The FM Demodulator with Negative Feedback

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This paper describes the design and theory of operation of the wideband FM demodulator with feedback (FMFB receiver) used at the Andover, Maine, earth station for Telstar satellite communications tests. Performance data for the FMFB receiver indicate a clear advantage over the conventional FM receiver in many cases. The principal advantage lies in the ability of the FMFB receiver to raise the threshold at which "breaking" occurs for TV and other wideband signals.

AUTHOR

I. INTRODUCTION

The FM demodulator with negative feedback (FMFB)* is an outgrowth of an FM receiver circuit first described in 1937.[1] A demodulator of this type was used during the communication tests with the Echo I satellite, which started in August, 1960.[2] Compared with a conventional or standard FM demodulator, the principal advantage to be derived from this circuit is its ability to improve the threshold at which "breaking," resulting from excessive noise, will occur.

The receivers mentioned above were capable of demodulating a single 4-kc voice channel. Since it was required that the Telstar system be capable of handling television and other wideband signals, the design of feedback receivers became a problem of much greater complexity as a consequence of the relatively wide baseband.

The FMFB receiver to be described in this paper is used in the Andover ground station for the reception of wideband signals from the Telstar satellite.[3] A different FMFB receiver, described elsewhere,[4] was used in the Holmdel station during Telstar experiments. The basic elements of the feedback receiver are shown in the block diagram of Fig. 1. The incoming RF signal is combined in a mixer with the output of a

* Sometimes called "FM feedback receiver" or "frequency compression demodulator."

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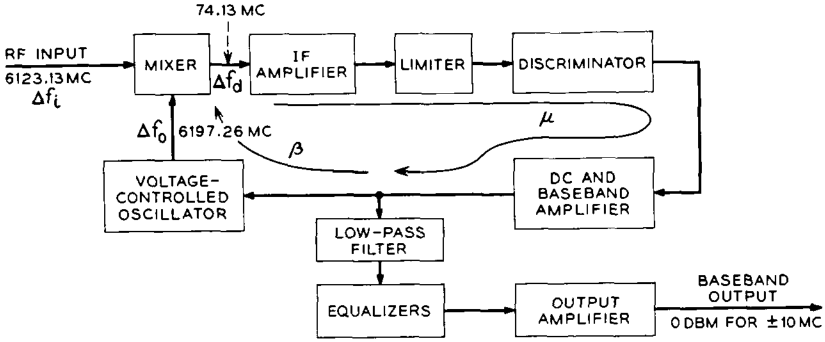

Fig. 1 — Block diagram of FMFB receiver.

voltage-controlled oscillator (VCO) to produce an intermediate-frequency signal. The frequency modulation which has been imparted to this signal, after going through an IF amplifier and a limiter, is then demodulated by the discriminator. The recovered baseband signals are amplified in the dc-and-baseband amplifier and are fed back to the VCO in such a way as to reduce significantly the instantaneous frequency difference between the two signals going into the mixer. The modulation index of the IF signal is thereby diminished. If this reduction is substantial, the resulting FM wave will occupy a much narrower band than the incoming RF signal. This makes it possible to restrict the bandwidth of the IF amplifier. In a sense, the feedback receiver can be considered as a "tracking" filter whose bandwidth is substantially narrower than that required to transmit the complete incoming FM wave; being narrower, the receiver is more immune to incoming noise.

The baseband signal to be fed to the output terminal of the receiver is taken from the input to the VCO and then passed through a low-pass filter, amplitude equalizers, and an output amplifier. The amplitude equalizers correct for deviations in the closed-loop transmission characteristics which are present at low and at high frequencies.

II. THEORY OF THE FMFB RECEIVER

2.1 *Operation above the Breaking Point*

The operation of the FMFB receiver is well understood in the region where the input carrier-to-noise ratio is large, which means well above the breaking point. For analytical purposes, it is convenient to represent the RF and IF circuits in the feedback loop by the baseband analogs which are valid for FM signals. Simplified analog circuits are normally used which are accurate only for low-index FM signals. Such analogs do not describe the nonlinear distortions of the FM signal which are actually produced in the RF and IF circuits. Nevertheless, they are good enough for a linear analysis above threshold.

The FMFB circuit can then be represented as a conventional feedback circuit with the $\mu(s)$ and the $\beta(s)$ paths defined as in Fig. 1. The mixer is the element which forms the instantaneous difference between the frequencies of the incoming and the fed-back RF signals, i.e., $\Delta f_d = \Delta f_i - \Delta f_o$. The expression $\mu\beta$ is called the open-loop transmission characteristic. It is of fundamental importance in the design of a feedback circuit, since it determines the stability of the loop through its effect on the closed-loop transmission characteristic

$$\frac{\Delta f_o(s)}{\Delta f_i(s)} = \frac{\mu\beta}{1 + \mu\beta} = \frac{\mu\beta}{F}, \tag{1}$$

where $F = 1 + \mu\beta =$ feedback. In any practical wideband FMFB circuit, $\mu\beta$ is a function consisting of many poles and zeros. Analytical methods and actual measurements, together with the optimizing procedures given by Bode,[5] have been used here to design and shape the feedback loop.

The fact that so many stages (or poles) are necessary makes the design of a truly wideband FMFB circuit very difficult. The design problem is reminiscent of the one found in building amplifiers with maximum gain-bandwidth product, to which is added the complicating requirement of holding the over-all phase shift well below 180°. The problem was approached by carefully selecting each part of the FMFB circuit to have as little phase shift and as much gain as possible. In addition, it is advantageous to use high-level signals at the two inputs of the mixer of Fig. 1. This reduces the amount of amplification required inside the loop and can therefore be considered to be the same as obtaining gain without phase shift. This method is limited mainly by the power handling capabilities of the circuits.

In the circuit of Fig. 1 the transmission factor $\beta(s)$ is associated with the VCO alone. In the design to be described in Section III, the VCO consists of a reflex klystron whose transmission characteristic can be assumed to be a constant. The over-all input-output characteristic of the FMFB receiver is therefore obtained by multiplying the closed-loop transmission characteristic with the characteristics of the low-pass filter and the equalizers.

The signal-to-noise ratio (SNR) at baseband for a conventional FM receiver for carrier-to-noise ratios (CNR) at RF well above breaking is known to be

$$\mathrm{SNR} = \frac{3}{2} m^2 \frac{B}{b} \mathrm{CNR}, \tag{2}$$

where B = RF bandwidth, b = baseband width, and m = modulation index. The RF and baseband transmission characteristics are assumed to be rectangular. It has been shown in Ref. 6 that (2) remains unchanged for an equivalent FMFB receiver. The feedback receiver, therefore, has no noise advantage if operated above the breaking or threshold point. This point, characterized by a certain CNR, is reached when the FM noise improvement given by (2) can no longer be obtained. The breaking point shall be loosely defined here as the CNR where the actual SNR has dropped 1 db below the value given by (2).

2.2 *Operation near the Breaking Point*

No theory is at present available which completely describes the behavior of an FMFB receiver near the threshold of noise improvement. Enloe[6] has introduced a two-threshold concept which has given good results in practical cases. Strictly speaking, the results of Ref. 6 are valid only under the assumption of an unmodulated carrier. We shall use this concept here in order to find simple mathematical expressions for the open-loop and the closed-loop threshold. Relations will then be derived for the optimum amount of feedback and the upper limit of the threshold improvement. These results are then compared with actual measurements taken under various conditions of modulation.

