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Noise Limited Receiver Sensitivity Measurement Technique

John P. VanDuyne, *Engineering Manager*

Types of Noise

The fundamental limitation of noise on the useful amplification of a network has long been known. For the first several decades of the radio receiver design art, the noise which supplied this limitation was out of the control of the receiver engineer. Prior to 1940, most receivers in use operated below 30 mc, a region in which atmospheric and man-made electrical disturbances supplied the noise which limited practical receiver sensitivities.[1] Laboratory determination of the sensitivity of such receivers then consisted of measuring the input level for a predetermined output. Since the external noise levels mentioned above were usually much higher than the internal noise in well designed receivers, most sensitivity specifications were written in terms of microvolts input, behind a stated dummy antenna, for a given power output.

During the 1930's, as the useful high frequency communication spectrum pushed above 10 mc, the receiving tubes then in use were found wanting, in that they generated amounts of noise which exceeded the noise from external sources. As a result, new tube designs were developed to reduce tube shot noise and induced grid noise. Although tube designers were successful in keeping tube noise lower than the external noise at frequencies below 30 mc until 1940, the problem became acute enough that the I. R. E. took cognizance of internal receiver noise in its "1938 Standard for Measurement of Radio Receivers".[2]

As the communication industry pushed above 30 mc, especially as stimulated by the pressures of World War II, it was found that the external noise sources had dropped to negligible proportions, compared to internal receiver noise, above 100 mc. The need for greater and greater radar sensitivity, VHF communication range, and less "snowy" TV pictures rapidly pushed internal receiver noise down to the level where cosmic radio background radiation (popularly called cosmic "noise", due to its randomly fluctuating character) limits useful receiver sensitivity from 30 to 100 mc, or down to 10 mc in the absence of man-made and atmospheric noise. Recent advances in low-noise receivers and high-resolution antennas used by radio astronomers are resulting in distribution maps of this radiation.[3]

In spite of their widely different origins, cosmic noise and internal receiver thermal agitation voltages are similar in character and pose similar measurement problems. Man-made and atmospheric noise are very different, in that they have a discrete impulse nature and require different measurement and system evaluation technique. The reader is referred to A. S. A. Standards C63.2 and C63.3 for a discussion of their measurement. The term "noise" used in the following discussion refers to thermal agitation phenomena unless otherwise qualified.

Measurement Objectives

A general word should be said with regard to the effect of the objective of the measurement on the choice of measurement technique. These objectives fall in several categories, the most common of which may be stated as follows:

1. The comparision, on a uniform basis, of equipments of the same design, or from the same process.
2. The comparision of equipment

of basically different design with similar desired preformance, for purposes of selecting the superior design.
3. The study of equipment performance with the intent of improving it by redesign.
4. The study of equipment performance for the purpose of learning more about the physical principles on which it operates, or to evaluate the extent to which the measured performance approaches the theoretical limit.

The techniques to be discussed require increasing degrees of skill and precision as the objective changes from (1) to (4) above.

Noise Limited Sensitivity Criteria

As previously mentioned, the reception of weak radio signals below 30 mc was limited by external noise. Hence, when measured under laboratory conditions, such receivers seemed "noiseless", since they had only sufficient amplification to produce rated output on signals supplied by a much noisier source than the laboratory signal generator. Therefore, the early concept of sensitivity was a specification of the input required, behind some specified network (dummy antenna) to produce a prescribed output. Long range communications operators found such receivers inadequate when used in quiet locations on well designed directional antenna systems. Consequently, amplifications were in-

creased, and it soon became possible to provide full rated output with no input supplied, due to the receiver's internal noise sources. An early solution to this problem was to state the sensitivity, as previously noted, but qualified by a statement of the minimum signal-to-noise ratio tolerable at rated sensitivity. A more standardized form of this was introduced in the I. R. E. 1938 Standard for Measurements of Radio Receivers.[2]

Equivalent Noise Sideband Input

This standard introduced the concept of "Equivalent Noise Sideband Input" or "ensi" as it is usually abbreviated, for the measurement of broadcast receivers. Ensi is measured by supplying an unmodulated carrier, of a specified level (E_s), through an appropriate source impedance to the receiver under test and noting the output noise power (P_n). 30% 400 cps A. M. is then applied and the output 400 cps power (P_s) is measured with the aid of a bandpass filter to eliminate the noise power. Then,

