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MEASUREMENT OF TRANSMITTER NOISE AT HIGH FREQUENCY THROUGH MICROWAVE FREQUENCIES

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A tutorial review of the basis for transmitter noise measurements shows that noise is best described and measured as AM and FM noise. The contribution of AM noise to RF spectrum shape is determined by the power spectral density shape of the AM noise. The contribution of FM noise to the RF spectrum is to make the shape that of an RLC resonant circuit response rather than a delta function with a sideband structure. AM noise is measured with a direct detector diode. FM noise for frequencies above 5 GHz is measured with a discriminator based on a one-port cavity. FM noise below 15 GHz is measured with an improved over

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transmission line discriminator which is described in detail. The most widely used baseband analyzer is the constant bandwidth superheterodyne wave or spectrum analyzer. The baseband spectrum analyzer has been automated. Complete calculator programs with brief descriptions are presented in the Appendix.

CONTENTS

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The work presented in this report is the result of 26 years of experience by the authors which makes the task of proper acknowledgement difficult. Whenever a theoretical problem was encountered, the answer or a method of attack was found in the writings of Dr. James Mullen and Dr. David Middleton. Dr. Mullen and Dr. Middleton deserve commendation for their excellent research. The measurement work which inspired our contributions to this subject has been performed in the Gainesville, Florida and Clearwater, Florida laboratories of the Sperry Rand Corporation and the Restone Arsenal, Alabama laboratories of the US Army Missile Command. Without the expensive physical resources of these laboratories, these accomplishments would not be significant. Correspondence and conversations with colleagues too numerous to call by name have added significantly to our knowledge of this topic and deserve appreciation.

Dr. Ashley acknowledges the benefit of many years of collaboration with Mr. Frank Palka of the Sperry Microwave Electronics Division of the Sperry Rand Corporation.

The measurement work performed at Redstone Arsenal, Alabama in the US Army Missile Research, Development and Engineering Laboratory has been greatly aided by the untiring work of Mr. R. Hammond. Mr. W. L. Low, Chief Radar Technology, Advanced Sensors Directorate has greatly supported this work through patience and words of encouragement.

Development of the automated measurement technique required experiments performed at the Hewlett-Pakard Loveland Instrument Division, the US Army Missile Command Laboratories, and at the University of Colorado. Mr. Charles Reynolds of Hewlett-Packard taught us how to use the interface bus and write the software.

I. BACKGROUND CONCEPTS

Noise in a microwave transmitter is an unwanted signal which degrades the ability of the system to transmit communications or other information. Stochastic signals and coherent spurious signals (usually related to bias supply ripple or mechanical vibration) must be included when considering transmitter noise. In the first stages of receiving systems, concern is directed toward a ratio of desired signal to noise in the order of unity. In a transmitter, the ratio of carrier to noise is of the order of 10^3 or more. Most of the amplifying or oscillating devices in a transmitter are operating in a saturated mode; thus, some of the linear system theory ideas that are so useful in the study of receiving systems must be used with care in studying a transmitter system.

The use of a microwave transmitter would also require a detection system which may respond to either amplitude modulation or angle modulation (an FM discriminator), but not both. The essential question is, "How much of the noise in a transmitter will come out of a detection system?" The kind of a detector being used will determine if the user is concerned with what has come to be called "FM noise" or "AM noise." Fortunately, the study of oscillator theory as pioneered by Middleton [1], Mullen [2], and others yields answers in terms of FM or AM noise; and even more fortunately, the measurement techniques which yield the best data are based on separately detecting AM and FM noise [3, 4].

The measurement of transmitter noise is accomplished by applying a suitable sample of the output to detectors which respond to either AM or FM. It is easy to visualize an AM detector which rejects a moderate amount of incidental FM on the signal being tested. It is not very easy to visualize a microwave discriminator without a limiter stage rejecting incidental AM on the signal being tested; however, the cavity resonator discriminators of Marsh and Clare [5], Ashley, et al [3], and the new transmission line discriminators of Ashley, et al [6] do reject incidental AM. Any competitive method must reject the unwanted form of modulation in measuring the desired form.

