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Description and Analysis of 890-MHz Noise-Measuring Equipment

K. W. Biber
A. C. Whittlesey

W. Shipley, Manager
Environmental Requirements Section

JET PROPULSION LABORATORY
CALIFORNIA INSTITUTE OF TECHNOLOGY
PASADENA, CALIFORNIA

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ABSTRACT

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This Report describes a narrow-band, low-noise system used with the *Ranger* spacecraft system for measuring noise at a frequency of 890.046 MHz ±50 kHz*. It includes a description of the equipment and its operation, a mathematical evaluation of system performance capabilities and of particular circuit constraints upon system operation, and a procedure for system calibration and operation.

*Hertz (Hz) has been adopted by the National Bureau of Standards to denote cycles per second.

I. INTRODUCTION

A. The Problem

The system (see Table 1) was designed to meet the goals listed below and to achieve compatibility with existing data-detection, recording, and analysis procedures. The system was to be used in the same manner as would a normal Radio Interference–Field Intensity (RI–FI) meter, i.e., for measuring the output of other subsystems to verify that they do not generate signals with a frequency and amplitude that could interfere with the spacecraft receiver (operating at 890.046 MHz).

Although noise-measuring equipment for the 890-MHz frequency range is common, the particular requirements for testing the compatibility of *Ranger* spacecraft subsystems with respect to the *Ranger* receiver could not be met by commercial equipment for the following reasons:

1. The receiver is designed to communicate over long ranges and is very sensitive.
2. Since the receiver is traveling at high speed with respect to the Earth, the doppler effect may cause the receiver frequency to shift.
3. Because of its phase-lock-loop design, the receiver has a narrow bandwidth ($\approx$100 Hz) for the frequency involved (890 MHz).
4. The very high ambient noise at 960 MHz from the television-camera transmitter could overload the RI–FI meter.

Table 1. Equipment list[a]

Item	Manufacturer	Model
Diplexer	Rantec	FLL-400B
Filter	Rantec	FLL-204B
Step attenuators	Stoddart	90506-43, 90540-6
890-Mc converter	REL	Type 864
IF attenuator	In-house	—
Receiver	Collins	R 390A/URR
Attenuator	In-house	—
Meter	Ballantine	320
X-Y recorder	Moseley	—
Motor scan	In-house	—
Noise tube	Airborne Instruments Laboratory	Type 70
Automatic noise-figure meter	Airborne Instruments Laboratory	Type 74A

[a] Serial numbers not listed because equipment may vary from test to test.

B. Design Goals

Based on the considerations given in Section A, the requirements and design goals for a *Ranger* noise-measuring system included:

1. Low noise figure at 890.046 MHz: in the vicinity of 10 db (-154 dbm threshold in 100-Hz bandwidth)
2. Narrow bandwidth: 100 Hz
3. Tracking range: 100 kHz
4. Rejection of 960 MHz
5. On-the-spot noise-figure calibration capability
6. Real-time readout of data
7. Permanent record of data for later analysis

II. SYSTEM EVALUATION SUMMARY

The mechanization of the L-band measuring system, as described in Part I, adequately met the design objectives. Several situations were encountered in which trade-offs were necessary, but these were optimized. A synopsis of the more important system parameters is given below.

1. *Scan rate.* The scan rate is the slowest rate which still allows the full frequency range to be covered in the allotted time. The maximum possible continuous-wave (CW) sensitivity of the system is thus achieved within the constraints of time and frequency range.
2. *Degrees of freedom.* The system has a reasonable number of degrees of freedom, so that the statistical approach in analyzing test results is valid.
3. *Time constants.* Time constants have been examined and their effects on continuous-wave sensitivity, statistical uncertainties, and sweep rates have been considered.

Even though the design goals were achieved, care must be taken in evaluating the overall capability of the system: for instance, because of its slow response, the system is not well suited for measuring rapidly changing spectra. Also, it cannot detect changing levels at frequencies adjacent to the scan window. Only when the spectral characteristics of the signal under observation change at a rate considerably slower than the scan period can the system be employed to its greatest advantage. The system can therefore be used to best advantage in measurements of a signal whose parameters are either stationary, or very slowly varying with time and frequency. At such times, the statistical treatment of the data obtained is valid, and the results will have a high level of confidence.

III. EQUIPMENT DESCRIPTION AND OPERATION

In this Part the important features of the 890-MHz receiver system will be described. Although some of the components (particularly the X-Y plotter) are changed from one test to another, the description is generally applicable to the whole system.