$$(1) \quad \text{ensi} = 0.3 \, E_s \sqrt{P_n/P_s}.$$

There are several possible sources of error in this measurement which must be eliminated or corrected if similar results are to be repeated at different locations (objectives 1 and 2) with different equipment, or if anything approaching an absolute performance measurement is desirable (objectives 3 and 4). First, the meter used to read P_n and P_s must be a true rms reading device, such as a thermocouple milliammeter. This meter must have reasonably constant sensitivity over the output pass band of the receiver being tested. This is necessary for a proper summation of the noise power spectrum to permit its proper comparison with the 400-cps demodulated sideband power. A second potential source of error lies in the linearity of the receiver being tested.
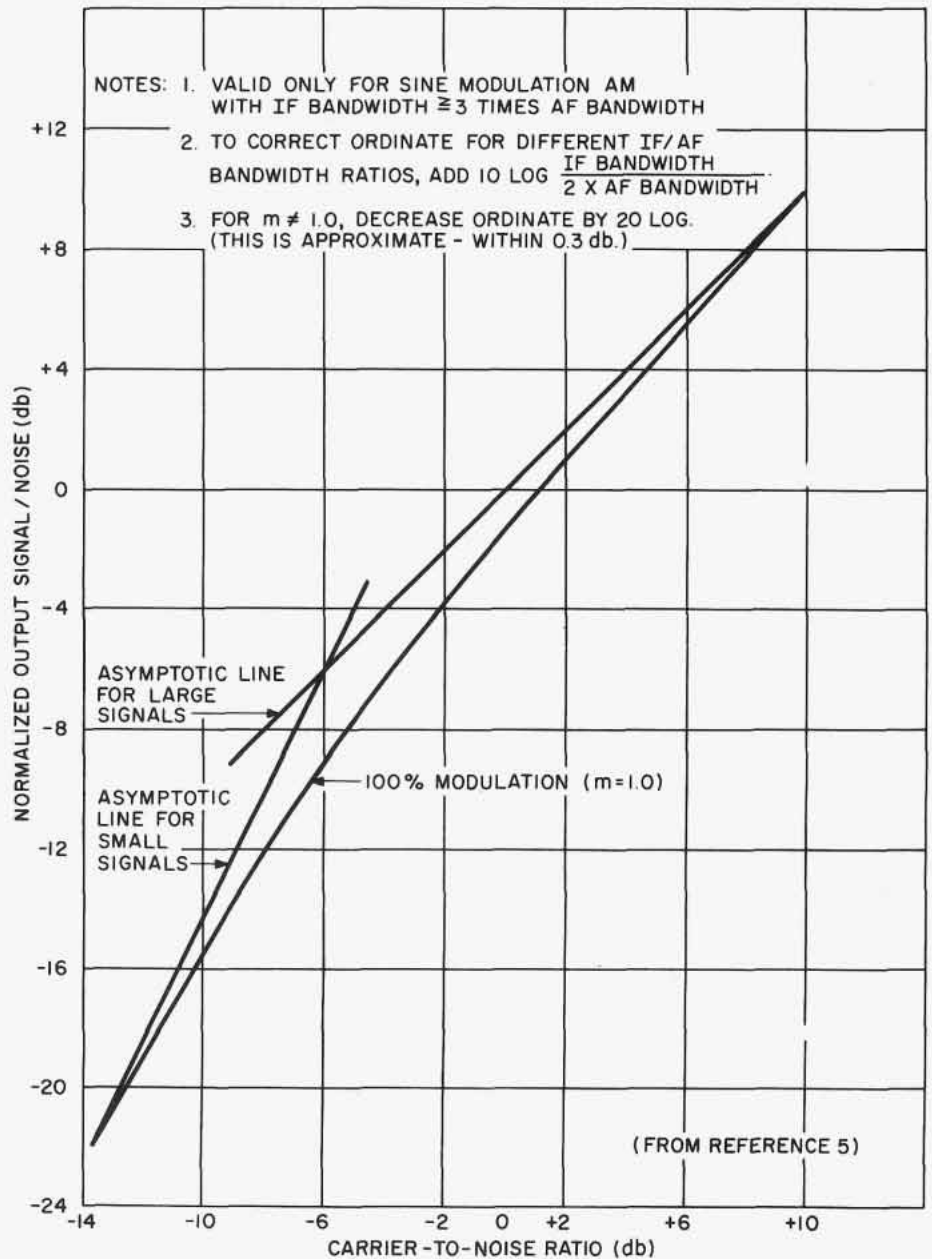

NOTES: 1. VALID ONLY FOR SINE MODULATION AM WITH IF BANDWIDTH ≧ 3 TIMES AF BANDWIDTH

2. TO CORRECT ORDINATE FOR DIFFERENT IF/AF BANDWIDTH RATIOS, ADD 10 LOG $\frac{\text{IF BANDWIDTH}}{2 \times \text{AF BANDWIDTH}}$