2.2.1 *The Open-Loop Threshold*

Under the assumptions that the incoming carrier is unmodulated and that the VCO generates a steady carrier (open loop), the CNR after the narrow-band IF amplifier (of bandwidth B_{IF}) in the loop will be greater than that at RF (bandwidth $B > B_{\mathrm{IF}}$). Since FM discriminators break at about the same CNR, it can be seen that the insertion of the narrow-band filter results in a threshold improvement over a standard, wide-band FM receiver. The breaking point of a conventional FM receiver varies slightly with the ratio B_{IFN}/b (or B/b). B_{IFN} is the noise bandwidth of the IF filter, and we assume it to be 20 per cent higher than the 3-db bandwidth B_{IF}. B is assumed to be the 3-db bandwidth as well as the noise bandwidth. The empirical formula

$$\mathrm{CNR}_0 = \left[250\left(\frac{B_{\mathrm{IFN}}}{2b} - 1\right)\right]^{\frac{1}{2}} \tag{3}$$

was found to approximate well the breaking points shown in Fig. 9 of Ref. 6. The expression gives good results for B_{IFN}/b as low as 2.5. The threshold, as measured in the RF band B, then appears at

$$\mathrm{CNR}_1 = \mathrm{CNR}_0 \frac{B_{\mathrm{IFN}}}{B} = \frac{B_{\mathrm{IFN}}}{B}\left[\left(250\left(\frac{B_{\mathrm{IFN}}}{2b} - 1\right)\right]^{\frac{1}{2}}. \tag{4}$$

Despite the fact that we consider the threshold under conditions of no modulation, we shall introduce here the modulation index $m = \Delta f/b$, where Δf is the peak frequency deviation caused by full sine-wave modulation at the top baseband frequency b. We can now express B in new terms, employing the so-called Carson's rule

$$B = 2(1 + m)\, b. \tag{5}$$

Under the assumption of uniform feedback F over the band b, the 3-db bandwidth of the IF filter in the loop should be $B_{\mathrm{IF}} = 2[1 + (m/F)]\, b$. Using these relations and $B_{\mathrm{IFN}} = 1.2\, B_{\mathrm{IF}}$, we can rewrite (4) in terms of m and F only

$$\mathrm{CNR}_1 = 1.2 \frac{1 + (m/F)}{1 + m}\left\{50\,[1 + (6m/F)]\right\}^{\frac{1}{2}}. \tag{6}$$

2.2.2 *The Closed-Loop Threshold*

The CNR at which the closed-loop threshold occurs is directly proportional to B_c, which is equal to twice the closed-loop noise bandwidth at baseband. If we again measure the second threshold in the RF bandwidth B, we find by using equation (1) of Enloe's paper

$$\mathrm{CNR}_2 = 4.8 \frac{B_c}{B}\left(\frac{F - 1}{F}\right)^2. \tag{7}$$

In order to obtain values for B_c, the open-loop characteristic $\mu(s) \cdot \beta(s)$ has to be specified. The Bode-type open-loop characteristic,[5] which provides constant feedback F up to a baseband frequency b and from there on a constant phase, can be considered to be near optimum for our purposes. It will be used for the following derivation. Although not strictly realizable in practice, this characteristic can be approximated by a technique, also due to Bode, which employs the so-called step and fillet. Fig. 8 of Ref. 6 has curves which give B_c as a function of the phase margin with F as parameter. A phase margin of 50° is considered to be near optimum, because it minimizes B_c over a wide range of F. The empirical relation $B_c = 2bF^{0.8}$ was found to agree well with these curves for a phase margin of 50°. With this relation and (5), we can write (7) in the following form:

$$\mathrm{CNR}_2 = 4.8 \frac{F^{0.8}}{1+m}\left(\frac{F-1}{F}\right)^2. \tag{8}$$

2.2.3 *Optimum Feedback and Upper Limit for Threshold Improvement*

It can be seen from (8) that an increase in feedback F will increase the closed-loop threshold, mainly due to the increase in B_c. Equation (6), on the other hand, indicates a decrease in threshold with increasing feedback. For a certain value of F, the two thresholds will therefore be equal. Such an operating point will tend to optimize the over-all threshold of the FMFB receiver. Equations (6) and (8) were equated and the result is plotted in Fig. 2 as F_{OPT} in decibels versus the modulation index $m = (B/2b)-1$. For a practical range of m from 3 to 10, we find feedback values varying over the narrow range from 14.2 to 18.2 db.

We can now determine the improvement I which an FMFB receiver provides over the threshold of a standard FM receiver. The standard

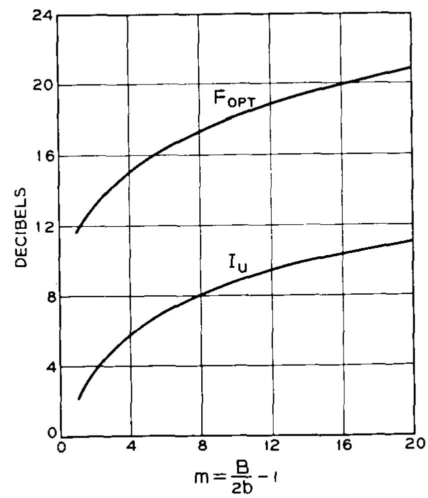

Fig. 2 — Optimum feedback F_{OPT} and upper limit of threshold improvement I_u.

receiver has a threshold given by (3) if B_{IFN} is replaced by B. Using (5), the threshold becomes

$$\mathrm{CNR}_3 = \sqrt[3]{250\, m}. \tag{9}$$

Inserting $F_{OPT}(m)$ into either (6) or (8) yields CNR_{OPT}. The improvement is therefore

$$I \leqq \frac{\mathrm{CNR}_3}{\mathrm{CNR}_{OPT}}. \tag{10}$$

The equals sign indicates an upper limit I_u for the improvement, which could be reached only if the two thresholds occurred abruptly and independently of each other. This will most likely not be the case and the improvement will always be below this upper limit. I_u was calculated using the values of F_{OPT} previously found. The result is plotted as a second curve in Fig. 2. In the range of $m = 3$ to 10, upper bounds for the improvement reach from 4.9 to 8.8 db.

2.2.4 *Comparison with Measured Results*

We shall now compare the formulas derived above with measurements made on the FMFB receiver which are described in Section IV. The pertinent measured quantities are: $B = 25$ mc, $b = 2$ mc, $B_{IFN} = 7$ mc, $F = 15$ db (at 0.5 mc) and $B_c = 15$ mc. Then we find from (4) and (7), $\mathrm{CNR}_1 = 2.05$ db and $\mathrm{CNR}_2 = 2.90$ db, respectively. Therefore, the closed-loop threshold is governing by the small margin of 0.85 db. This is not quite optimum, and the design could be improved by a slight widening of the IF filter in the loop. Since the over-all threshold of the feedback demodulator is above either of the two thresholds, it must be greater than 2.9 db. It actually occurs at 4.5 db, as indicated by point A on curve 1 of Fig. 12 (Section 4.3.2). It can also be seen from this figure that the threshold is adversely affected by the application of modulation, especially at the higher baseband frequencies. We also wish to compare the measured results with the optimum design. Using (5), we obtain $m = 5.25$, and from Fig. 2 we find an optimum value for the feedback of $F_{OPT} = 16$ db and an upper limit for the threshold improvement of $I_u = 6.65$ db. The optimum setting of F was experimentally found to be the aforementioned 15 db at 0.5 mc. From Fig. 12, the threshold improvement amounts to 4.5 db, equal to the difference between points A and B. It can therefore be concluded that the actual design of the FMFB receiver has come to within 1 db of the calculated F_{OPT} and within 2.15 db of I_u.