A. System Components and Operation

A block diagram of the complete 890-MHz receiver system is shown in Fig. 1. Two functionally similar systems have been constructed for dual capability. In the

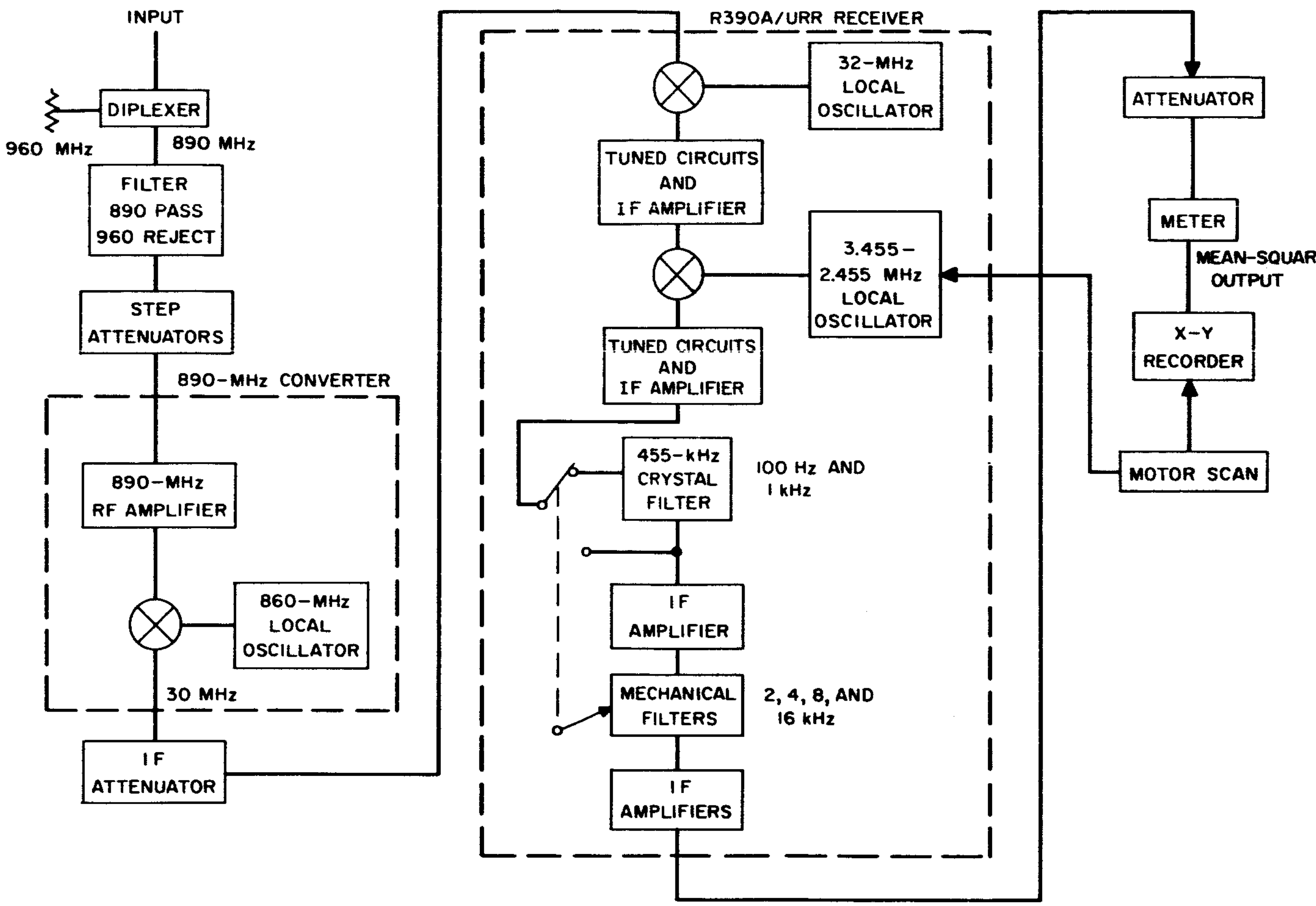

Fig. 1. 890-MHz receiver system

operation of each of these systems, the incoming signal is fed to a diplexer, which separates the relatively high-level 960-MHz noise (the television-transmitter carrier) from the desired 890-MHz signal and ensures that the system has a 50-Ω input at both 960 and 890 MHz. The following filter further rejects the 960-MHz noise. Step attenuators are available for increasing the dynamic range of the system and for use in calibration.

The filtered signal is then amplified in the 890-MHz converter, whose 30-MHz output is fed through an IF attenuator (necessary to match power levels) to the input of an R 390A/URR receiver. The resulting signal is mixed (in the R 390) with a crystal-controlled local-oscillator signal, and the difference signal is fed to a third mixer.

The third mixer has a local oscillator (LO) (2.955 ±0.5 MHz) with a variable frequency that is changed by a motor-driven front-panel control. The output of this mixer stage is a 455-kHz center frequency signal whose bandwidth can be selected from the values 100 Hz, or 1, 2, 4, 8, or 16 kHz. By varying the local-oscillator frequency, it is possible to scan the spectral components of the original input signal (890 MHz); the maximum scan range is ±250 kHz, which is limited by the 500-kHz bandwidth of the IF amplifier situated in front of the filters.

For the *Ranger VIII* and *IX* tests, the scan range has been ±50 kHz about the center frequency of 890.046 MHz, giving a total width of 100 kHz. Up to this point, the voltage amplitude is directly proportional to the input voltage (if we operate in the linear region of the receiver).

The 455-kHz output of the R 390 is subsequently fed to a true root-mean-square (RMS) voltmeter (listed accuracy 5% to 500 kHz), where the amplitude information is detected. If the mean-square output of the voltmeter is taken, a voltage output level is obtained that is proportional to the power contained in the original input signal in the bandwidth selected. Finally, the voltmeter output, along with the frequency data (derived from a potentiometer on the variable LO drive), is used to drive the Y- and X-axes, respectively, of an X-Y plotter; this completes the information detection and display process. The resulting display is a plot of power in a given bandwidth vs. frequency (and time, because of the time taken to scan the frequency range).