3. FOR m ≠ 1.0, DECREASE ORDINATE BY 20 LOG. (THIS IS APPROXIMATE – WITHIN 0.3 db.)

ASYMPTOTIC LINE FOR LARGE SIGNALS

ASYMPTOTIC LINE FOR SMALL SIGNALS

100% MODULATION (m=1.0)

(FROM REFERENCE 5)

Figure 1. Universal Curve for Output Signal-to-Noise Ratio

Since thermal noise has a peak-to-rms ratio of about 13 db[1], the receiver (and the output meter) must not overload at voltage levels up to 4.5 times that of the demodulated sideband. The receiver detector is another element, the transfer linearity of which must be considered if the ensi measurement is to be of maximum value. The stated definition of ensi ignores detector non-linearities. This is justified if the usual high-level diode peak detector (so called "linear" detector) is used.

In the description of the ensi measurement, the carrier value E_s was mentioned as "specified". As a general rule, this value should be from 3 to 10 times the resulting ensi value. The 1938 I. R. E. Standard states that the measurement is to be made at a level of 5 μv if the "absolute sensitivity" of the receiver is 5 μv or less and at 50 μv if the sensitivity is between 5 and 50 μv. These precautions are necessary, due to the fact that the output signal-to-noise ratio of an A. M. detector is a non-linear

function of the carrier-to-noise ratio. It is also a function of the I. F.-to-A. F. bandwidth ratio and it varies with the shape of the amplitude-vs-frequency response curves of these two portions of the receiver. In general, however, if the carrier-to-noise ratio into the detector exceeds 10 db, the errors in the output signal-to-noise ratio becomes negligible. Figure 1, taken from reference 4, illustrates this fact. For further detail on this matter, the reader should consult references 4, 5, and 6.

It should be noted, especially, in comparing different receiver designs for weak signal performance, that the effects of the second detector on output signal-to-noise ratio will not show up with a standard ensi measurement. For this reason the ensi measurement has given way in many specifications, to the measurement of sensitivity as the input, behind a specified source impedance, required to produce a stated output with a specified minimum signal-to-noise ratio. In evaluating equipments under such a specification, the second detector effects will usually be negligible if the output signal-to-noise ratio is 10 db or more and if the noise-free transfer characteristic of the detector is linear.

An additional proposal for a sensitivity figure which combines the concept of maximum gain with the limitations imposed by the receiver internal noise level has been suggested by J. M. Pettit.[7] In this proposal, the concept of "standard gain setting" is introduced. This is defined as the setting of the gain control which permits the delivery of a previously decided upon standard noise output to a specified load. This standard noise output must be specified for a given class of service and for specific equipments. For example, it might be specified as 0.5 milliwatts in 600 ohms as is typical for some communications equipment. The procedure for measurement is to connect the test equipment to the receiver, but with the signal generator adjusted for zero output. The receiver gain control is then adjusted for standard noise output. The output of the signal generator is then increased until "standard output" is obtained on the output indicator. The level from the generator is then noted as the receiver sensitivity. If standard noise output is not achieved, then a "maximum gain" or a "maximum sensitivity" criteria is used as previously discussed. This procedure is a formalization of
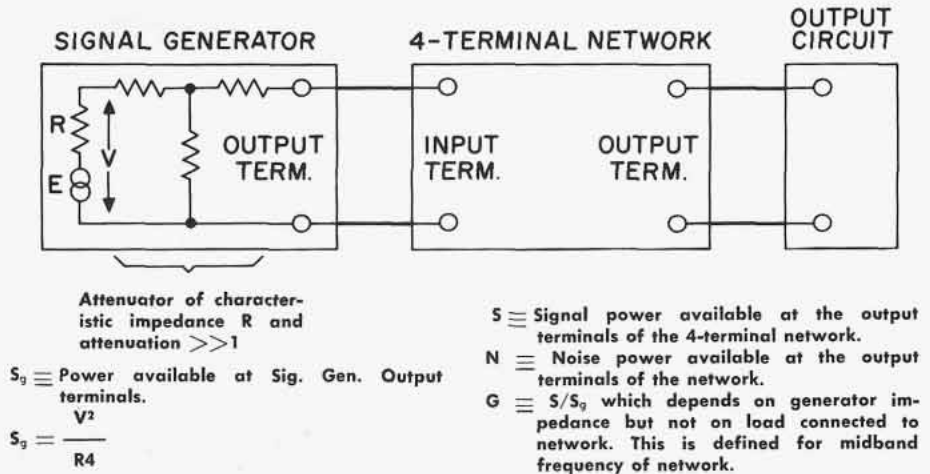

SIGNAL GENERATOR   4-TERMINAL NETWORK   OUTPUT CIRCUIT

Attenuator of characteristic impedance R and attenuation $\gg 1$

$S_g \equiv$ Power available at Sig. Gen. Output terminals.