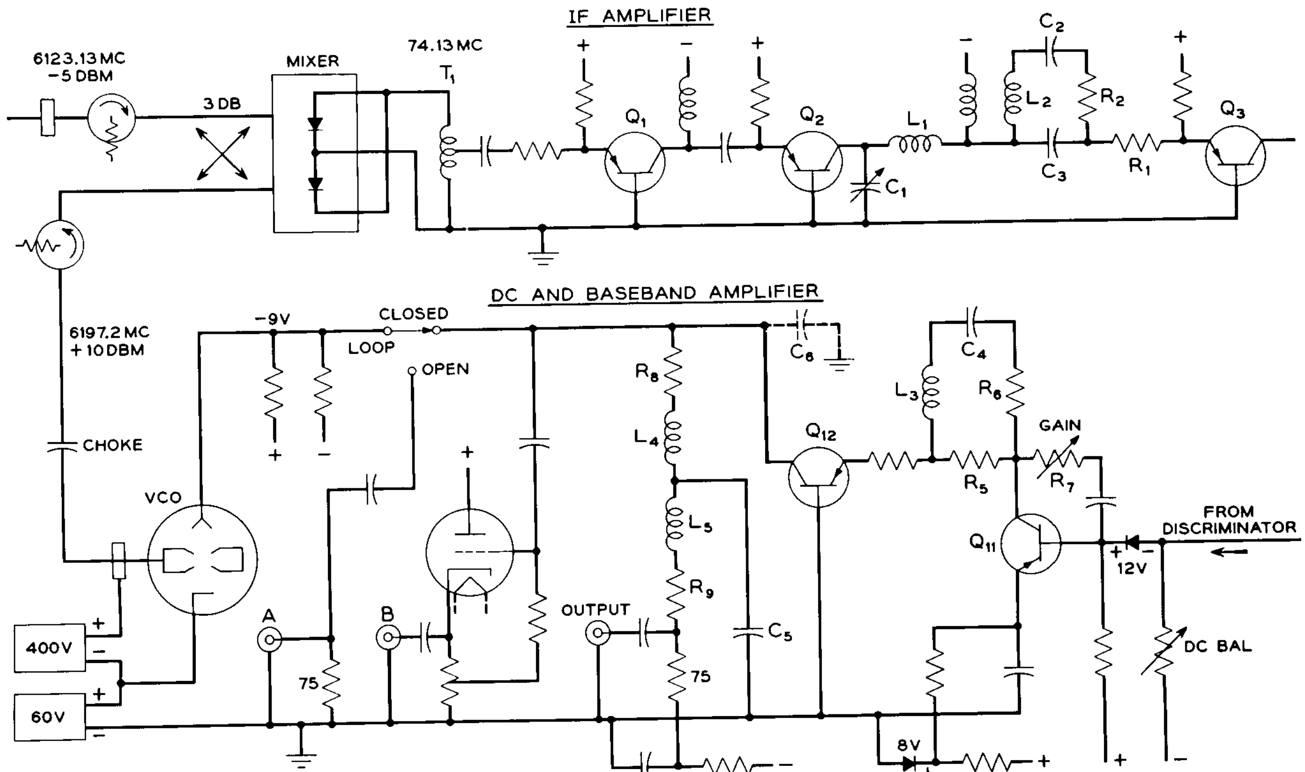

Fig. 3 — Circuit diagram of mixer, IF amplifier, VCO and dc-and-baseband amplifier.

III. CIRCUIT DESCRIPTION

Simplified schematic diagrams of the feedback receiver are shown in Figs. 3 and 4. For simplicity, conventional filtering and power supply circuit details have been omitted.

The decision to use a microwave frequency as the input to the feedback circuit was dictated by the choice of a 6-gc reflex klystron for the VCO (Fig. 3). A klystron gives very little signal delay, has good modulation sensitivity, and its linearity is excellent. The klystron used is a modified version of the WE450A tube, in which the modulation sensitivity was increased from 1.5 to 5.0 mc/volt by altering the operation from the 2-3/4 to the 3-3/4 mode, decreasing the loaded Q slightly and changing the repeller geometry.*

The repeller capacitance of only 3 pf tends to keep the delay or phase shift in the driving circuit at a low value. A second, although very small, source of delay is found in the klystron itself. It is estimated that the modulating signal is delayed by the total travel time of the electrons in the repeller field. Since the tube is working in the 3-3/4 mode, this time is about 0.6 ns at the operating frequency of 6197 mc. No delay is produced by the cavity resonator, however. The nonlinearity of the klystron is very low and amounts to a change in modulation sensitivity of less than 4 per cent at a frequency 10 mc away from midband. Since the feedback in the receiver tends to reduce the nonlinear effects of the other circuits, the klystron can be considered the main source of nonlinear distortion. Since this klystron is sensitive to a small degree of mismatch, an isolator is provided between its output and the input to the mixer. The isolator has a signal delay only slightly greater than a straight piece of waveguide of the same length.

In order to maintain the driving transistor circuits at ground potential, the negative terminal of the repeller bias source is grounded. Consequently the cavity of the klystron is at high potential, necessitating the use of an insulating waveguide choke.

The mixer is a version of the receiving modulator used in the TH radio system, modified to operate efficiently at higher local oscillator and input levels. The mixer is designed to deliver the 74-mc IF signal into a 75-ohm unbalanced load. Since the emitter of the grounded-base transistor Q_1 has a much lower impedance, a broadband autotransformer (T_1) having a 2:1 turns ratio[7] is inserted, which partly corrects this mismatch. These transformers have a bandwidth of about 150 mc.

* This modification was carried out at the Allentown Laboratories of Bell Telephone Laboratories.

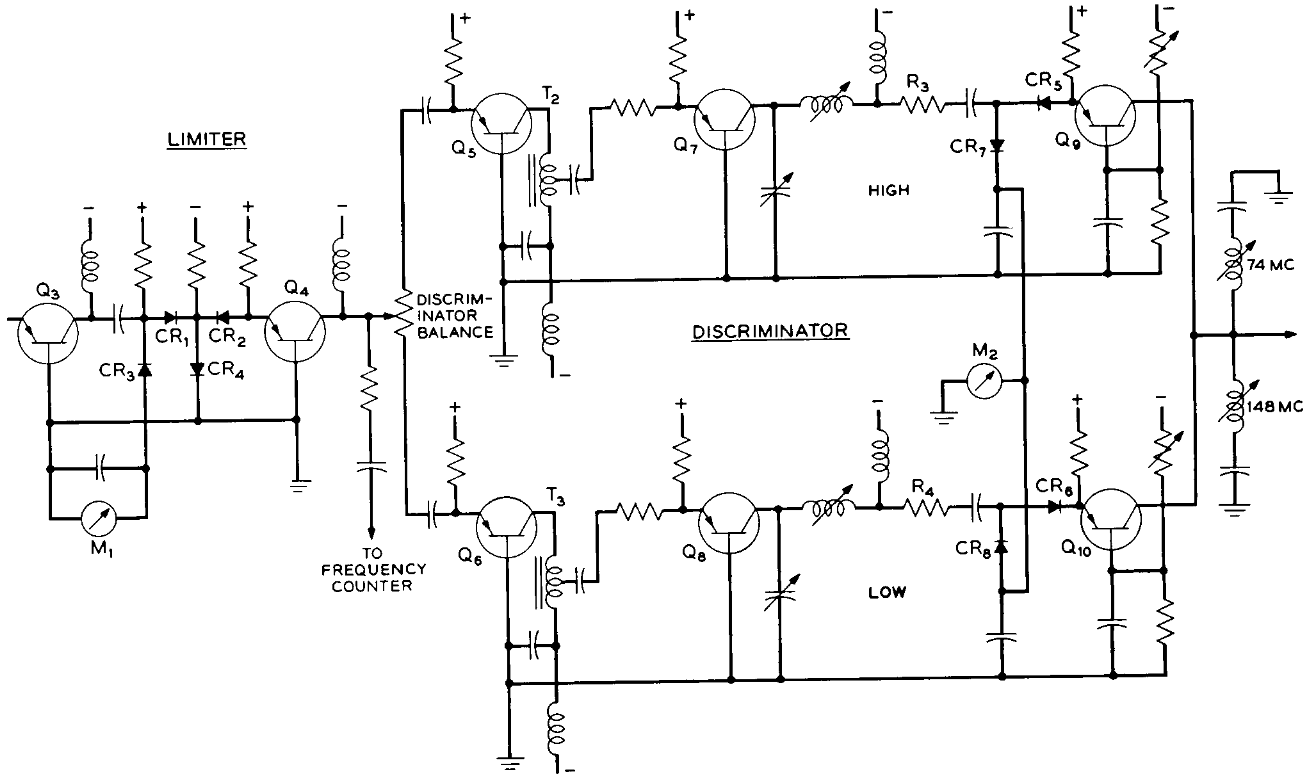

Fig. 4 — Circuit diagram of limiter and discriminator.

Diffused-base germanium transistors with alpha cutoffs above 500 mc were used throughout the circuit.

It was found that the relatively high impedance of the narrow-band IF filter network in the collector circuit of transistor Q_2 produced marked variations in the input impedance of this stage in the vicinity of resonance, even with a transistor having an $r_b'C_c$ product as low as 10 ns. In order to provide a suitable termination for the secondary of autotransformer T_1, it was necessary to introduce a buffer stage, represented by the first transistor, Q_1, to mask this effect.

The network shown in the output of the second transistor (Q_2) provides a bandpass transmission characteristic which limits the noise power delivered to the limiter-discriminator.