B. Operational Modes

As described in Section A, a motor drive is used to change the frequency in the R 390 receiver. Because the X-drive of the plotter corresponds to frequency and the Y-drive to power, a visual trace of amplitude vs. frequency is obtained. In an alternate mode of use, the R 390 receiver is set at a fixed frequency and the plotter X-axis is used as a time base. The calibration and operation procedures for the system are given in the Appendix.

IV. SYSTEM PERFORMANCE CAPABILITIES

A. Background

Many of the measurements in which the 890-MHz systems were used were made on signals whose general characteristics closely approximated the nature of random noise. Whenver this was the case, the detected signal constituted only an estimate of the true value, in that the accuracy of the estimate is influenced by such factors as the bandwidth of the filter and the averaging time available for each sample. It therefore becomes desirable to know something further about the nature of the estimates and the premises on which they are based.

First, it is necessary to establish that there are two discrete sets of data: (1) the true set, which describes the process under observation; and (2) a set which consists of the measured data points. Data from the known values of the second set will be used to make an estimate of the values describing the characteristics of the first, and unknown, set.

Next, some of the characteristics of the unknown random-noise process must be postulated. While its exact value is unknown, this process will have an average, or mean, value (sometimes called its expected value). In addition, the value of the random signal at any given time varies about the mean, so that some measure of the dispersion about the mean is required. This dispersion is generally measured by the mean-square deviation, or variance, and by its square root, the standard deviation. The variance and standard deviation are written as (Ref. 1)

$$\sigma^2 = E\left[(x-m)^2\right] = \overline{x^2} - m^2 \tag{1}$$

and

$$\sigma = \sqrt{\sigma^2} \tag{2}$$

where

E = the expected value

x = the instantaneous value of the random variable

m = the mean value of the random variable

and

σ^2 = the variance of the random variable

σ = the standard deviation of the random variable

$\overline{x^2}$ = the average value of the square of the random variable

In a practical situation the exact values of m and σ cannot be determined because of the limited amount of observation time. It is possible, however, to make estimates of their values, designated $\bar{x}$ and s, respectively. This brings to light one of the main reasons for this analysis. It is obvious from the above that the estimates $\bar{x}$ and s are themselves random variables (as they vary from sample to sample), whereas m and σ are constant parameters if the process is a true stationary random variable. For any sample of N observations, we will therefore probably obtain different values for $\bar{x}$ and s. It is important to note here that the mean, or expected, value of the random variable s is σ; that is, the bigger the samples become, the less spread out the distribution of x and s will become, and the more probable it is that x and s will be close to m and σ.

The calculations of x and s follow from the definition and are given for large values of N by (Ref. 2)

$$\overline{x} = \sum_{j=1}^{N} x_j \tag{3}$$

and

$$s^2 = \frac{\sum_{j=1}^{N} (x_j - \overline{x})^2}{N} = \frac{\sum x_j^2 - N(\overline{x})^2}{N} \tag{4}$$

where

$\overline{x}$ = the sample mean

x_j = the value of the jth sample

s^2 = the sample variance

The relationship between the estimate s and the parameter σ is readily computed:

$$\begin{aligned} E(s^2) &= E\left[\frac{1}{N}\sum_{j=1}^{N}(x_j - m)^2 - (\overline{x} - m)^2\right] \\ &= \sigma^2 - \text{variance } \overline{x} \end{aligned} \tag{5}$$

Since the variance of $\overline{x}$ is known to be σ^2/N, Eq. (5) can be rearranged:

$$\begin{aligned} E(s^2) &= \sigma^2 - \frac{\sigma^2}{N} \\ &= \sigma^2\left(1 - \frac{1}{N}\right) \end{aligned} \tag{6}$$

From Eq. (6), it can be seen that the sample variance is a biased estimate of the signal variance. The amount of bias is equal to $-\sigma^2/N$, a factor that can be easily removed. Consider the following operation:

$$\widetilde{s^2} = \frac{1}{N-1}\sum_{j=1}^{N}(x_j - m)^2 - (\overline{x} - m)^2 \tag{7}$$

This is $N/(N-1)$ times the sample variance, and

$$E(\widetilde{s^2}) = \frac{N}{N-1}E(s^2) = \sigma^2 \tag{8}$$

The amount of bias in the sample variance is negligible for estimates involving large N, so that $\sqrt{s^2}$ is still a good estimate of σ.

Next, the factor N must be determined. The sampling theorem states that if a function $f(t)$ contains no frequencies higher than B cps, it is completely determined by its ordinates being given at a series of points spaced $\frac{1}{2}B$ sec apart in the time domain (Ref. 3). In this analysis, we have a function of bandwidth B, but the sampling theorem is still valid. If the function exists for a given time period T and is sampled $\frac{1}{2}B$ sec apart, $2BT$ independent sample points will be generated (the number N in the above equations). This number of independent variables in the signal, i.e., the number of sample points, is also considered to be the number of degrees of freedom of the signal function.