$S_g = \dfrac{V^2}{R4}$

$S \equiv$ Signal power available at the output terminals of the 4-terminal network.

$N \equiv$ Noise power available at the output terminals of the network.

$G \equiv S/S_g$ which depends on generator impedance but not on load connected to network. This is defined for midband frequency of network.

Figure 2a. Equivalent Circuit for Noise Figure Definition

the previously mentioned general class of sensitivity specifications which state an input signal level for a specified signal-to-noise ratio.

The measurement precautions mentioned under the discussion of ensi apply to any sensitivity measurement involving the ratio of single-frequency power to the power in a noise spectrum.

Noise Figure

In an effort to work out a more basic measure of receiver performance when limited by random noise, several workers proposed rating the noise characteristics of a receiver, independent of its amplification. The specific proposal which has come into general use is that by D.O. North[8]. A later paper by Friis[9] developed a more rigorous general definition of Noise Figure. This allowed the concept to be applied to networks generally. In addition, Friis developed techniques for handling the noise performance of networks in cascade. This work permits calculation of a system noise figure from that of its components or vice versa.

Figure 2a shows the general circuit analyzed. The concept of available power is used in this analysis to avoid loss of generality due to dependence on the receiver input impedance or the load connected to the receiver output. Friis defines Noise Figure of a network (sometimes called Noise Factor and Excess Noise Ratio) as "the ratio of the available signal-to-noise ratio at the signal generator terminals to the available signal-to-noise ratio at its output term-

inals". From Figure 2b, we have the available signal-to-noise ratio of the generator as S_g/KTB and that for the network as S/N.

Available Thermal Noise = P_n Power

$P_n = \dfrac{4KTRB}{4R} = KTB$ watts.

$e_n =$ equivalent Thermal noise voltage.
$e_n^2 = 4KTB$ where: $K = 1.38 \times 10^{-23}$.
Thus, the available signal-to-noise ratio for the above generator is S_g/KTB, and the network output available signal-to-noise ratio is S/N.
(From Fig. 1, p. 419 of Reference 9)

Figure 2b. Available Thermal Noise Power from a Resistor

If $F \equiv$ Noise Figure, we have by our statement above

$$(2) \quad F = \frac{\dfrac{S_g}{KTB}}{\dfrac{S}{N}} = \frac{S_g}{S} \times \frac{N}{KTB};$$

$$(3) \quad \text{but } G \equiv \frac{S}{S_g};$$

$$(4) \quad \text{so } F = \frac{N}{GKTB};$$

(5) or $N = FGKTB$, which includes the amplified signal generator thermal noise power $GKTB$ so the available output noise, due only to sources in the network, is $(F-1)$ $GKTB$.

For simplicity, the usual measurement method is to adjust the attenuator A (Figure 2a) such that the output noise power is doubled by the generator signal. Under this condition, $S = N$ and from equation (2):

$$(6) \quad F = \frac{S_g}{KTB} = \frac{V^2}{RAKTB},$$

$$\text{if } e^2_s = \frac{V^2}{A}$$

where e_s is in microvolts delivered to a load from an R ohm source — (implicit due to our use of the available power concept).

$K = 1.38 \times 10^{-23}$ (Boltzman's constant)

$B = $ bandwidth in kc

$T = 290°$ Kelvin $(17°C)$ (for arithmetic convenience)

$R = $ generator source in ohms

$$F = \frac{e^2_s}{4RB} \times 10^6 \quad \text{(as a power ratio)}$$

$$(7) \quad \text{or } F_{db} = 10 \log \frac{(e^2_s \times 10^6)}{4RB}.$$

It is important to note that the bandwidth B introduced in Figure 2b represents the bandwidth of an equivalent rectangular power pass band of gain G and an area equal to that under the actual power gain vs frequency curve of the device being tested. In mathematical notation,

$$(8) \quad B = \frac{1}{G} \int_0^\infty G_f df$$

where $G_f \equiv$ available power gain at frequency f.
$G = $ available power gain at the frequency of the CW measurement.