In the choice of a filter suitable for use at this point, several factors must be taken into consideration. From the standpoint of noise rejection, the use of a highly selective bandpass structure would be advantageous. However, such a filter would introduce a prohibitive amount of delay into the feedback loop. A simple single-pole network, on the other hand, has a much more favorable phase characteristic, but its effective noise bandwidth is 1.57 times the bandwidth between half-power points. The circuit shown is a compromise between these extremes. The tuned circuit formed by elements C_1, L_1, and R_1 alone would constitute a one-mesh filter circuit adjusted to resonate at the center of the IF band. The addition of a similar tuned circuit formed by elements C_2, L_2, R_2, loosely coupled to the first circuit through the mutual impedance of capacitor C_3, leads to a more nearly flat-topped transmission characteristic, shown in Fig. 5 by solid lines. For comparison, the characteristics of the one-mesh circuit of the same 3-db bandwidth are given by the dashed curves of Fig. 5. Although the two gain curves are superimposed in the figure, the two-mesh circuit produces a current gain which is 3 db greater than that of the one-mesh circuit. Note that the asymptotic phase shift with respect to midband is 90° for both circuits. The noise bandwidth of the two-mesh circuit is 7 mc, which is only 25 per cent higher than its 3-db bandwidth. The current gain is about 24 db. All the current gain in the IF amplifier is achieved by means of stable passive circuitry, with the common-base transistors acting as impedance transformers.

The common-base stage Q_3 feeds the IF signal to the series-type diode limiter which is shown in Fig. 4. Diodes CR_1 and CR_2 are normally conducting except at the peaks of the input wave. Diodes CR_3 and CR_4 normally do not conduct, except when diodes CR_1 and CR_2, respectively, are nonconducting.

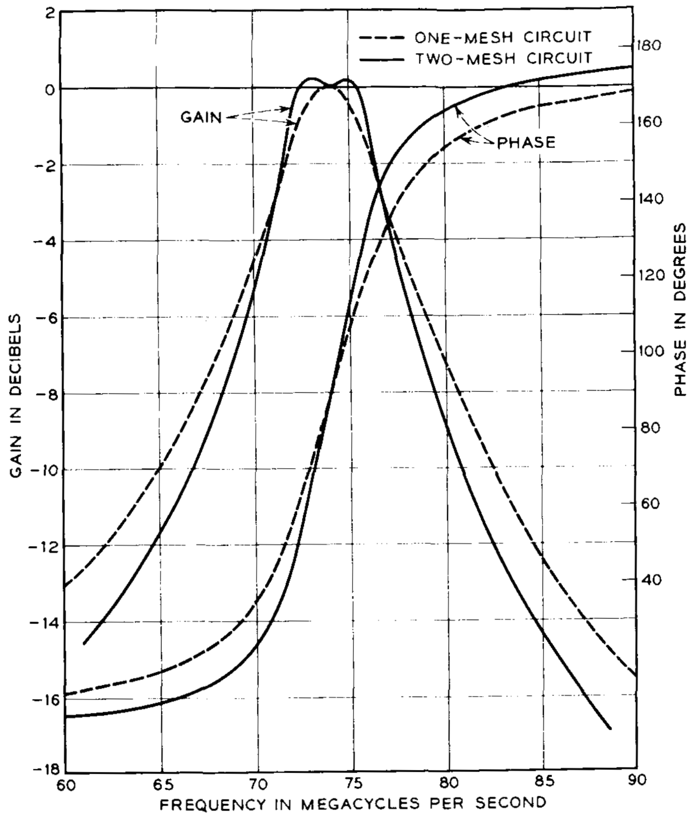

Fig. 5 — Gain and phase characteristics of narrow-band IF filter.

The limiter has no reactive elements and its impedance level is extremely low. This makes it a wideband circuit which exhibits very small signal delay. Envelope variations appearing on the input wave are reduced about 22 db by the circuit.

A dc meter M_1 in series with diode CR_3 serves as a convenient means for adjusting and monitoring the input to the limiter. The level is generally adjusted (by changing the 6123-mc input signal) so that clipping takes place about 3 db below the peaks of the IF signal.

Current delivered by the limiter is divided into two equal parts which are then subjected to 6 db of current gain in transformers T_2 and T_3 before being applied to the two inputs of a balanced discriminator. This discriminator consists of two simple resonant circuits connected to the collectors of transistors Q_7 and Q_8. One circuit is tuned to a frequency

12 mc above and the other a similar amount below the center frequency of the IF band. Damping of each circuit is adjusted by resistors R_3 and R_4 for a bandwidth of about 24 mc, so that the current gain of each branch is equal to 7 db at 74 mc.

When the incoming signal is frequency-modulated, frequency deviations give rise to amplitude variations within the two tuned circuits. These are detected by diodes CR_5 and CR_6. Diodes CR_7 and CR_8 provide a dc return in each case. These dc currents return to ground through a differential meter M_2, which gives a zero indication at midband when the two currents are equal. In order to avoid unequal biasing of diodes CR_5 and CR_6 by the emitter voltages of transistors Q_9 and Q_{10}, a bias is applied to the bases of these two transistors to make the resulting emitter voltages equal to zero.

The output of the discriminator is taken from the collectors of transistors Q_9 and Q_{10} and delivered to the dc-and-baseband amplifier shown in Fig. 3. Amplitude variations occasioned by noise will tend to cancel in the output line, thus supplementing the smoothing action of the amplitude limiter. Series-resonant circuits in the output line shunt to ground the 74- and 148-mc components which remain after the detection process.

The dc-and-baseband amplifier provides gain for the demodulated signal and delivers it to the repeller of the klystron as well as to the outgoing video line. Automatic center frequency control is accomplished by providing dc coupling within the amplifier.

In order to achieve adequate gain and phase margins within the feedback loop, it is necessary to control the open-loop gain characteristic over a wide range of frequencies — a problem common to all feedback amplifiers. This shaping of the loop has in general followed the techniques proposed by Bode.[5] The main correcting network is located in the collector circuit of transistor Q_{12} (Fig. 3). The part of the circuit consisting of C_6, R_8, and L_4, with return to ground over the large capacitor C_5, has a transmission peak at 30 mc which ensures the presence of the correct "Bode step" in the open-loop transmission characteristic. Below 30 mc, the gain of the last stage decreases and eventually rises again to a constant value at about 1 mc because of the presence of another shunt peaking network L_5, R_9, and C_5.

The common-emitter stage Q_{11} provides some minor shaping by means of the local feedback circuit consisting of R_7 together with R_5, R_6, C_4, and L_3. Increased gain is obtained for the "Bode fillet" at 1.25 mc, which is the resonant frequency of the series circuit R_6, C_4, L_3. The gain of Q_{11} otherwise is very flat up to about 40 mc, and can be

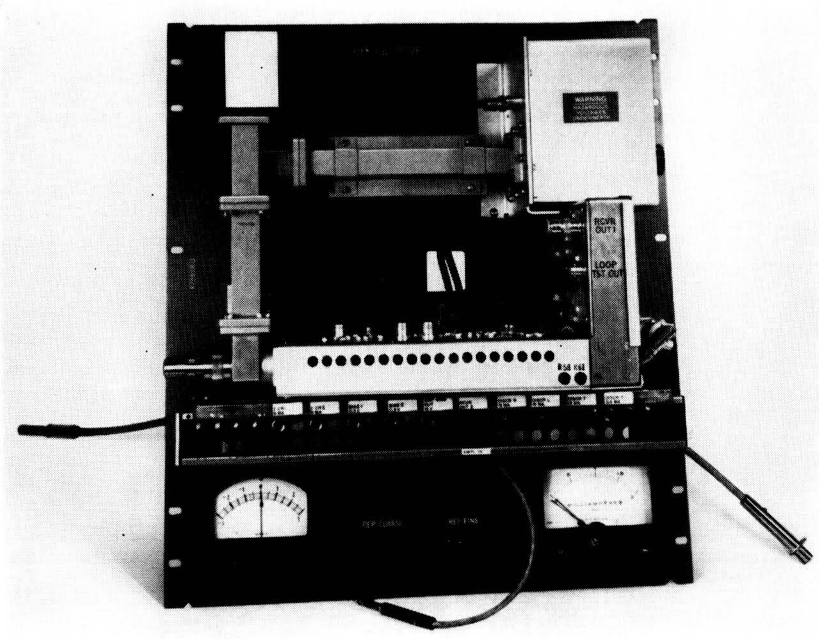

Fig. 6 — Photograph of FMFB receiver.