One final point must be considered in the analysis before the performance of the 890-MHz system can be evaluated. The bandwidth B and the averaging time T have so far been treated as idealized quantities — especially T, which is assumed to involve true averaging. The Ballantine meter output, however, is the equivalent of an RC network output, so that a relationship stating the effects of true and RC averaging will have to be expressed.

It has been shown (Ref. 4) that for sample records four or five times as long as the time constant τ of the RC network, the continuous RC-averaged measurement will, at any instant, have an uncertainty equivalent to the uncertainty of a single measurement obtained by true averaging over a time interval $T = 2\tau$. For shorter sample records,

$$T = \frac{2\tau(1 - e^{-T_r/\tau})^2}{1 - e^{2T_r/\tau}} \tag{9}$$

where τ = the time constant of the RC detector in the Ballantine meter ($\approx$ 0.125 sec, from Ref. 5) and T_r = the length of the sample record (the time a CW signal would spend in the filter bandwidth at a given sweep rate, $\approx$ 0.6 sec in our system).

This means that the expression for the number of degrees of freedom of the signal will also be modified. For this case,

$$\begin{aligned} N &= 2BT \\ &= 4B\tau\frac{(1 - e^{-T_r/\tau})^2}{1 - e^{-2T_r/\tau}} \end{aligned} \tag{10}$$

which gives a value of about 50. Since a value of 100 is a commonly used gage, the 890-MHz system can be said to have a reasonably sufficient number of degrees of freedom.

It is interesting to note that there is a limit to the number of degrees of freedom that the 890-MHz system can possess in its present configuration. If one were to allow τ of the RC network to become much larger than T_r, the equation

$$N = 4B\tau \frac{(1 - e^{-T_r/\tau})^2}{1 - e^{-2T_r/\tau}} \tag{11}$$

would reduce to the form

$$N = 2BT_r \approx 120 \tag{12}$$

(where $e^{-x} = 1 - x + x^2/2 - \cdots$). Of course, sweep rate and transient response are lost by lengthening τ.

B. Standard Error

One of the drawbacks to using the standard deviation is that this quantity is determined in part by the amplitude of the mean value, and a recomputation is necessary every time a new average value appears. What is needed is a dimensionless quantity, such as a percentage or db-level variation, that could be applied to all the measurements obtained during a test series.

The expression for the standard deviation of the measured spectral density $\hat{G}(f)$ is given in Ref. 4 as

$$S_{\hat{G}(f)} = \frac{G(f)}{\sqrt{BT}} \tag{13}$$

where T = the quantity described in Eq. (9) and $G(f)$ = the true power spectral density of the random signal. This means that for about 67% of the measurements obtained,

$$|\hat{G}(f) - G(f)| < S_{\hat{G}(f)} \tag{14}$$

If the standard deviation is then divided by the value of $G(f)$, thus normalizing the expression, a dimensionless quantity results. This value ϵ is called the standard error for the measurement:

$$\epsilon = \frac{S_{\hat{G}(f)}}{G(f)} = \frac{1}{\sqrt{BT}} \tag{15}$$

As computed for the 890-MHz recording system for a 100-Hz bandwidth filter, the standard error is

$$\begin{aligned} \epsilon &= \frac{1}{\sqrt{2B\tau}} \\ &= \frac{1}{\sqrt{200 \times 0.125}} = \frac{1}{\sqrt{25}} \\ &= 0.2 \end{aligned} \tag{16}$$

(Since $T_r/\tau = 0.6/0.125 = 4.8$, $e^{-4.8} = 0.008$, and $e^{-9.6} = 0.00007$. Therefore, $T \approx 2\tau$ as derived from Eq. 10). By rearranging and substituting into Eqs. (13), (14), and (15), it is possible to derive a confidence interval of about 67% for the measurement of $G(f)$ in terms of the standard error:

$$\frac{\hat{G}(f)}{1+\epsilon} < G(f) < \frac{\hat{G}(f)}{1-\epsilon} \tag{17}$$

Since the measurements for the power spectral density are in dbm, it is convenient to convert the ratio to db.

$$\frac{1}{1+\epsilon} = \frac{1}{1.2} = 0.834 = -0.8 \text{ db } (1.0 = 0 \text{ db})$$

$$\frac{1}{1-\epsilon} = 1.25 = +1.0 \text{ db}$$

One is now able to say, with a 67% confidence level, that the true spectral density is within +1.0 and −0.8 db of the actual measured value at any given instant (in the case of noise-type signals).

C. Confidence Levels

Estimation by standard error is a good method for arriving at initial estimates but is rather cumbersome and inaccurate when higher confidence levels are desired. In more detailed calculations, it often is advisable to consider the specific sampling distribution for the measurement and to proceed from this point. In this method, a statistical approach in terms of the number of degrees of freedom possessed by the system is used.

Consider the nature of the signal. As it emerges from the R 390 receiver at the 455-kHz IF frequency, the signal is proportional to the voltage as seen at the antenna input terminal and can be considered to be the sum of N independent random variables, each of which has a standard normal distribution (zero mean and unit variance, as shown in Fig. 2), where $f(x)$ is the density function.