In a practical case, B may be determined by plotting the squared ordinates of the voltage gain vs frequency curve as a function of frequency, calculating the area of the resulting curve graphically, and dividing by the value of G at the specific frequency, within the passband, to which the generator in Figure 2a is tuned.
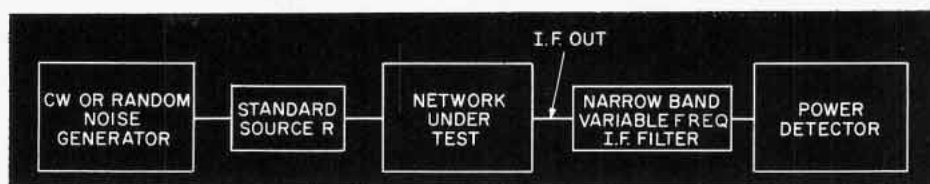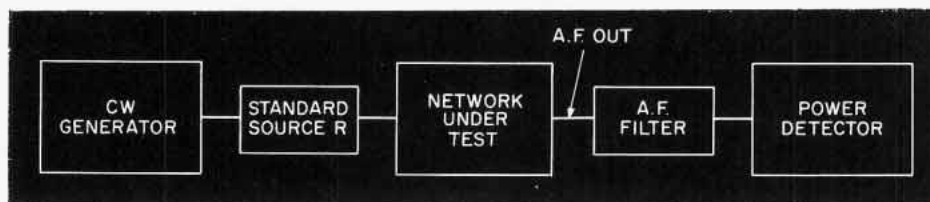

Figure 3. Single Frequency Noise Figure — I. F. Filter Method

Alternative to the use of a CW signal generator is the use of a random noise generator such as a temperature limited diode. This permits a simpler measurement to be made, without the need to determine B, but yields an answer which is actually the mean value of F. For purposes of receiver "front-end" evaluation, this is often sufficient, but for the evaluation of the noise figure of networks involving modulation or demodulation, or cascaded networks of different bandwidths, or networks in which the source resistance is a variable with frequency, the measurement of "single frequency" noise figure is often necessary if proper evaluation is to be made.

There are several types of measure-

I. F. noise spectrum is much wider than $\frac{1}{2}$ of the I. F. response (Reference 6), thus when using the I. F. filter method of Figure 3, a much narrower than expected I. F. filter is required. By the use of a CW generator and A. F. filter (Figure 4), useful data on the variation of single frequency noise figure through a network pass band can be derived rather simply. In solving equation 8 it should be noted that the value for G is that which corresponds to the frequency of measurement in the pass band. The resulting value of B is used in equation (7) to solve for the single frequency noise figure.

Note: the standardization of the output power detector for a 3db output power increase should be accomplished by introducing a known 3db change in IF gain.

Figure 4. Single Frequency Noise Figure — A. F. Filter Method

The single frequency noise figure concept as distinct from the mean noise figure (which is what has been described above) is thoroughly discussed mathematically in reference (10). The classical method of measurement is to insert a variable center frequency narrow-band filter between the network being measured and the power detector. The single frequency noise figures which result from measurement through each center frequency of the narrow filter are then weighted by their relative gains and averaged. See Figure 3 for this set-up. Since this technique is often inconvenient, a frequently useable approximation results from the use of a CW generator, a narrow-band A. F. filter, and a power detector as in Figure 4. The A. F. noise spectrum resulting from wide-band demodulation of an

ment for which the CW noise figure measurement is necessary. A typical one is the case of a receiver with a selective filter ahead of the first noise source which is comparable to the I. F. selectivity. In this case, a temperature limited diode measurement with a wide-band power detector may indicate a poor noise figure (mean noise figure). Measurement with a single frequency method may disclose a good band center noise figure, but with rapid deterioration toward the band edges, which indicates an incorrect choice of R. F. selectivity, impedance match compromise, or too wide a coupling circuit between the first and second stages of amplification if the second stage contributes appreciable noise.