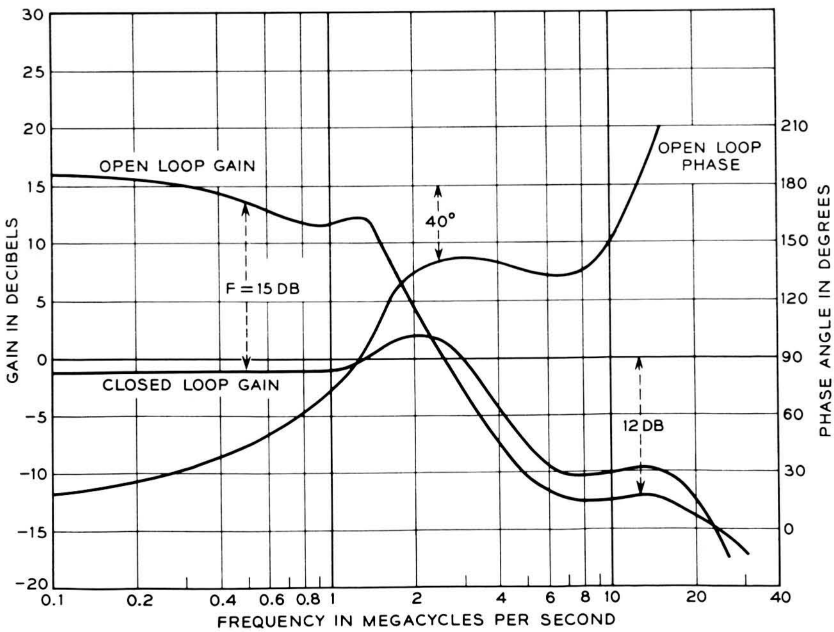

Fig. 7 — Open- and closed-loop transmission characteristics.

varied by R_7 over the range of 2 to 12 db. The normal setting is around 6 db, which results in a feedback of 15 db at 0.5 mc.

The gain characteristic of the loop is also well controlled down to dc by various circuits — not separately described — in the dc-and-baseband amplifier.

Baseband signals to be delivered to the outgoing line are taken from a 75-ohm impedance point within the main equalizing network. Transmission to this point differs markedly from the actual loop characteristic at frequencies above the usual baseband, in that frequencies higher than 3 mc are greatly attenuated. This effect is symbolized by the low-pass filter shown in Fig. 1.

The actual open-loop performance is measured by opening the loop with the switch in Fig. 3, applying baseband signals to the repeller at point A, and observing the voltage which appears at the collector of Q_{12}. Since this is a relatively high-impedance point, a cathode follower is permanently connected to the collector. The cathode impedance of this device is sufficiently low that standard low-impedance measuring devices can be connected at point B. The closed-loop characteristic is likewise measured at point B, with the loop switch in the closed position and a frequency-modulated 6123-mc signal applied to the input of the mixer.

A photograph of the FMFB receiver is shown in Fig. 6. The box housing the klystron (VCO) is shown on the upper right. The mixer on the lower left is attached to the horizontal part of the chassis, which contains the IF amplifier, limiter and discriminator. The dc-and-baseband amplifier is located in the vertical part of the chassis. The equalizers and the output amplifier are not shown in the photo.

IV. MEASUREMENTS

4.1 *Transmission Characteristics*

Gain and phase characteristics of the open loop are shown in Fig. 7. With the feedback F set to 15 db at 0.5 mc, the phase and gain margins are seen to be 40° and 12 db, respectively. Not shown in the figure is the transmission characteristic below 100 kc. Towards the very low frequencies the loop gain is permitted to increase slowly, so as to obtain about 5 db of additional gain at 15.75 kc. This is advantageous from the standpoint of television signals.

The gain with the loop closed is also shown in Fig. 7. In this case, a reflex klystron was used as the source of the 6123-mc FM signal. The closed-loop characteristic exhibits the typical region of gain enhancement in the vicinity of the open-loop gain crossover. This region is strongly affected by the gain in the feedback loop. More gain would increase the transmission bulge around 2 mc, and in turn increase the closed-loop noise bandwidth. Less gain would cause insufficient frequency compression in the loop. With a feedback of 15 db, the closed-loop noise bandwidth at baseband was found to be about 7.5 mc, or $B_c = 15$ mc.

Since the effect of the gain enhancement around 2 mc is also seen at the useful output terminal of Fig. 3, a compensating high-frequency equalizer is included ahead of the output amplifier, as indicated in Fig. 1. The increase of gain in the open loop below 50 kc affects the closed-loop characteristic by about 1.3 db. A low-frequency equalizer makes the transmission flat to about 0.2 db down to a few hundred cycles. Transmission to the output of the output amplifier is shown in Fig. 8.

4.2 *Excess Phase Problems*

The problem of excess phase will now be dealt with in some detail because of its importance to present and future wideband feedback receivers. The term "excess phase" is used to indicate phase shift in excess of the theoretically determined minimum phase associated with the open-loop transmission characteristic. An obvious contribution to excess phase arises from the physical dimensions of the loop structure itself. In the present instance it is estimated that this can account for a delay of about 3.5 ns. In terms of phase, this would introduce a linear phase shift

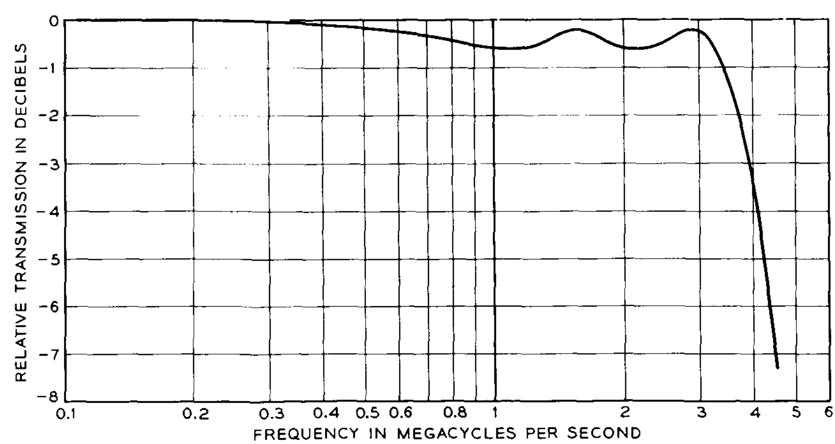

Fig. 8 — Over-all transmission characteristic.

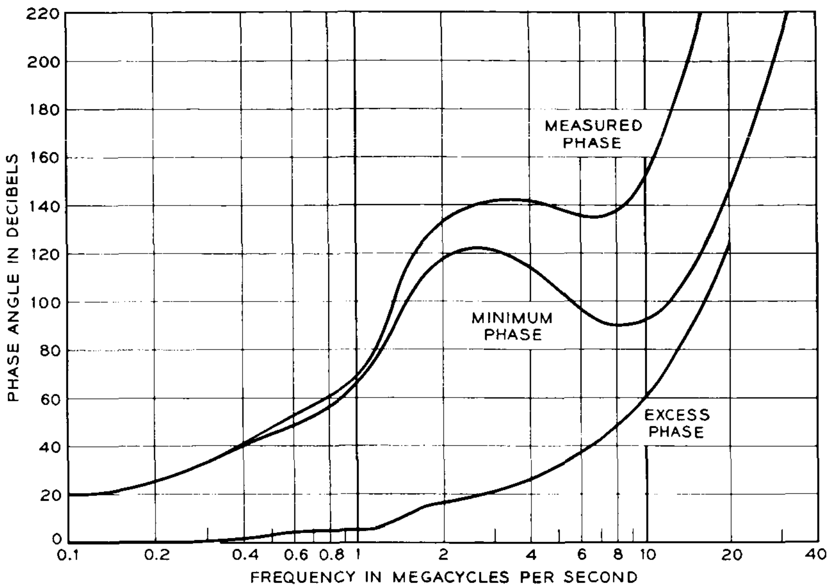

Fig. 9 — Measured and computed minimum phase characteristic of the open loop.

amounting to 1.27 degrees per megacycle. Though not negligible, the restriction thus imposed upon loop performance is rather small. Actually, a much greater amount of excess phase has been observed. This is illustrated in Fig. 9, which shows the computed minimum phase characteristic based upon the measured open-loop transmission. This curve is an approximation because of some degree of uncertainty regarding the cutoff performance at very high frequencies, in the region between 30 and 100 mc. However, conditions at these rather remote frequencies have only a minor influence upon the region of greatest interest.