Current microwave transmitters include transmitters with a simple high power oscillator (such as a magnetron) as the only device, those which include direct microwave oscillators [reflex klystrons, transferred electron oscillators (TEO), IMPATT], driving amplifiers (amplitron, klystron, traveling wave tube (TWT), locked IMPATT oscillator, transistor), and those which synthesize the microwave signal from sources controlled by VHF or UHF crystal oscillators and then increase the power with amplifiers. To measure noise in transmitters, one must be able to study direct oscillators, amplifiers, and the VHF or UHF oscillators, amplifiers, and frequency multipliers.

The range where the demodulated noise must be studied varies with the applications. Doppler radar systems are degraded by excessive noise in the modulation range 10 Hz to 100 kHz. The usual noise specification covers a significant portion of this range. Communications systems using FM multiplex the subcarriers into a band typically 10 MHz wide, and this sets the upper limit on the measurement of demodulated noise. These modulation frequencies are sometimes described as distance from the carrier with an implied idea of the RF spectrum spread.

In FM communications, the nonlinearity in amplifiers can cause cross modulation of multiplexed subcarriers which is a form of distortion. This will appear in the system output as unwanted signals in the individual channels. This meets the definition of noise as being unwanted. However, this problem will not be treated further in this work.

Understanding noise and the mathematics for studying noise requires careful definitions and explanations. Some of the intuitive ideas about noise based on modulation theory using sinusoids only are imprecise. The following section presents some of the concepts needed.

II. MODULATION NOISE RELATIONSHIPS

Most of the relationships to be presented are tersely abstracted from the works of Dr. David Middleton [7] and by Dr. Athanasious Papoulis [8]. The proofs and many details are presented in the works of Middleton and Papoulis.

For amplitude modulation transmitters, the output can be written

$$v(t) = A_o[1 + \lambda V_s(t)] \cos(\omega_c t + \phi) \tag{1}$$

where

t = time

A_o = unmodulated amplitude, peak volts

$\omega_c = 2\pi\nu_c$ = carrier frequency, rad/sec

ν_c = carrier frequency, Hz

ϕ = phase constant usually set to zero by proper choice of coordinates.

and $\lambda V_s(t) \leq 1$ is the information to be communicated. Interest is directed toward the noise floor which exists when $V_s(t) = 0$. Thus, $\lambda V_s(t)$ is replaced by $N_A(t)$, where $N_A(t)$ is the AM noise and must be treated as a random variable.

To distinguish the frequency variable near the carrier from frequency in the baseband region where modulation noise must be described, ν will be used for frequency (Hz) near the carrier frequency, ν_c. In the baseband region from zero to a few megahertz, f will be used to designate frequency (Hz).

For angle-modulated transmitters, the output can be written:

$$v(t) = A_o \cos(\omega_c t + \Phi(t) + \phi) \quad , \tag{2}$$

where the information to be transmitted is carried in the instantaneous phase $\Phi(t)$. The instantaneous frequency, ν_i, of this signal is:

$$\nu_i = \frac{d}{dt} (\omega_c t + \Phi(t) + \phi) = \omega_c + \dot{\Phi}(t) \quad . \tag{3}$$

Interest is in the residual $\Phi(t)$ or $\dot{\Phi}(t)$ for the unmodulated state which will be called $\Phi_N(t)$ or $\dot{\Phi}_N(t)$. Again, $\Phi_N(t)$ must be considered as a random variable. From a measurement standpoint it is fortuitous that $\dot{\Phi}_N(t)$ is usually measured with a discriminator, and the results can be integrated to find $\Phi_N(t)$.

Combining these equations, the signal from an unmodulated transmitter can be written

$$v(t) = A_o(1 + N_A(t))\cos(\omega_o t + \Phi_N(t) + \phi) \quad , \tag{4}$$

where

$$|N_A(t)| \ll 1 \tag{5}$$

and

$$|\Phi_N(t)| \ll 1 \quad . \tag{6}$$

The task is to learn something about $N_A(t)$ and $\Phi_N(t)$ by measurements made on $v(t)$.

Except for components caused by bias supply ripple or deterministic mechanical vibrations, the signals $N_A(t)$ and $\Phi_N(t)$ are not deterministic and must be studied from a statistical point of view. Any dc average is absorbed in either A_o or ω_c, therefore it is appropriate to consider that $N_A(t)$ and $\Phi_N(t)$ have zero mean, and a study of the density function, variance, autocorrelation function, and band-limited spectrum must be made.