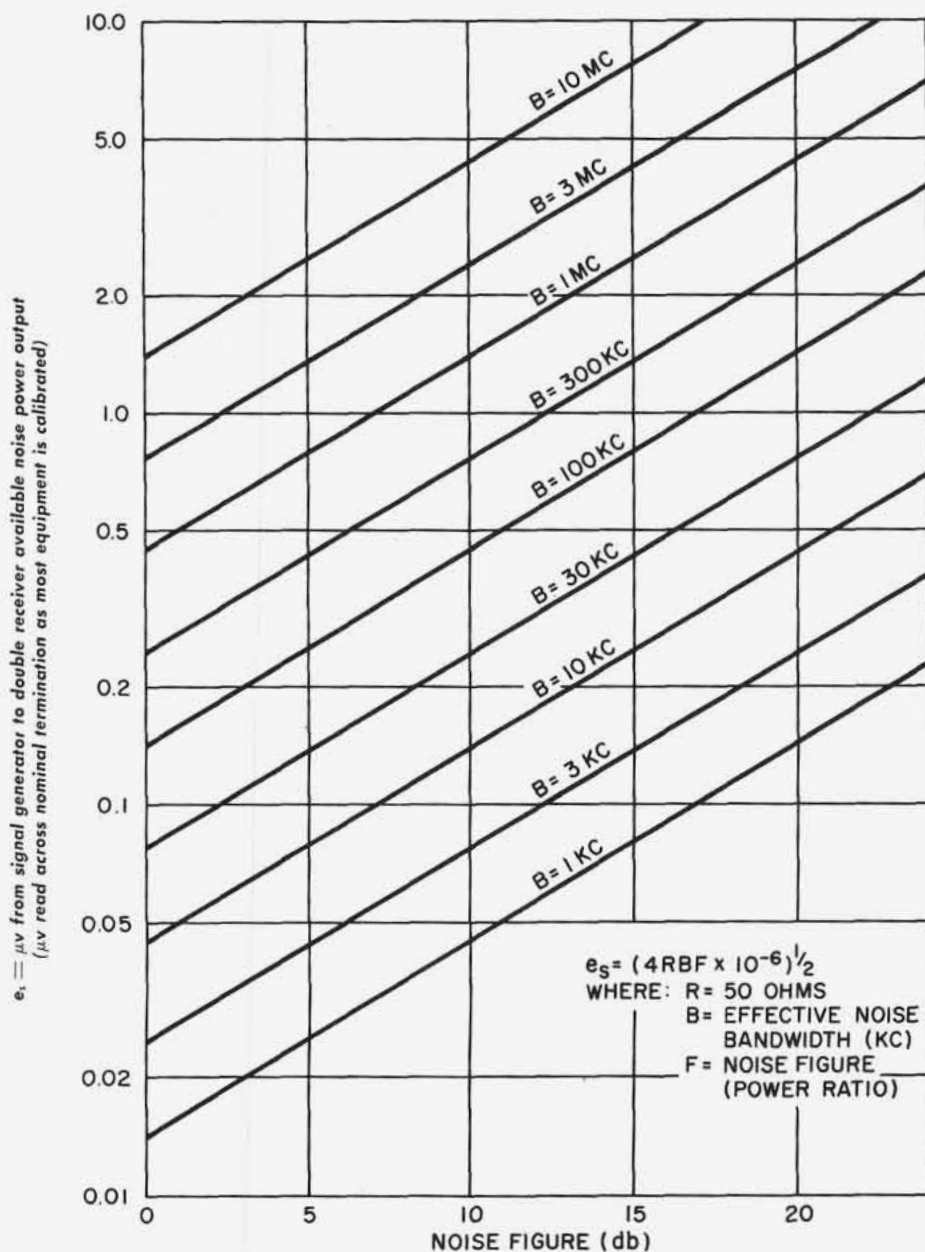Some interesting data results from solving equation (7) for e_s.

$$e_S = (4RBF \times 10^{-6})^{1/2}$$
WHERE: R = 50 OHMS
B = EFFECTIVE NOISE
BANDWIDTH (KC)
F = NOISE FIGURE
(POWER RATIO)

Figure 5.   Values of e_s for Various Noise Figures and Bandwidths

(9) $e_s = \sqrt{4RB \times 10^{-6} \text{ antilog } F_{db}}$.

Figure 5 plots these results for several bandwidths and for R = 50 ohms. Note, for example, that with a noise figure of 3 db and a bandwidth of 3 kc, the input signal, e_s, required to double the available noise power output is 0.035 microvolts. This is approximately true for a voice frequency VHF A. M. communications receiver of good design. Obviously, a generator of extremely low leakage and good low-level calibration is required to make this measurement.

An external 40-db pad is usually required to reduce the usual 1.0 —μv minimum level to the order of 0.01 μv.

There are several sources of error in noise figure measurements. Typical are those associated with the super-heterodyne selectivity of the receiver under test. In general, the I. F. and image frequency rejections must exceed 10 db if significant error is to be avoided. It is interesting to note that such spurious responses will give a pessimistic (high) noise figure if a CW generator is used

for the measurement, but an optimistic reading if a broad-noise spectrum generator is used. This is due to the fact that in the former case, the receiver is exposed to unwanted noise generators (I. F. or image) which have no signal counterpart.

Another source of error may occur in the determination of the doubled output noise power. This is not as critical as in the case of the ensi measurement, since the detector output with a CW signal at a 3-db signal-to-noise ratio is largely composed of noise sidebands. The best method is to calibrate the I. F. amplifier for a 3-db gain differential, so that the detector operating point stays the same. If the device permits, insertion of a 3-db attenuator is best, if it can be done without an accompanying change in bandpass. Alternatively, the gain control can be calibrated for a 3-db gain change. Either of these two methods can be used to calibrate the detector characteristic to answer the question of its power response. It should be noted that any attempt to use a modulated signal for noise figure measurement is beset with all the errors of the ensi method and should be avoided.