Shown for comparison is the measured phase characteristic. The difference between these two curves closely follows the linear law of about 6 degrees per megacycle. This is five times as great as that estimated from the physical dimensions of the feedback loop. Several additional sources have been identified. Transistors are known to introduce a measure of excess phase. Measurements made at intermediate frequencies with a single common-base stage, such as is used at many points within the receiver, show excess phase amounting to about 0.2° per megacycle. Since there are in effect eight such stages within the loop, the total contribution of 1.6° per megacycle is not negligible.

TABLE I — SOURCES OF EXCESS PHASE

Source	Estimated Phase
Loop dimensions	1.3°/mc
8 common-base stages	1.6°/mc
Limiter	0.5°/mc
Common-emitter stage	2.0°/mc
Total	5.4°/mc
Measured	6.0°/mc

Considerable excess phase is associated with the common-emitter stage Q_{11} of the baseband amplifier. This amounts to about 2° per megacycle.

The limiter was found to have an essentially linear phase characteristic which, allowing for the input and output transistors, indicated a phase contribution of 0.5° per megacycle. Negligible phase shift is caused by the estimated 0.6 ns of delay in the klystron.

Adding these several contributions, we can account reasonably well for the difference between the minimum phase and the observed open-loop performance as indicated in Table I.

4.3 *Measurements near Threshold*

In order to obtain a better insight into the rather complicated mechanisms responsible for the performance near threshold, several types of measurements were performed on the FMFB receiver and on a conventional or standard FM receiver. The latter was taken from the TH radio system.[8] In order to produce the desired CNR's near threshold, flat thermal noise of 25-mc bandwidth was mixed with the FM signal at the intermediate frequency of 74 mc. This signal was then directly applied to the standard FM receiver, but translated to 6123 mc in a modulator for application to the FMFB receiver.

4.3.1 *Baseband Noise Spectra*

Baseband noise spectra were measured for CNR's ranging from 0 to 22 db. For this purpose, a sensitive analyzer with a 4-kc bandwidth was connected to the output of the receivers and readings were taken over the range from 100 kc to 5 mc. The solid curves of Figs. 10 and 11 show samples of these measurements taken on the standard and the FMFB receivers, respectively, for the case where the carrier is unmodulated. The noise readings were corrected where necessary so as to simu-

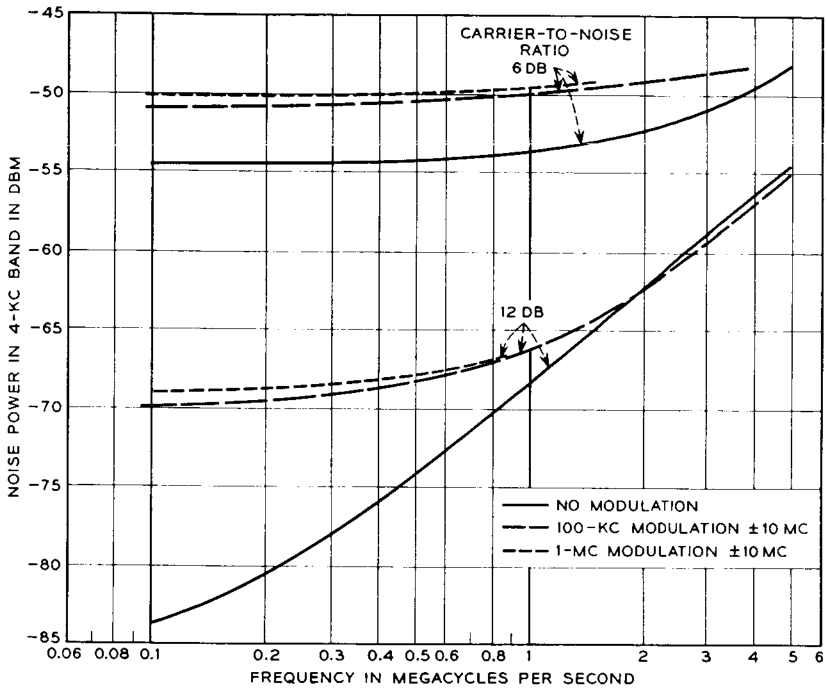

Fig. 10 — Noise spectra for standard FM receiver.

late absolutely flat receiver gain characteristics from 100 kc to 5 mc. At the higher CNR's, and for frequencies above a few hundred kilocycles, the spectra show the typical triangular characteristic of FM noise. When the CNR is decreased, the noise rises much faster at low frequencies, to produce ultimately the fairly flat noise spectra which were calculated by S. O. Rice[9] for the standard FM receiver. It has been known for some time that the baseband noise increases to some extent when modulation is applied to the FM carrier,[10] but the literature on the subject is very sparse. It was therefore decided to measure noise spectra in the presence of full ±10-mc sine-wave modulation. Such measurements are of somewhat limited practical value, because a single tone will only rarely be transmitted with the maximum frequency deviation. This is the case, however, in a number of TV test patterns, especially the "multiburst" pattern. Occasionally, such a situation may also arise in live TV signals for short periods of time. Examples of the results of modulation measurements are also shown in Figs. 10 and 11. A drastic difference exists between the standard and the FMFB re-

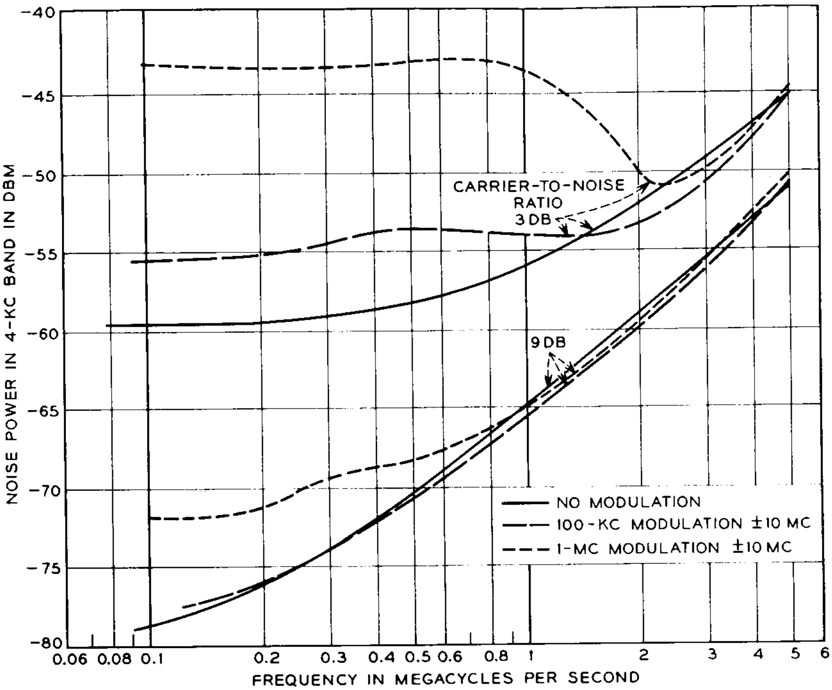

Fig. 11 — Noise spectra for FMFB receiver.

ceiver. The former shows an increase in noise which is nearly independent of the modulating frequency. The FMFB receiver, on the other hand, is little affected by a 100-kc modulating signal but much more so by a 1-mc signal, especially at low CNR's. The strongest effect of the modulation is seen at the low-frequency end of the noise spectrum, particularly in the case of the standard FM receiver at high CNR's.

4.3.2 *Threshold Curves*

The behavior of an FM system above and around the threshold region is often described by plotting the ratio of the signal power to the total noise power (SNR) in a baseband b, as a function of the CNR at RF in the bandwidth B. The total noise power in a 2-mc baseband was first determined for the case of no modulation by integrating a set of curves, of which the solid ones in Figs. 10 and 11 are a part. The result is shown in Fig. 12 as curves 1 and 3.