The probability density function for $N_A(t)$ or $\Phi_N(t)$ is difficult to measure. It is usually assumed, by implication, to be Gaussian because this is the mathematically best known function. In terms of physical processes in oscillators and amplifiers, this is not an unreasonable assumption.

The usual definition of an autocorrelation function can be applied to $A_N(t)$ and $\Phi_N(t)$

$$R_A(\tau) = \lim_{T \to \infty} \frac{1}{2T} \int_{-T}^{T} N_A(t)\, N_A(t + \tau)dt \tag{7}$$

$$R_\phi(\tau) = \lim_{T \to \infty} \frac{1}{2T} \int_{-T}^{T} \Phi_N(t)\, \Phi_N(t + \tau)dt \tag{8}$$

$$\tau = t_2 - t_1 \quad , \tag{9}$$

and for the real random variables $N_A(t)$ and $\Phi_N(t)$

$$R(-\tau) = R(\tau) \quad . \tag{10}$$

The autocorrelation function is used as the mathematical link between the time domain and frequency domain descriptions of random variables. The Wiener-Khintchine theorem states that for ergodic and well-behaved real processes the power spectral density $S(f)$ and the autocorrelation functions $R(\tau)$ are a Fourier transform pair:

$$S(f) = \int_{-\infty}^{\infty} R(\tau) \cos(2\pi f\tau)d\tau \tag{11}$$

$$R(\tau) = \int_{-\infty}^{\infty} S(f) \cos(2\pi f\tau)df \quad . \tag{12}$$

The power spectral density, $\underset{\sim}{S}(f)$, is an even, real function and is non-negative

$$\underset{\sim}{S}(-f) = \underset{\sim}{S}(f) \qquad (13)$$

$$\underset{\sim}{S}(f) \geq 0 \quad . \qquad (14)$$

$\underset{\sim}{S}(f)$ carries no phase information*. The power spectral density of the noise modulations is band-limited to a small fraction of the carrier frequency by resonant circuits in the signal generation source.** This limited band is usually termed the "baseband."

In terms of the measurement problem, applying the source to demodulators as shown in Figure 1 yields voltages proportional to $N_A(t)$, $\Phi_N(t)$, and $\dot{\Phi}_N(t)$ which are functions of time. The simplest data processing in the time domain is to use a cathode ray or mechanical oscillograph to record instantaneous amplitude, phase, or frequency versus time. The records are often called "stability" although the word instability might be more appropriate. In terms of frequency or phase instability, the time domain processing can be quite sophisticated if long term stability (time period of days or more) is important, e.g., timekeeping applications. The use of computing counters to determine the "sigma versus tau" characteristic is described by Shoaf, et al [15].

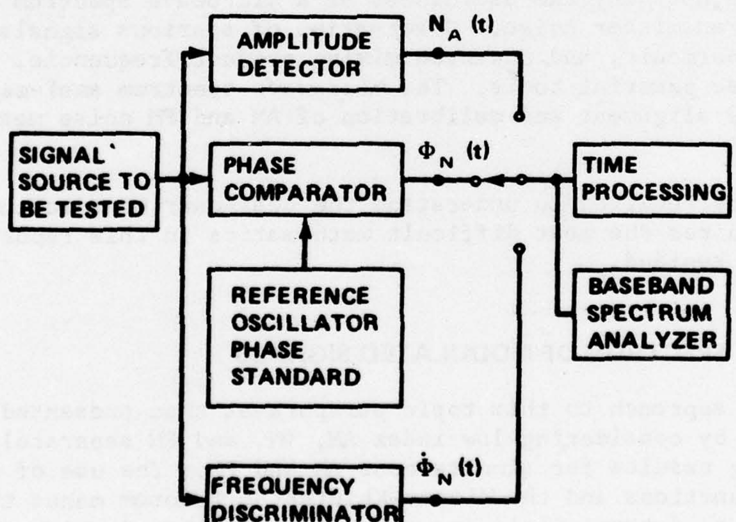

Figure 1. Schematic representation of modulation noise measurements.

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*Note: Phase information, not phase modulation.

**This is not true for TWT amplifiers or backward wave tube oscillators.

In theory, it should be possible to apply the output of the demodulators to autocorrelation computing equipment. There are no known published results of such an experiment. One experiment which would be of great significance in oscillator noise theory would be to determine the cross correlation between amplitude and phase noise modulation.