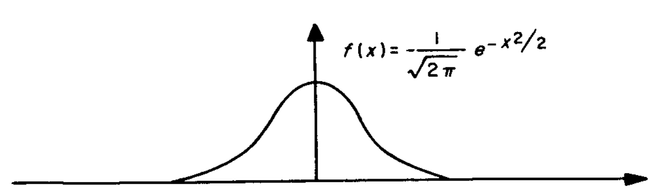

Fig. 2. Standard normal distribution

As the signal is put through the squaring circuitry of the Ballantine meter, a new function is generated. This new random variable, χ_n^2, can be defined as the sum of the squares of the N independent standard normal variables (the cross terms average out to zero), and its density function can be plotted. Naturally, since we are squaring and summing the individual terms, the shape of the distribution will vary, depending on the number of degrees of freedom N, as shown in Fig. 3 (Refs. 2 and 6).

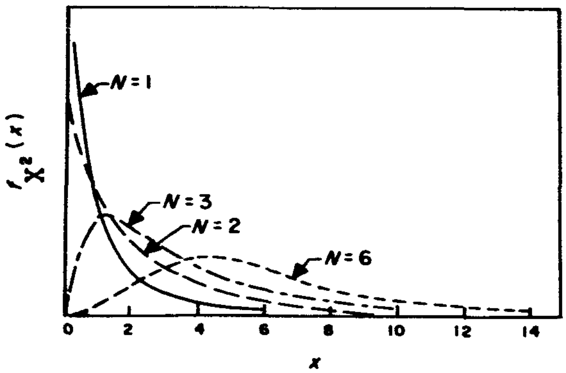

Fig. 3. Chi-square distribution vs. *N*

The general trend is now discernible. As N increases, the mean shifts to the right and the variance increases; the general shape, however, again approaches that of the normal density (central limit theorem).

A short digression is now necessary. We have let the input signal be represented by x, actually $x(t)$ and, by implication, we have let the average of x^2, or $\overline{x}^2$, be proportional to the power contained in the signal. Since the mean-square value for a random signal is equal to the total variance plus the square of the mean value (Ref. 7),

$$\overline{x}^2 = \sigma_x^2 + (\overline{x})^2 \tag{18}$$

we can determine the power in the signal by determining σ_x^2 ($\overline{x}$ is zero since the IF signal is an AC signal); $G(f)$ is now easily shown to be σ_x^2/B.

It is finally possible to establish the confidence limits; χ_N^2 is defined as

$$\chi_N^2 = \frac{\Sigma\,(x_j - m)^2}{\sigma^2} \tag{19}$$

However, m is not known and has to be estimated by the sample average x. This costs one degree of freedom, and a new expression evolves:

$$\chi_{N-1}^2 = \frac{\Sigma\,(x_j - \overline{x})^2}{\sigma^2} = \frac{(N-1)\,s^2}{\sigma^2} \tag{20}$$

where s^2 is the sample variance, and, by implication, proportional to $\hat{G}(f)$. Since σ^2 is proportional to $G(f)$, we have, after a suitable rearrangement,

$$\hat{G}(f) = \frac{G(f)\,\chi_{N-1}^2}{N-1} \tag{21}$$

The values for the chi-square distribution are listed in most tables of statistics, and are given in terms of the probability distribution vs. the degrees of freedom. Since these terms are either specified or known, one may determine any given interval $(1-\alpha)$ by the use of the chi-square distribution function and the following relationship (Ref. 4):

$$\frac{(N-1)\,\hat{G}(f)}{\chi^2_{(N-1);\,(\alpha/2)}} < G(f) < \frac{(N-1)\,\hat{G}(f)}{\chi^2_{(N-1);\,(1-\alpha/2)}} \tag{22}$$

Several representative values (listed in Table 2) afford a comparison between the interval limits and the number of degrees of freedom in a system.

Table 2. Confidence levels for various degrees of freedom

Degrees of freedom (N-1)	Confidence intervals								Method
	$(1-\alpha) = 0.67$		$(1-\alpha) = 0.90$		$(1-\alpha) = 0.95$		$(1-\alpha) = 0.99$		
	Lower limit	Upper limit	Lower limit	Upper limit	Lower limit	Upper limit	Lower limit	Upper limit	
25	0.780	1.39	0.662	1.71	0.615	1.90	0.532	2.38	Multiply the limits by the measured value of $\hat{G}(f)$ to obtain the confidence interval for the actual $G(f)$.
50	0.834	1.25	0.741	1.44	0.699	1.54	0.629	1.78	
100	0.878	1.16	0.806	1.28	0.769	1.35	0.714	1.48	
25	−1.08	+1.44	−1.79	+2.33	−2.10	+2.80	−2.74	+3.77	Add the decibels to the measured value of $\hat{G}(f)$ to obtain the confidence interval for the actual $G(f)$.
50	−0.79	+0.98	−1.30	+1.59	−1.56	+1.88	−2.01	+2.50	
100	−0.56	+0.65	−0.93	+1.07	−1.14	+1.30	−1.46	+1.71	

V. SYSTEM PERFORMANCE PARAMETERS

A. Measured Sensitivity

The system sensitivity varies from test to test, depending on the warmup time and the particular converter used. The system containing the Radio Engineering Laboratories (REL) converter can be peaked up to yield a −145-dbm sensitivity in a 100-Hz bandwidth for a 3-db signal-plus-noise to noise ratio $[(S+N)/N]$. This is equivalent to a 9-db noise figure. During frequency scanning, the CW sensitivity can be degraded as described in Section C; any losses between the receiver and the point being measured reduce the sensitivity further.