When modulation is applied, two effects are to be noted. One is the

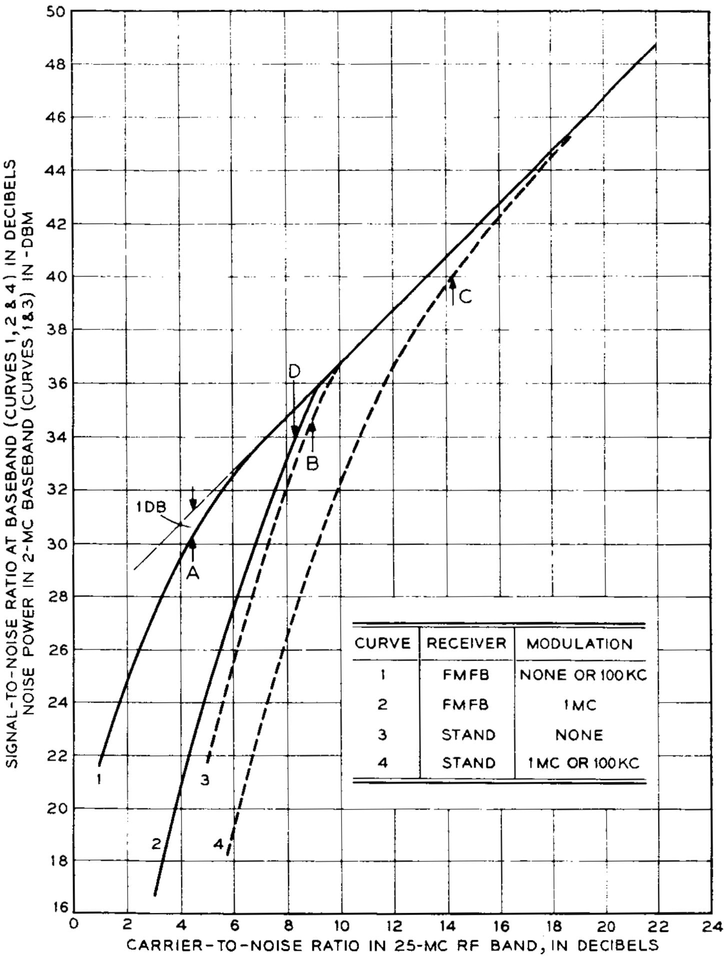

Fig. 12 — Signal-to-noise ratio at baseband.

increase in noise mentioned above. The other is a reduction in recovered signal level when the CNR is rather low. These effects shall be called noise increment and signal reduction, respectively. Noise increments for various CNR's were obtained by integrating spectra such as those shown in Figs. 10 and 11 for the case of full modulation. The results are shown in Fig. 13 for the two types of receivers. The increment is actually negative under some conditions — meaning that the total baseband noise is slightly less than for the case of the unmodulated carrier. This can be seen from the modulation curves of Fig. 11.

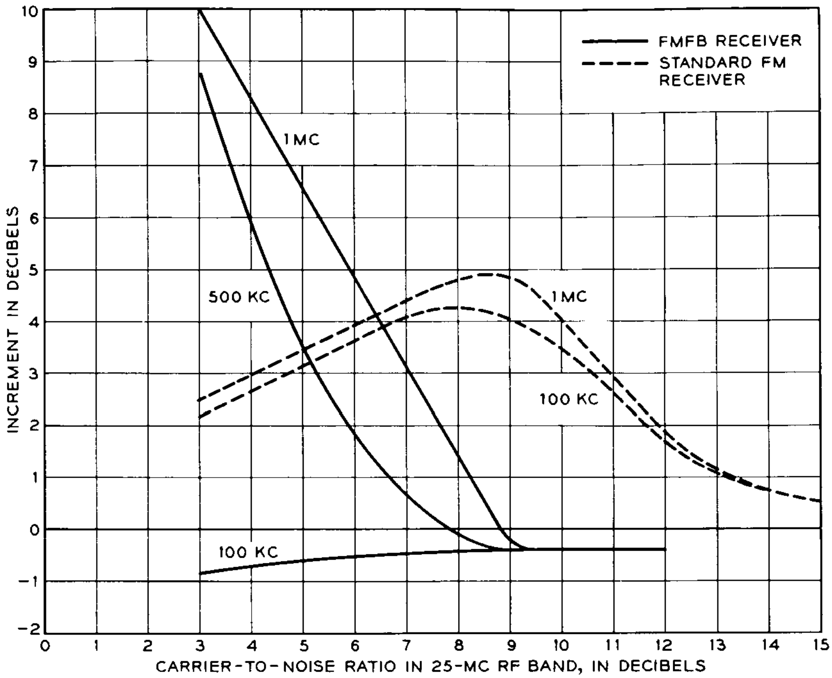

Fig. 13 — Noise increments.

Signal suppression was determined with the aid of a selective analyzer tuned to the signal frequency as the CNR was varied. Fig. 14 shows the performance which was observed.

The actual SNR in the 2-mc baseband can now be determined. The

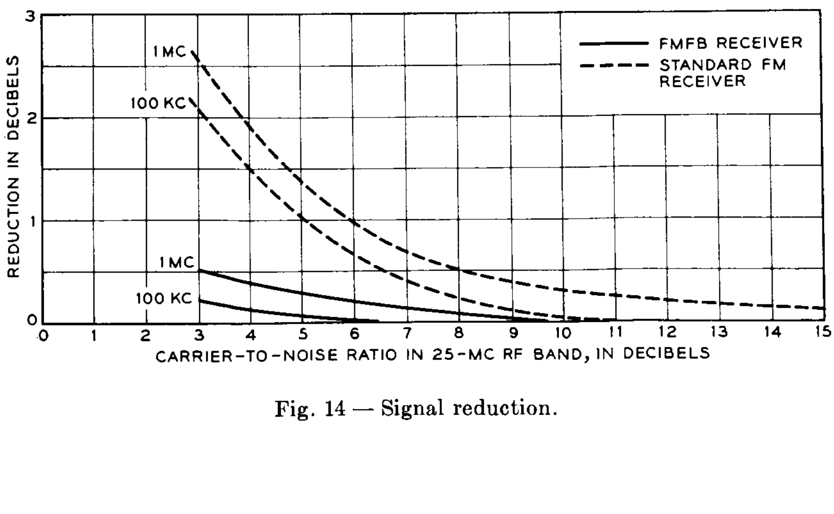

Fig. 14 — Signal reduction.

results are given in Fig. 12 as curves 1, 2, and 4. A few simplifications were made in this figure. Curves 1 and 4 are accurate only for no modulation and 1-mc modulation, respectively, and the 100-kc curves would actually be a few tenths of a db higher. Slight differences of about 0.5 db between the actual curves 1 and 3 and the calculated curve at high CNR's were neglected and made equal to the theoretical value given by equation (2).

The curves of Fig. 12 show clearly the advantage of the FMFB receiver over the standard receiver in the threshold region. An advantage of 4.5 db is obtained for the case of no modulation by taking the difference between the threshold points A and B. For 100-kc modulation the threshold improvement is 9.8 db ($C - A$), and for 1 mc it is 5.9 db ($C - D$). Curves similar to 3 and 4 were calculated by Rice.[10] No theory is yet available which would produce curves 1 or 2 for the FMFB receiver.

The threshold situation for actual signals such as TV and carrier multiplex requires further exploration by subjective tests and noise loading measurements, respectively. Such tests are described in the following two sections.

4.3.3 *Television Threshold Tests*

Fig. 15 shows the results of TV threshold tests. The peak-to-peak deviation produced by the television signal (including the sync pulse) was set to be 14 mc. An aural subcarrier at 4.5 mc, which deviated the main carrier by 2.8 mc peak-to-peak, was also transmitted. The television spectrum was restricted on the transmitting side by the insertion of a well-equalized 3-mc roll-off filter. Fig. 15 shows that the television patterns are still recognizable down to very low CNR's, especially when the FMFB receiver was used. A large number of subjective tests have shown threshold improvements of about 4 to 5 db. These pictures, as well as all the data described above, were taken with the FMFB receiver adjusted for a feedback of 15 db at 500 kc and its discriminator balance set for equal numbers of white and black noise dots. The 15-db figure was found to be a good compromise, giving a small enough closed-loop bandwidth (and therefore low noise) and still enough feedback to make high-frequency breaking due to lack of feedback not too conspicuous. A live TV picture occasionally contains very strong high-frequency components which show up as sharp vertical lines. Since there is not enough feedback available to keep the frequency deviation within the passband of the narrow IF filter, the signal at the discriminator will be below the breaking point at the deviation peaks. Breaking