Because most microwave transmitters are used in either radar or communication systems, the most useful analysis of the modulation noise is in the frequency domain. The baseband analysis equipment is usually called a spectrum analyzer if frequency is automatically swept, and a wave analyzer if the frequency is manually set. Because of their importance, these equipments are discussed in a later section.

Nothing has been presented concerning the microwave spectrum of v(t). The main reason is that the measurement of the signal spectrum will not predict the effects of transmitter noise on the amplitude or angle demodulators in the receiver. It is not possible to determine from the signal spectrum alone whether a particular portion of a sideband is caused by $N_A(t)$ or $\Phi_N(t)$. If one needs to know the spectrum structure in the region of the carrier, it will be found hidden under the noise floor of the spectrum analyzer. Thus, $N_A(t)$ and $\Phi_N(t)$ must be measured and used to compute the near-carrier spectrum.

This does not deny the usefulness of a microwave spectrum analyzer in studying transmitter noise. Observation of spurious signals at subharmonic, harmonic, and unwanted mixing product frequencies is best done with these powerful tools. The microwave spectrum analyzer will be used in the alignment and calibration of AM and FM noise measuring equipment.

The theory required to understand the near-carrier spectrum structure requires the most difficult mathematics in this report, but should not be avoided.

III. THE SPECTRUM OF MODULATED SIGNALS

The approach to this topic paraphrases that presented by Middleton [7] by considering low index AM, ΦM, and FM separately, and then combining results for simultaneous AM and FM. The use of autocorrelation functions and the Wiener-Khintchine theorem makes the results apply to deterministic (usually sinusoidal) and stochastic modulation signals. First, for amplitude modulation only,

$$v(t) = A_o(1 + \lambda V_M(t)) \cos(2\pi\nu_c t) \quad , \tag{15}$$

where

$$N_A(t) = \lambda V_M(t) \quad .$$

The spectrum, $\underset{\sim}{S}_{RF}(\nu)$ is found from (Middleton [7], equation 12.10b)

$$\underset{\sim}{S}_{RF}(\nu) = A_o \delta(\nu - \nu_c) + \lambda^2 A_o^2 \int_0^\infty R_M(\tau) \cos 2\pi(\nu - \nu)\tau \, d\tau \quad , \tag{16}$$

where $R_M(\tau)$ is the autocorrelation function of the modulation, $V_M(t)$. The delta function, $\delta(\nu - \nu_c)$, indicates a spectral "line" at ν_c. If $V_M(t) = \cos 2\pi f_a t$ and $\lambda < 1$, then

$$S_{RF}(\nu) = \frac{A_o^2}{2} \delta(\nu - \nu_c) + \frac{A_o^2\lambda^2}{8} \delta[\nu - (\nu_c + f_a)] \tag{17}$$

$$+ \frac{A_o^2\lambda^2}{8} \delta[\nu - (\nu_c - f_a)] \quad ,$$

which is the familiar carrier with a pair of side frequencies located at $\nu_c \pm f_a$. If V_M is a more complex signal which can be expressed as a Fourier series of sinusoids, then pairs of side frequencies will appear for each Fourier component. There will be no cross products as long as the total modulation index is less than one (i.e., no over-modulation).

A very useful result can be deduced for the case of any form of low index amplitude modulation. If the spectrum of V_M were found from its autocorrelation function, the following would be evaluated

$$\underset{\sim}{S}_{AM}(f) = 2 \int_0^\infty R_M(\tau) \cos 2\pi f\tau \, d\tau \quad , \tag{18}$$

while the sideband portion, $\underset{\sim}{S}_{SB}(\nu)$, of Equation (16) requires evaluation of

$$\underset{\sim}{S}_{SB}(\nu) = \int_0^\infty R_M(\tau) \cos 2\pi(\nu - \nu_c)\tau \, d\tau \quad . \tag{19}$$