A precaution which is important to all sensitivity measurements, but especially so in the case of noise figure, is the need to accurately control the generator source impedance and noise temperature. In the absence of contrary system requirements, a resistance should be used equal to the nominal transmission line impedance for which the receiver is designed and corrected to 290° K from the actual temperature.

Conclusion

Sensitivity measurements made on receivers which have sufficiently low internal noise to detect thermal noise in the source are among the most exacting which can be made on a radio receiver. Good results require good equipment, careful set-up, and careful experimental technique. Most important, however, is a thorough understanding of the theoretical basis of the measurement and the use to which the results are to be put. The precautions and suggestions in the preceding discussion have all been thoroughly proven by extensive use in the laboratory and are offered to the reader as a guide to better experimental procedure.

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More About Transistor Measurements With The HF-VHF Bridge

GEORGE P. McCASLAND, *Sales Engineer*

The response to the transistor measurement article, which appeared in Notebook No. 19, was gratifying indeed. Many helpful suggestions and criticisms were received from Notebook readers. Your author hopes that the information presented here will serve to correct and clarify several points in question.

h_{21b} and h_{12b} Formulas

The derivation of h_{21b} as explained in the appendix to the original article is not entirely correct. Diagram 4 on page 5 of Notebook No. 19 shows e_1 to be positive toward the emitter terminal of the transistor. In diagram 7, on the other hand, e_1 is shown positive toward the base terminal. Actually, current flow shown in both cases requires that e_1 be positive toward the emitter terminal. Therefore, the derivation of the h_{21b} formula should have been given as:

$$h_{11e} = \frac{-e_1}{i_1} \text{ from diagram (7)}$$

$$i_1 = -(1-\alpha)\,i_e \text{ from diagram (7)}$$

$$h_{11e} = \frac{-e_1}{-(1-\alpha)\,i_e} = \frac{e_1}{(1-\alpha)\,i_e}$$

Now: $h_{11b} = \dfrac{e_1}{i_e}$ from diagram (4)

$$h_{11e} = \frac{h_{11b}}{1-\alpha}$$

$$-\alpha = \frac{h_{11b}}{h_{11e}} - 1 = h_{21b}$$

In deriving the formulas for h_{21b} and h_{12b}, certain approximations were made which were not specifically mentioned in the original article. Formal network theory shows these approximations. For example, from the table of Matrixes of 3-Terminal Networks on page 506 of *Reference Data for Radio Engineers*, by I.T.T. we find that:

$$h_{11e} = \frac{h_{11b}}{d}$$

From page 503 of the same book:
$$d = h_{11}\,h_{22} - h_{12}\,h_{21} - h_{12} + h_{21} + 1$$
$$\triangle^b = h_{11}\,h_{22} - h_{12}\,h_{21},$$

and

for junction transistors:
$$\triangle^b << h_{21}$$
$$h_{12} << 1$$
$$d \approx 1 + h_{21}.$$

Therefore:
$$h_{11e} = \frac{h_{11b}}{1+h_{21b}}.$$

$$h_{21b} = \frac{h_{11b}}{h_{11e}} - 1 = -\alpha.$$

Converting RX Meter Readings to Admittance and Impedance

The RX Meter directly reads out resistance in parallel with a capacitance $(+C_p)$ or an inductance $(-C_p)$. If RX Meter readings are converted to admittance, $+C_p$ converts to a positive susceptance and $-C_p$ converts to a negative susceptance. When RX Meter readings are converted to impedance, $+C_p$ converts to a negative reactance and $-C_p$ converts to a positive reactance. These changes of sign must be remembered when using the series-parallel conversion chart. Considering these changes in sign then, correct rectangular Z's in Table IV on page 4 of Notebook No. 19 are $144-j52$ and $65+j66$ while correct polar Z's are $153\underline{/-20°}$ and $93\underline{/45°}$. The signs and j's preceding C_p (ohms) in the same table are somewhat misleading and should therefore be disregarded.

Calculation of h_{21b} and α

Using the corrected h_{21b} formula and corrected values from Table IV, the new sample calculation of h_{21b} should read as follows:

$$h_{21b} = -\alpha = \frac{h_{11b}}{h_{11e}} - 1$$

$$= \frac{93\underline{/45°}}{153\underline{/-20°}} - 1 = .61\underline{/65°} - 1$$

$$= .26 + j.55 - 1$$

$$= -.73 + j.55$$

$$h_{21b} = -\alpha = .93\underline{/143°}$$

$$\alpha = .93\underline{/-37°}$$

MEET OUR REPRESENTATIVES
BIVINS AND CALDWELL
HARRY J. LANG, *Sales Manager*

Bivins and Caldwell, Inc. was founded by John F. Bivins and David J. Caldwell shortly after their return to High Point, North Carolina at the close of World War II, in the belief that electronics and radio communications would play an increasingly important role in the future industrial development of the South. The specific need for a technical group to provide local customers with sales and application engineering services on complex electronic equipment was apparent and the partners formed their organization to represent leading manufacturers of communications and electronic laboratory test equipment.