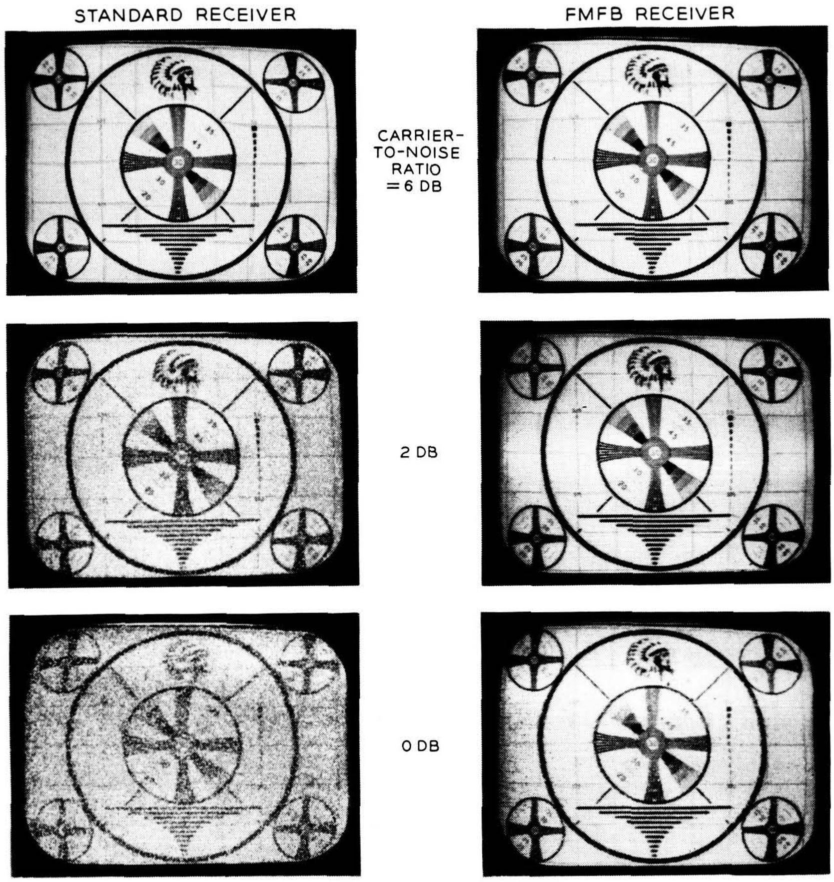

Fig. 15 — TV threshold tests.

which starts at a white peak produces a trailing black dot, and correspondingly a black peak generates a white noise dot. The nonrandomness of this breaking process makes it appear as a characteristic "sparkling" on the screen. It should be mentioned that the sparkling effect is not responsible for the strong increase in noise shown in Fig. 13 for 500 kc and 1 mc. Randomness of the noise in this case is still maintained, as observed on the oscilloscope.

The foregoing description makes it clear that TV pre-emphasis should not be used in a system containing the FMFB receiver of the present design. It is even desirable to insert a baseband roll-off filter in the transmitter in order to restrict the high-frequency components of the picture.

Tests with TV pre-emphasis were made, and it was found that the sparkling effect was strongly enhanced. This is due to the fact that the relatively short bursts of noise which cause the sparkle are integrated or lengthened in the de-emphasis network.

The FMFB receiver has also been tested with color TV signals. The behavior near threshold was found to be excellent if program material was transmitted, but rather poor if the color-bar pattern was used.

4.3.4 *Noise Loading Tests*

Tests were made to describe the behavior of the two types of receivers in the threshold region under conditions of a simulated telephone load. The telephone multiplex signal is simulated for this purpose by a noise signal with a flat spectrum, which in turn frequency modulates the RF carrier. Tests were made with 120, 240, and 600 channels. The noise-power ratio was measured at the receiver outputs in a number of frequency slots as a function of the total rms frequency deviation. The

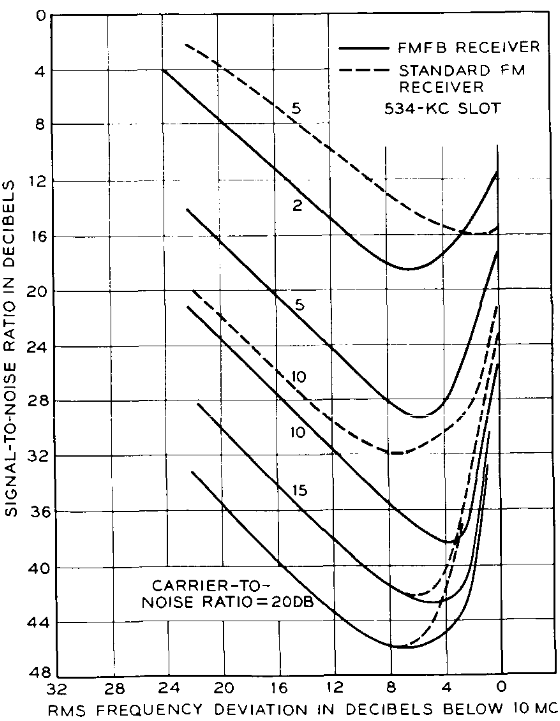

Fig. 16 — 240-channel noise loading.

noise-power ratio is the ratio of signal-plus-noise to noise, with the noise consisting of thermal and intermodulation noise. This ratio is essentially equal to the SNR for values greater than 10 db. Conditions existing in the vicinity of the threshold were established by injecting a 25-mc band of noise into the signal path. The bandwidth of the signal path exceeded the noise bandwidth by about 10 mc. Nonlinear distortions due to band limitations preceding the FM receivers were therefore reduced.

Fig. 16 shows results of 240-channel noise loading tests with a noise spectrum extending from 60 to 1052 kc and the test channel located at 534 kc. The FMFB receiver was found to have slightly more random noise, which is believed to be due to the $1/f$ noise generated in the transistors of the feedback circuit. In Fig. 16 the excess noise effect was subtracted from the original data of the FMFB receiver. Fig. 16 shows a clear advantage of using the FMFB receiver. For 600 channels the SNR should be greater than 31 db, which is consistent with recommendations of CCIR for satellite systems. This corresponds to 40 dbrn at the zero

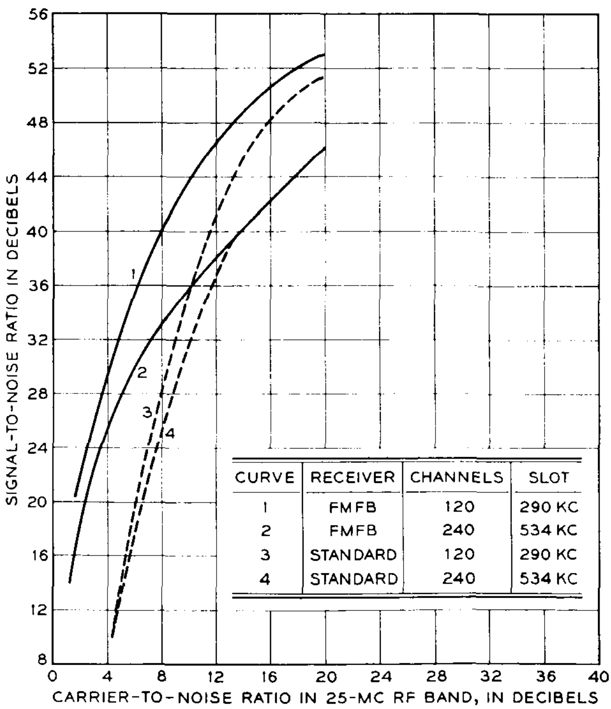

Fig. 17 — Noise loading tests; threshold region.

transmission level point. Curves were also taken at 70 and 1002 kc. They are not different in character, except that the values for SNR are higher for 70 kc and lower for 1002 kc than the ones given in Fig. 16.

With 600-channel noise loading, intermodulation in the feedback receiver begins to appear at lower modulation levels. This is due to the insufficient amount of feedback available in the circuit at higher frequencies. No useful advantage can be obtained from the present FMFB receiver in this case.

In order to show the behavior near threshold more clearly, the curves of Fig. 17 were plotted. They show the SNR in a channel as a function of the CNR at RF. An rms frequency deviation 8 db below 10 mc, or 4 mc, was chosen for these curves. The curves will eventually reach a horizontal asymptote at high CNR's. The noise will then be entirely due to intermodulation. Threshold improvements of about 4 to 5 db can be observed in Fig. 17.

V. ACKNOWLEDGMENT

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