B. Circuitry-Limiting Response of the System

1. RF to IF Circuitry

All RF and IF circuits are broadband, and it is assumed that they contribute very little to the overall rise time (or time constant).

2. IF Filter

The IF filter is a high Q, 100-Hz-bandwidth filter. Its time constant is calculated below on the basis of an exponential response to a suddenly applied signal of the form given in Eq. (23):

$$V_{in} = U(t)\, e^{j\omega_0 t} \tag{23}$$

$$V_{out} = U(t)\,(1 - e^{-t/\tau})\, e^{j\omega_0 t} \tag{24}$$

where

$U(t)$ = the unit step function

$e^{j\omega_0 t}$ = the applied sinusoid of radian frequency ω_0

τ_1 = the time constant of the IF filter circuitry

From circuit theory, some approximations valid for high Q circuits are used to evaluate τ_1 (Ref. 8):

$$Q_0 \approx \frac{\tau_1\,\omega_0}{2} \text{ and } Q_0 \approx \frac{\omega_0}{\Delta\omega} \tag{25}$$

where Q_0 = the Q of the circuit and $\Delta\omega$ = the 3-db bandwidth of the circuit. From Eq. (25), we have

$$\tau_1 \approx \frac{2}{\Delta\omega} \tag{26}$$

In our case, $\Delta\omega$ is 100 Hz and τ_1 is calculated to be

$$\tau_1 \approx \frac{2}{\Delta\omega} = \frac{2}{100 \times 2\pi} = 0.00316 \text{ sec} \tag{27}$$

3. Mean-Square Detector

The mean-square detector is a device that converts an input voltage to a current proportional to the square of the input voltage. Associated with the process is a time lag that can be described by an exponential function. Given a sudden input, the output current can be approximated as follows:

$$V_{in} = U(t)\, E\,(1 + m \sin \omega_m t) \sin \omega_c t \tag{28}$$

$$I_{out} = U(t)\, KE^2\,(1 + m \sin \omega_m t)^2 \sin^2 \omega_c t\,(1 - e^{-t/\tau_2}) \tag{29}$$

where

E = the input voltage amplitude

m = the modulation index

ω_m = the modulation frequency

ω_c = the carrier frequency (445-kHz IF)

K = the proportionality constant

τ_2 = the time constant of the mean-square detector circuitry

Equation (29) can be expanded to the following form:

$$I(t) = KE^2\, U \left(1 + m \sin \omega_m t + \frac{m^2}{2} + \frac{m^2}{2} \cos 2\omega_m t - \cos 2\omega_c t - 2m \sin \omega_m t \cos 2\omega_c t - \frac{m^2}{2} \cos 2\omega_c t + \frac{m^2}{2} \cos 2\omega_m t \cos 2\omega_c t\right)(1 - e^{-t/\tau_2}) \tag{30}$$

Because m is small, and by using a low pass filter after the detector, the following terms remain:

$$I(t) \approx U(t)\, KE^2 \,(1 + m \sin \omega_m t)\,(1 - e^{-t/\tau_2}) \qquad (31)$$

The manufacturer's instruction manual (Ref. 5) gives the time constant τ_2 of the RMS voltmeter to be 0.125 sec.

4. Conclusions About the Reponse-Limiting Circuitry

By far the largest time constant associated with these circuits is that of the detector; therefore, only its time constant will be considered when the effects of a CW signal are evaluated.

C. *Effects of Noncontinuous CW Signals*

1. Response to a Transient Signal

As shown in Section B, the rise time of the IF filter is much faster than that of the detector, so the detector rise time is the limiting factor in the detection of a signal. From Eq. (31), the envelope of the detector current is

$$I(t) = KE^2 (1 - e^{-t/\tau_2}) \qquad \text{for } t > 0 \qquad (32)$$

The time to reach 50% of the final current can be found from the equation

$$\frac{I(t)}{I(\infty)} = \frac{1 - e^{-t/\tau_2}}{1 - e^{-\infty/\tau_2}} = 1 - e^{-t/\tau_2} = 0.5 \qquad (33)$$

Solving Eq. (5), we have

$$t = 0.7\,\tau_2 \qquad (34)$$

For the detector time constant $\tau_2 = 0.125$ sec, we have

$$t\,(50\%) = 0.7\,(0.125) = 0.0875 \text{ sec} \qquad (35)$$

Therefore, for a signal to cause at least a 50% response in the detector (corresponding to a meter reading 3 db lower than the actual signal power), it must remain in the IF bandwidth for a minimum of 0.0875 sec.

2. Response During Swept Measurements

We will calculate the length of time a CW signal remains in the IF bandwidth as the receiver is swept across its frequency range. Equation (32) is a general statement of response vs. time in the IF bandwidth; from this, we will calculate the loss of CW response due to a swept receiver frequency.