An obvious change of variable yields

$$\underset{\sim}{S}_{RF}(\nu) = \frac{A_o^2}{2} \delta(\nu - \nu_c) + \frac{\lambda^2 A_o^2}{2} \underset{\sim}{S}_{AM}(\nu - \nu_c) \quad , \tag{20}$$

which reveals that the spectrum of the modulating signal is shifted from being symmetrical about F = 0 to being symmetrical about $\nu = \nu_c$. Therefore, if the spectrum of the AM noise, $\underset{\sim}{S}_{AM}(f)$ can be measured, then a prediction of the contribution of AM noise to the RF spectrum can be made using Equation (20). This important concept is illustrated by Figure 2. The measured noise spectrum of the AM, $\underset{\sim}{S}_{MAM}(f)$, is shown

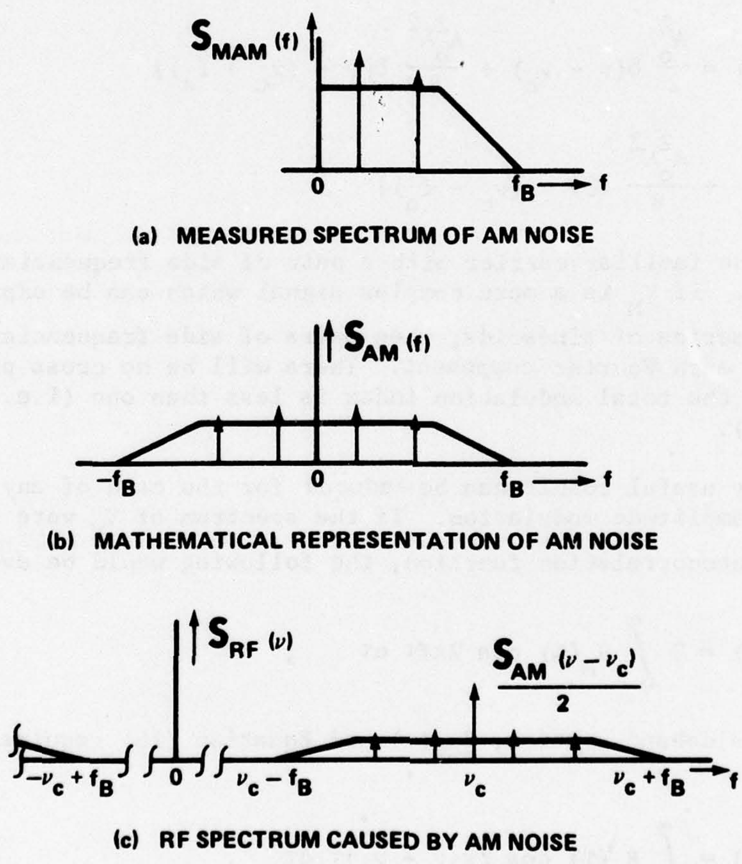

Figure 2. Example of AM noise spectrum.

as it would be measured in Figure 2(a). This is a power spectral density for the baseband range. Mathematically, the negative frequency portion of the spectrum must be carried, which is the mirror image of the positive portion as shown in Figure 2(b). The fact that $\underset{\sim}{S}_{AM}(f)$ is the AM spectrum means that the RF spectrum can be found by translating $\underset{\sim}{S}_{AM}(f)$ to the regions near $\pm\nu_c$ and adding the delta function representing the carrier [Figure 2(c)]. This example of AM noise spectrum has delta functions to represent sinusoids and a continuous portion to represent random noise.

For the case of angle modulation only, the RF signal can be written

$$v(t) = A_o \cos[2\pi\nu_c t - \Phi(t)] \quad , \tag{21}$$

where $\Phi(t)$ is chosen for phase or frequency modulation as

$$\Phi(t)_{\Phi M} = D_\Phi V_M(t) \tag{22}$$

$$\Phi(t)_{FM} = D_F \int^t V_M(t')dt' \quad . \tag{23}$$

D_Φ has units of rad/V and D_F has units of angular frequency (rad/sec) per volt. For the case where V_M is sinusoidal, let

$$\Phi(t) = \mu_{\Phi M} \cos 2\pi f_a t \tag{24}$$

for phase modulation and

$$\Phi(t) = \mu_{FM} \sin 2\pi f_a t \quad , \tag{25}$$

where $\mu_{\Phi M} = \Delta\Phi$, the magnitude of the phase deviation and $\mu_{FM} = \frac{\Delta\nu}{f_a}$, the frequency deviation divided by the modulation frequency are all the appropriate modulation indicies for ΦM and FM. With little effort and using the Bessel function manipulation, Middleton ([7], eq 14.10b ff) found the autocorrelation function and the spectrum

