertical scales are the same for both pictures.
The effect of feedback on an interference condition between two 1000 cycle modulated signals, one one-half the amplitude of the other, and 50 kcs. lower in center frequency. Both signals are deviated ± 75 kcs.
FIG. 9

The effect of feedback on an interference condition between 1000 cycle modulated signal, and a weaker signal deviated at a 50 cycle rate. Notice how the asymmetrical situation becomes symmetrical with addition of feedback due to the effect of the feedback on tank circuit parameters.
FIG. 10

The effect of feedback on the random noise emerging from the receiver under no-signal condition, simulating in-between-station conditions. The decrease in amplitude as measured on a VTVM was to 1/15 or a reduction of 94%.
that when the circuit was tuned so as to have the characteristics shown in Fig. 7, the apparent effect of feedback on interference rejection was not the maximum that could be obtained if the circuits were tuned for minimum interference directly. Also, as seen in Fig. 9, the effect of feedback seemed to change an unsymmetric interference condition into one that was symmetric. This was due to the aforementioned changes in tank parameters that appeared when the value of grid bias on the feedback tube was altered. It seemed impossible to find a cause of these troubles due to the interaction between all elements in the feedback loop. After many long hours of tinkering with the circuit, it was finally decided to build up a new unit, especially designed to minimize the circuit interdependencies that had made quantitative measurements on the previous circuit meaningless.
DESIGN OF NEW UNIT

In designing the new unit, we must think in terms of the limitations of the old circuit, and endeavor to circumvent those same problems.

The limitations were:

1. The circuit worked at its best for only one level of input signal.

2. Tuning of the discriminator transformer detuned the feedback loop.

3. Measurement of dynamic i.f. characteristics before and after limiting could not be easily done without disturbing the feedback loop.

4. There was very little isolation of the critical feedback loop from the rest of the circuit.

After much careful thought in connection with the preceding points, the unit whose circuit diagram appears in Fig. 11 was constructed. In order to keep the tuned circuits as simple as possible, standard miniature f.m. i.f. and discriminator coils were used throughout.

It is seen that the feedback loop is entirely isolated from the rest of the receiver by limiters, and that feedback in this case is only around one stage. The use of the limiter preceding the loop successfully solves the problem of feeding the circuit only its optimum input level. This stage was designed to saturate at just this value. The grid r-c circuit of the pre-limiter makes an
FIG.

Schematic Diagram of the

6AU6 I.F.

6AU6 I.F.

6AU6 I.F.

6AU6 PRE-LIMITER
ALL RESISTORS IN OHMS
ALL CAPACITORS IN MFD.

FREQUENCY OF ALL TANK CIRCUITS
IN VICINITY OF 10.7 MCS.
ideal point from which to observe the i.f. characteristics before limiting.

In the same manner, the post-limiter r-c circuit provides for dynamic measurements of the signal after the feedback has taken place, in addition to its isolation of the loop from the discriminator transformer. This latter point was important, for it would be impossible to use the primary of the discriminator transformer in the circuit effectively if stagger tuning of the limiters were called for.

The circuit shown was built up, and after the usual amount of debugging, was finally gotten to operate in a stable manner. The first order of business was the alignment of the i.f. stages to give a flat frequency response over the 150 kc range required in a conventional f.m. system. This was accomplished by a stagger tuning procedure, and the resulting dynamic curve as measured at the grid of the pre-limiter is shown in Fig. 12.

The feedback loop circuits were aligned next. Various values of capacity coupling were inserted between the feedback tube transformer secondary, and the limiter grid circuit in order to obtain the value that would serve to feed back enough signal level, yet not couple the two tanks together in such a way as to have the tuning of one affect the tuning of the other to any great extent.