The IF filter passband can be approximated by a rectangular shape of 100-Hz width. It is, in fact, a high Q filter, and this should be a reasonable approximation. Given a sweep rate x, in Hz/sec, the time that a CW signal remains in the IF bandwidth Δf is

$$T = \frac{\Delta f}{x} \qquad (36)$$

The relative response at time T from Eq. (32) is

$$\text{response} = \frac{I(t)}{I(\infty)} = 1 - e^{-T/\tau_2} \qquad (37)$$

The response in decibels is

$$L = 10 \log (1 - e^{-T/\tau_2}) \qquad (38)$$

Equations (36) and (38) can be combined to give the response in terms of sweep rate:

$$L = 10 \log (1 - e^{-\Delta f / x \tau_2}) \qquad (39)$$

The value L has been calculated for various values of x using $\Delta f = 100$ Hz, and $\tau_2 = 0.125$ sec, which are the values of those parameters in the 890-MHz system. The results are presented in Fig. 4.

SWEEP RATE x, Hz/sec	RESPONSE L, db
No sweep	0
510	-1
800	-2
1142	-3
1568	-4
2105	-5
2580	-6
3636	-7
4705	-8
6153	-9
8000	-10

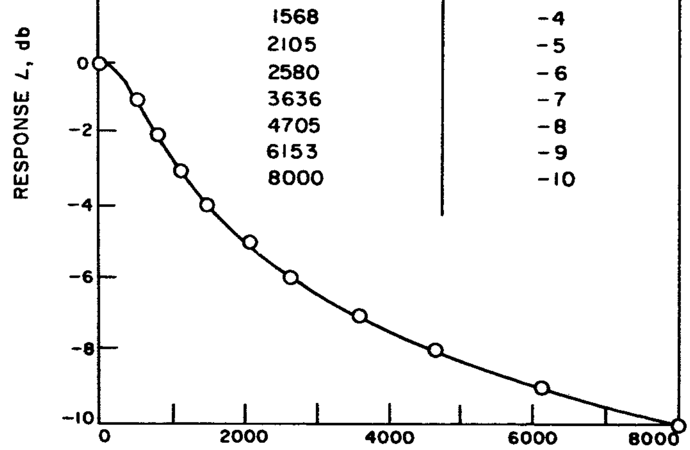

Fig. 4. Response degradation vs. sweep rate

The sweep rate of the system in its present configuration is 100 kHz/10 min = 167 Hz/sec. Figure 4 shows that the system is operating with negligible loss to CW signals: if other, faster, sweep rates were used, it might be necessary to correct the CW data for the loss due to the sweep rate.

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APPENDIX

Procedures for Use of the 890-MHz Noise-Measuring Equipment With Frequency Scan

The following procedure is used for taking readings with the 890-MHz noise-measuring rack (NMR).

A. *Pretest Calibration*

1. **Center-Frequency Calibration**

   Obtain the 890.046-MHz beacon or a frequency standard of similar stability. Connect it to the NMR and set the R390A/URR frequency so that the translated 890.046-MHz signal is within its 100-Hz passband. (For the NMR system with the REL converter, the 890.046-MHz frequency is converted to approximately 30.1932 MHz at the R 390A/URR input.)

2. **Sweep-Drive Adjustment**

   a. Loosen unit from front of R 390 by means of the Allen screws; this will disengage the gears.
   b. Turn the Helipot gear to the center of its rotation limits; it will now be capable of sweeping an approximately equal frequency region above and below the center setting when the gears are re-engaged.
   c. Set the R 390 to 30.1932 MHz to receive the beacon frequency; this will determine the center point of the frequency band to be swept.
   d. Remount the drive unit.
   e. Adjust the microswitches (high- and low-limit switches) so that the receiver sweeps about 50 kHz above and below the center frequency established by the beacon.

3. **X-Axis Adjustment of the Plotter**

   a. Connect the sweep-drive unit wires to the *X*-axis input of the plotter (black and white leads).
   b. Place the X-range switch to 750 mv.
   c. Connect two standard mercury-type cells (1.34 v) in series and attach the red and black leads.
   d. Run the sweep drive to its high-frequency range, place the X-range in variable gain mode, and adjust the gain to the amount desired.
   e. Run the sweep drive to the low-frequency range and adjust the X-zero where desired.
   f. Repeat steps (d) and (e) until 1 in. horizontally corresponds to 10 kc of frequency range on the receiver.

4. **Y-Axis Adjustment of the Recorder**

   a. Adjust the RF gain control of the R 390A/URR near 9 so as to yield 0 db on the voltmeter (on any convenient range).
   b. Plug in the cable from the square law output of the voltmeter to the *Y*-axis input of the plotter.
   c. Adjust the *Y*-axis zero and gain so that 0 db is near the bottom of the record.
   d. Disconnect the cable from the voltmeter and adjust the R 390 RF gain for a reading of 10 db on the meter.
   e. Connect the cable to the voltmeter and adjust the gain for near-maximum on-scale pen deflection.
   f. Repeat the preceding steps, adjusting the zero and gain until the *Y*-axis pen deflection corresponds to 0–10 db.
   g. Without changing the recorder zero or gain, calibrate intermediate amplitudes for later data reduction (the recorded amplitude is nonlinear).
   h. Reset the R 390 RF gain for a 0-db amplitude reading before running test.