$$R_v(\tau) = \frac{A_o^2}{2} J_o\left(2\mu \cos \frac{2\pi f_a \tau}{2}\right) \cos 2\pi\nu_c \tau \tag{26}$$

$$\underset{\sim}{S}_{RF}(\nu) = \frac{A_o^2}{2} \sum_{m=o}^{\infty} J_m^2(\mu)\{\delta[\nu - (\nu_c + mf_a)] + \delta[\nu - (\nu_c - mf_a)]\} \quad ,$$

where J_m is an m^{th} order Bessel function. This is the familiar result with a multitude of side frequencies for large μ. If $\mu < 0.01$, only the m = 1 side frequencies are significant, and the approximation $J_1(t) = \frac{x}{2}$ is often useful. This theory is the basis for calibration of the discriminators used for FM noise measurements.

If more than one sinusoid is present in the modulation signal, then additional side frequencies are added for low values of μ. For large μ, the nonlinearity of angle modulation causes some cross coupling of the modulation. For example, if two sinusoids at 300-Hz and 1000-Hz angle modulate a carrier, the spectrum will have components located at 1300, 1600, 2300, 2900, etc., Hz away from the center frequency as well as at the harmonics of 300 and 1000 Hz. These facts and the expression $\mu_{FM} = \frac{\Delta\nu}{f_a}$ should indicate trouble when the modulation is not deterministic.

The theory is not a disappointment. The limiting behavior for small modulation is all that is necessary to consider in this study. Since the main concern is with the case where the modulation is the residual noise in the source, Middleton's equations ([7], Sec. 14.2-1) can be used and higher order terms abandoned because Φ is quite small. For phase modulation,

$$R_V(\tau) \approx \frac{A_o^2}{2} \cos 2\pi\nu_c\tau \,\{1 + R_{\phi M}(\tau)\} \quad , \tag{28}$$

which is Fourier-transformed to find the RF spectrum

$$\underset{\sim}{S}_{RF}(\nu) \approx \frac{A_o^2}{2}\,\delta(\nu - \nu_c) + \frac{A_o^2}{2}\int_o^\infty R_{\phi M}(\tau)\cos 2\pi\nu_c\tau \cos 2\pi\nu\tau \, d\tau \, . \tag{29}$$

In the integral, the identity for cos A cos B can be used (a term in $\cos 2\pi(\nu_c + \nu)$ abandoned by a stationary phase argument) and the same substitution used in the AM theory to show:

$$\underset{\sim}{S}_{RF}(\nu) = \frac{A_o^2}{2}\,\delta(\nu - \nu_c) + \frac{A_o^2}{4}\,\underset{\sim}{S}_{\phi M}(\nu - \nu_c) \quad . \tag{30}$$

This result shows a similar limiting behavior as was found in the AM case. A carrier line is the center frequency for the symmetrical phase modulation spectrum which is unchanged in spectral shape. Thus, $\underset{\sim}{S}_{\phi M}(f)$ is measured and very quickly translated to the RF spectrum.

It is discouraging to look at typical measurements of $\underset{\sim}{S}_{\phi M}(f)$ and study oscillator theory. Leeson [9] shows general forms for $\underset{\sim}{S}_{\phi M}(f)$ which show a behavior below approximately 1 kHz which is of the order $\frac{1}{f^2}$ or $\frac{1}{f^3}$. Reverting back to the theory with this knowledge of $\underset{\sim}{S}_{\phi M}(f)$ shows that all of the approximations were not very well justified. Those who use the "theorem" implied in Equation (31) would be in for a shock if $\underset{\sim}{S}_{RF}(\nu)$ could be measured for a signal where $\underset{\sim}{S}_{\phi M}(f)$ is known.