At these frequencies it is usually speedier to adopt this cut and try approach rather than attempt to calculate the circuit parameters from theory.
FIG. 12

Dynamic i.f. response curve measured at the r-c circuit of the pre-limiter. The total deviation is 480 kcs, and the distance between peaks is 130 kcs.
The dynamic bandpass characteristics of the combined i.f. and limiter stages measured at the grid r-c circuit of the post-limiter is shown in Fig. 13. It is seen that the addition of feed-back causes the characteristic to narrow beyond the 150 kc bandpass needed for the desired signal. The tearing of the curve at its extremeties is caused by the loss of control of the oscillating loop at distances greater than approximately 200 kcs. from the center frequency. This, of course, has no effect on regular operation which extends only ± 75 kcs. What cannot be easily seen from these curves is that the gain of the system in the center 150 kc region is increased, due to the regenerative feedback, making for more thorough limiting. The circuit was now ready for capture measurements.
Dynamic response curve over a 480 kc. range of both the i.f. and limiter stages as measured at the r-c circuit of the post-limiter. Horizontal and vertical scales are the same in both pictures.
MEASUREMENTS OF THE EFFECT OF FEEDBACK
ON THE CAPTURE RATIO

The method chosen for measurement of capture effects is outlined below. It was felt that this technique was the most straightforward of any considered, and provided for the most easily interpreted data. Reference to Fig. 5 will aid in understanding the procedure.

1. The two Boonton oscillators were set at the same frequency by feeding both through the receiver and observing the beat note.

2. The oscillators were set so that their output amplitudes were equal for equal attenuator setting by the following technique

   (a) Deviation on both was adjusted to be \( \pm 75 \text{kcs.} \) (as shown on the instrument meter). One was modulated at 5000 cps, the other at 400.

   (b) The output of one oscillator was set at a convenient value that was below the limiter threshold of the receiver, while the other’s attenuator was set at zero.

   (c) The value of negative grid bias developed at the r-c circuit of the pre-limiter was noted on the VTVM.

   (d) The generators were then put in the reverse position, and the r.f. level control on this second generation was adjusted in such a way as to give the same amount of grid bias as shown on the VTVM when its attenuator was set at the same convenient value.

3. The generator whose output was being deviated at a 5000 cps rate, was used as the desired signal, and was set at a reference point of 3,000 microvolts.
4. The generator, with the 400 cycle modulated signal, was used as the interference. Its output level was varied to provide a variety of conditions.

5. The wave analyzer and oscilloscope were connected to the output of the discriminator just after the de-emphasis network.

6. For the various settings of input signal ratios, as taken directly from the generator attenuators, the wave analyzer was adjusted to measure the amplitudes of the 5000 cycle and 400 cycle signals present. This was done both with and without feedback.

7. At the same time, pictures were taken of the scope patterns to serve as visual evidence of the effects of feedback.

The data taken is shown in Table 2 and the pictures are displayed in Figs. 14-20. It is seen from Table 2 that there is a definite increase in the capture effect exhibited by the receiver ranging from a factor of 6.5 at an input ratio of 15 to 4.8 at an input ratio of 3.3. At ratios of 3 and less, measurements on the wave analyzer become progressively meaningless, for the amount of hash produced by the interference became more and more significant as can be seen in Figs. 14 and 15 in the non-sinusoidal 400 cycle ripple that is present with feedback. In order to further investigate this hash, scope pictures were taken of 5000 cycle interference on top of a 400 cycle desired signal at high interference ratios. These are shown in Figs. 19 and 20. It is seen that the majority of the 5000 cycle ripple is removed after feedback, but
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<td>4.8</td>
</tr>
<tr>
<td>3.0</td>
<td>8.3</td>
<td>50</td>
<td>6.0*</td>
</tr>
<tr>
<td>2.1</td>
<td>4.5</td>
<td>40</td>
<td>8.9*</td>
</tr>
<tr>
<td>1.9</td>
<td>3.6</td>
<td>22</td>
<td>6.2*</td>
</tr>
<tr>
<td>1.5</td>
<td>2.2</td>
<td>6.7</td>
<td>3.0*</td>
</tr>
</tbody>
</table>