5. **Sensitivity Measurements**

   a. *Automatic noise-figure measurement.* These sensitivity measurements are made at the connector

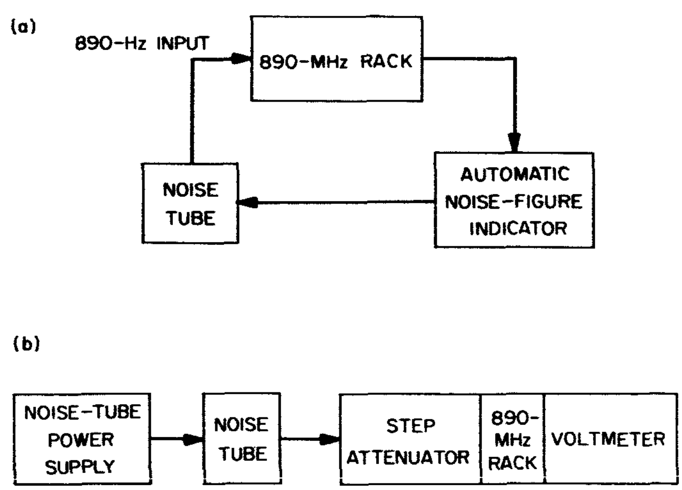

Fig. A-1. Sensitivity measurements: (a) automatic noise-figure measurement, (b) manual noise-figure measurement

cable input. If there are any additional cables in test configuration between the input and the point being measured, the receiver sensitivity as determined below must be corrected to reflect the cable losses. This is easily done as described in Ref. 6 (see Fig. A-1a). The noise figure is given in db: NF_{db}. The system sensitivity in dbm/Hz bandwidth for a 3-db signal-plus-noise to noise ratio is

$$\text{3-db sensitivity/Hz bandwidth} = -174 \text{ dbm} + NF_{db}$$

where -174 dbm is thermal noise KT at 290°K. In a 100-Hz bandwidth, the 3-db sensitivity is

$$\begin{aligned}\text{3-db sensitivity/100-Hz bandwidth} &= -174 \text{ dbm} + NF_{db} + 20 \text{ db} \\ &= -154 \text{ dbm} + NF_{db}\end{aligned}$$

b. *Manual 3-db sensitivity measurement* (see Fig. A-1b). The broadband noise tube has an excess noise output of 15.3 db. This is equivalent to a total output of -158.5 dbm/Hz bandwidth or -138.5 dbm in a 100-Hz bandwidth. With the noise tube as an input, the receiver attenuators are adjusted until the voltmeter reads 3 db above the no-input reading. If that attenuation was A_{db}, the 3-db sensitivity with no attenuation is

$$\text{3-db sensitivity/100-Hz bandwidth} = -138.5 \text{ dbm} - A_{db}$$

6. **Set Input Attenuation**

After determining the sensitivity (including cable losses), compare it to the particular specification sensitivity level. The input attenuators should be adjusted so that the resultant 3-db sensitivity is 5 db more sensitive than the specification level if possible; for example, if the specification requirement is -95 dbm and the unattenuated 3-db sensitivity is -145 dbm, the attenuators should be set at 45 db so that the resulting 3-db sensitivity is -100 dbm.

B. Test Procedure

1. Fill X-Y plotter pen with ink.
2. Allow the equipment to warm up and stabilize; ideally, 3 hr is the minimum time required.
3. Run a complete sweep with a 50-Ω termination on the cable; this should establish internal equipment performance levels.
4. Connect the cable to the antenna for system tests, and to the Division 33 communications rack for thermal vacuum tests.
5. Set the sweep to either the high or low point of its sweep region.
6. At the indicated time, turn on the sweep drive unit and plotter, and record the measurements; if possible, record any system changes, such as transmitter turn-on, on the graph.

C. Post-Test Procedure

1. Rerun a complete sweep with the 50-Ω termination to ensure that no equipment performance changes have occurred.
2. Turn off equipment.
3. Perform data reduction. The system has been found to have a certain 3-db sensitivity, S_{3db} in dbm in a 100-Hz bandwidth. The graph drawn by the plotter records 10 log $(S + N)/N$, where S is the signal power in milliwatts and N is the receiver noise power in milliwatts, which is equivalent to S_{3db} in milliwatts. By use of Fig. A-2, the ratio 10 log S/N can be determined. The signal power in dbm is then

$$S_{3db} + 10 \log S/N$$

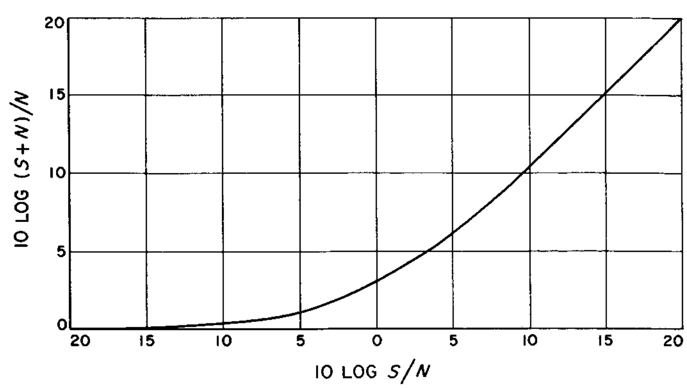

Fig. A-2. Linear receiver response to signal power (S) in the presence of receiver noise power (N)