The true understanding of the RF spectrum for small angle modulation comes from the study of limiting forms for FM noise modulation. The classic reference is Mullen and Middleton [10], although the word is well documented by Middleton ([7] Sec 14.2-1). The mathematical trouble originates in the value of an integral,

$$L = \int_0^\infty \frac{\underset{\sim}{S}_{FM}(f)}{4\pi^2 f^2}\, df \quad , \tag{31}$$

needed in finding the autocorrelation function. There are two cases which have been worked. The first case is for an idealized spectrum for $\underset{\sim}{S}_{FM}(f)$ sketched in Figure 3. When the noise spectrum does not extend to zero frequency, $L < \infty$ and the limiting form for the RF spectrum is ([7] Eq 14.42 b)

$$\underset{\sim}{S}_{RF}(\nu) = \frac{A_o}{2}\left[\delta(\nu - \nu_c) + \frac{1}{2}\,\frac{\underset{\sim}{S}_{FM}(\nu - \nu_c)}{(\nu - \nu_c)^2}\right] \quad , \tag{32}$$

which is valid only for

$$|\nu - \nu_c| < f_b \quad . \tag{33}$$

This is the kind of a result intuitively expected from using sinusoidal modulation theory for the small deviation case. Since this idealized spectrum, $\underset{\sim}{S}_{FM}$, does not go to zero frequency, the analogous sinusoidal modulation index $\Delta\nu/f_A$ does not go to ∞ and, therefore, presents intuitive trouble.

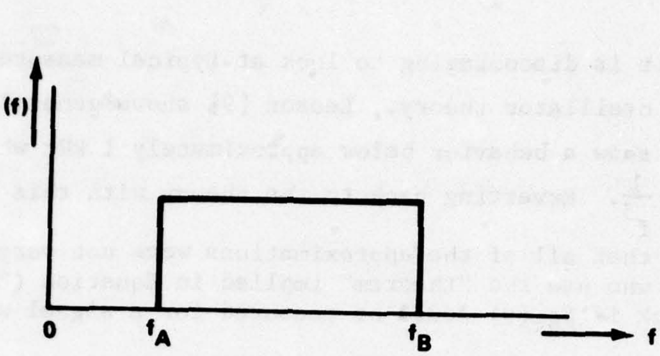

Figure 3. An idealized FM noise spectrum.

The real world case corresponds to the FM noise spectrum sketeched in Figure 4. This sketch is idealized because the oscillator should show a 1/f spectral shape near f = 0. If the FM noise spectrum is constant as $f \rightarrow 0$, then the integral L of Equation (31) diverges. Using a different mathematical route, Middleton ([7], Eq 14.45) obtains the RF spectrum as

$$S_{RF}(\nu) = \frac{\gamma A_o^2}{\gamma^2 + 4\pi^2 (\nu - \nu_c)^2} + \text{Correction Term} \quad . \tag{34}$$

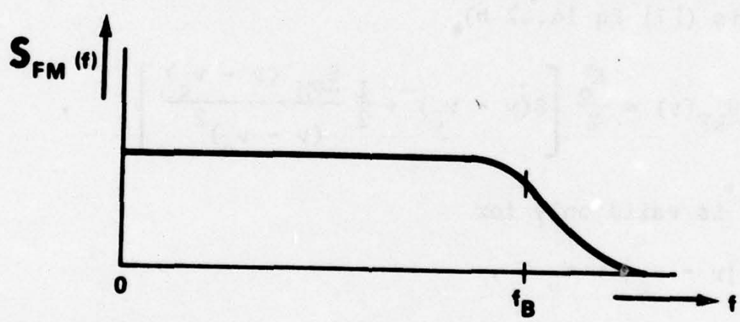

Figure 4. A more typical FM noise spectrum.

The "correction term" is significant only in the "tails" of the spectrum where

$$|\nu - \nu_c| > f_B \quad . \tag{35}$$

There is a surprising physical significance in this development. In the region

$$|\nu - \nu_c| < f_B \quad , \tag{36}$$

the shape of $\underset{\sim}{S}_{FM}(\nu)$ does not appear in the leading term for the spectrum. Also, there is no delta function in Equation (34), so a carrier for computing carrier to noise sideband ratios cannot be considered. Instead of the discrete carrier and sideband structure of AM or overly idealized angle modulation, the real world signal sources have a spectral shape resembling the resonant response of a series L-R-C circuit shown in Figure 5. The width of this spectral "line" is γ/π where γ is a measure of the total power in the FM noise spectrum; i.e., a factor proportional to the area under the $\underset{\sim}{S}_{FM}(f)$ curve. In practice, the line width is approximately a few hertz at the most and cannot be resolved with commercially available RF spectrum analyzers which might have a minimum bandwidth of 10 Hz. (Also, the noise floor in the spectrum analyzer is too high to see anything important.) *This* is why the failure of Equation (34) to match the intuitive extrapolation of sinusoidal FM theory has not been detected by measurement.