Both partners, by virtue of their backgrounds, brought a wealth of specialized experience to the new company. John Bivins majored in physics and engineering at Duke University and had been in the radio and communications business in High Point for nearly ten years prior to World War II. Dave Caldwell majored in physics and engineering at Davidson, taught physics at that school for two years after graduation, and later held posts in the production and cost accounting fields. During the War, Dave Caldwell served in the Planning Section (G3) of the Army, and John Bivins was employed as a special engineer with the Navy Department, dealing with planning and supervision of Radar installations on Naval vessels.

Industrial expansion throughout the South during the past eight years directly confirmed the early beliefs of Bivins and Caldwell and also brought about a decisive change in the Bivins and Caldwell organization. While broadcasting equipment accounted for the major portion of their business up until approximately 1949, the rapid growth of electronic manufacturers and related industries, created increasing demands for specialized services in the application of electronic instrumentation and the organization now handles precision electronic laboratory equipment exclusively. Bivins and Caldwell joined BRC in 1952 and has continuously handled our products since that time.

With increased business activity came the requirement for expansion of the company's personnel and facilities. C. M. Smith, Jr. joined the expanding com-

Headquarters of Bivins and Caldwell, Inc. in High Point, N. C.

pany in 1950. "Smitty" is an engineering graduate of North Carolina State College with extensive communications engineering experience. During World War II he served as an Electronics Officer with the U. S. Navy and was an instructor at the famed M.I.T. Radar School in Boston, Mass. Later additions to the staff have increased total personnel to over 21; 9 of which are field engineers who concentrate exclusively on customer problems.

The company's facilities have also been increased, with branch offices in Atlanta, Georgia, and Orlando, Florida. The new headquarters building in High Point is one of the most modern and best arranged office facilities in the area. The Service Department here, under the direction of Mr. Robert L. Moore, has complete repair and recalibration facilities and is well equipped to service all products that are currently handled. All offices are equipped

Shown in conference are, left-to-right, John Bivins, Douglas Severence, David Caldwell, and C. M. Smith, Jr.

with TWX service for efficient communication with all factories. The company also maintains a fleet of eight vehicles for use by field engineers. The Bivins and Caldwell Fall Road Show, which runs for about four weeks, covers some 4,000 miles and is an annual event attended by several thousand engineers in the local area.

Bivins and Caldwell believes that their organization must operate on the premise of fundamental engineering integrity. Their basic objective is to assist customers in solving engineering problems and to recommend the best available test equipment and techniques for the particular job to be done. We at BRC proudly salute Bivins and Caldwell for their faithful service to our many valued customers throughout the South.

EDITOR'S NOTE

John P. Van Duyne Appointed Engineering Manager at BRC

The appointment of John P. Van-Duyne as Engineering Manager, effective August 18, 1958, has been announced by Dr. George A. Downs-brough, President. Mr. Van Duyne comes to BRC with 15 years of experience in the fields of engineering and electronics. The major part of his experience has been in the development and production of electronic instruments.

After receiving his Bachelor of Science degree in Electrical Engineering from Rensselaer Polytechnic Institute of Troy, New York in December 1943, Mr. Van Duyne served with the U. S. Signal Corps where he instructed in radar and radio relay techniques. During his last year of service, he was engaged in the design of Radio Countermeasures epuipment in the Coles Laboratory at Fort Monmouth, New Jersey.

Following his discharge from the Signal Corps in 1946, Mr. Van Duyne joined the Measurements Corporation of Boonton, N. J. as Project Engineer and was engaged in the development of signal generators.

In August 1948, he became associated with the Allen B. DuMont Laboratories, East Paterson, New Jersey, serving successively as Senior Engineer, Section Head of the Advanced Development Section, and Section Head of the Circuit Design and Development Section.

From March 1953 until he joined BRC in August, 1958, Mr. Van Duyne held posts with the Westinghouse Electric Corporation in Metuchen, New Jersey. He served as Engineering Section Manager and was engaged in the design of color television receivers until January 1956 when he was appointed Manager of TV Engineering. He served in the latter capacity until he joined BRC.

An active radio "ham", Mr. Duyne is also a member of Tau Beta Pi, Eta Kappa Nu, Sigma Chi and the Institute of Radio Engineers.

JOHN P. VanDUYNE

Engineering ⓑⓡⓒ Representatives

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