*See text

Table 2

Ratios Measured on Audio Wave Analyzer between 5000 cps Modulated Desired Signal and 400 cps Modulated Interference.
the high frequency spikes remain. These produce a wide audio interference spectrum, and therefore cannot be taken into account on a single frequency wave analyzer. It is believed that the non symmetrical appearance of these spikes is due to slight asymmetry of the i.f. response curve.
FIG. 14

EFFECT OF FEEDBACK ON INTERFERENCE

Desired (stronger) signal - 5000 cycles
Interference - 400 cycles
Amplitude Ratio 1:5:1
FIG. 15

EFFECT OF FEEDBACK ON INTERFERENCE

Desired (stronger) signal - 5000 cycles
Interference - 400 cycles
Amplitude Ratio 2:1

(a) Without Feedback

(b) With Feedback
FIG. 16

EFFECT OF FEEDBACK ON INTERFERENCE
Desired (stronger) signal - 5000 cycles
Interference - 400 cycles
Amplitude Ratio 3:1

(a) Without Feedback

(b) With Feedback
FIG. 17

EFFECT OF FEEDBACK ON INTERFERENCE

Desired (stronger) signal - 5000 cycles
Interference - 400 cycles
Amplitude Ratio 6:1
FIG. 18

EFFECT OF FEEDBACK ON INTERFERENCE

Desired (stronger) signal - 5000 cycles
Interference - 400 cycles
Amplitude Ratio 10:1
EFFECT OF FEEDBACK ON INTERFERENCE

Desired (stronger) signal - 400 cycles
Interference - 5000 cycles
Amplitude Ratio 1:5:1
FIG. 20

EFFECT OF FEEDBACK ON INTERFERENCE

Desired (stronger) signal - 400 cycles
Interference - 5000 cycles
Amplitude Ratio 2:1
THE SQUELCH EFFECTS DUE TO FEEDBACK

Because of the higher gain of this unit, the quieting effect in-between stations was much more pronounced. The oscilloscope patterns are exactly the same as for the original unit as shown in Fig. 10. Voltage measurements of the noise showed a decrease of 97% when the feedback was applied and the limiter broke into oscillation.

The use of an oscillating limiter brings with it the problem of weak signal reception. If the received signal level is very low, it will not be able to control the oscillation, and it will either be squelched itself or else it will create an in-between-situation where it will semi-control the system and produce a much distorted version of itself at the receiver output.

This problem was solved in the unit to a great extent by the use of the pre-limiter. The limiting can be set to have a threshold below which a signal would not be well received in any case, due to noise. Further design can bring any signal above this threshold value up to a level where it will adequately control the oscillating stage. Since this pre-limiter saturates at this level, only one level of feedback is necessary in the oscillating stage for all signals.
GENERAL COMMENTS ON OPERATION

The receiver was connected to an 88-108 mc/s r.f. and converter unit so as to enable it to tune in standard broadcast signals. The first thing that strikes the operator is the complete lack of hiss in-between stations. One gets the feeling that the receiver lacks in sensitivity because of its muteness. Tuning across the band quickly changes ones opinion as station after station appears. The tuning-in of a signal is a bit more annoying than that in a conventional receiver due to the distortion produced by the semi-control of the oscillating stage when one is tuned off center frequency.

Because of this, it is recommended that a tuning indicator be used at all times in order to make on-center tuning as rapid as possible.

It was noted in reference to Fig. 13 that the skirt selectivity of the i.f.-limiter system was improved with the addition of feedback. This indicated that adjacent channel interference rejection, where the adjacent channel signal is considerably stronger than the desired signal, can be substantially reduced. Fig. 21 shows the results of feedback on the interference caused by a signal twice the magnitude and 200 kcs. higher in center frequency than the desired signal. Both are modulated at the standard ± 75 kcs. rate.

It would be an advantage to have a feedback control on the receiver, for the pre-limiter does not behave ideally, and consequently the level of signal reaching the oscillating state does vary according
FIG. 21

EFFECT OF FEEDBACK ON ADJACENT CHANNEL INTERFERENCE

Desired (weaker) signal - 5000 cycles
Interference - 400 cycles, 200 kcs. higher in frequency
Amplitude Ratio 1:2