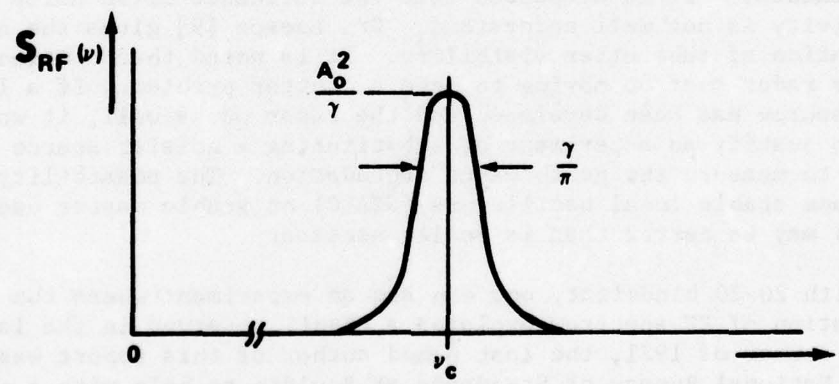

Figure 5. The limiting form for the spectrum with low FM noise of the shape shown in Figure 4.

Reflecting on what has been presented and comparing it with what is widely (and wrongly) written about the spectrum of a phase noise modulated signal, one will soon realize that a bomb has been dropped on the practicioners of conventional wisdom. All the plots of $\mathcal{L}(f)$ in the National Bureau of Standards reports, the spectrum plots in textbooks on radar, and many specifications of phase noise in terms of "dBC" are merely mathematical manipulation of the data; none are the actual RF spectrum. The problem with $\Delta\nu/f_m$ going to infinity as $f_m \rightarrow 0$ has been a cause for suspicion and this reverting to fundamentals (available since 1957[10]) has justified this suspicion.

It is important to note that one widely practiced data manipulation has not been destroyed. If the phase modulation spectrum, $\underset{\sim}{S}_{FM}(f)$, is measured, then the frequency modulation spectrum, $\underset{\sim}{S}_{\phi M}(f)$, can be derived using

$$\underset{\sim}{S}_{FM}(f) = \frac{1}{f^2} \underset{\sim}{S}_{\phi M}(f) \qquad (36a)$$

and vice versa. The proof of this is constructed by recalling the definition of instantaneous frequency, Equation (3), and using a well known theorem of communications theory quoted by Papoulis ([8] Table 10-1).

One might ask "why hasn't this shown up in radar sensitivity measurements?" It is suspected that the influence of FM noise on radar sensitivity is not well understood. Dr. Leeson [9] gives the usual explanation of subclutter visibility. It is noted that continuous wave Doppler radar must be moving to have a clutter problem. If a low noise source has been developed and the radar works well, it would be hard to justify an experiment of substituting a noisier source and trying to measure the performance degradation. The possibility exists that some stable local oscillators (STALO) or stable master oscillator (STAMO) may be better than is really necessary.

With 20-20 hindsight, one can see an experiment where the correct computation of RF spectrum explains a result observed in the laboratory. In the summer of 1971, the last named author of this report was called to the National Bureau of Standards at Boulder to help with a problem of detecting the beat between a 9.50817-GHz reflex klystron and various lasers [11]. Theory indicated that a beat should be detectable using 3- to 10-THz lasers.* Only the 0.964-THz line from a hydrogen-carbon-nitrogen (HCN) laser has been detected. The reflex klystron

*THz means Terahertz.

was stabilized with an HP 2650A synchronizer which electronically locks a reflex klystron 30 MHz from a harmonic of an approximately 100-MHz crystal-controlled oscillator. Figure 1 shows the measured FM noise from such a system. When these data were observed, it was not understood why the beats were not detectable. The upward bend at 10 kHz is beyond the 4 kHz half bandwidth of the receiving system. A decision was made to try a cavity-stabilized klystron which had the greatly reduced FM noise shown in Figure 2. It was quickly found that one could detect a beat with the 3.821774-THz water-vapor laser if one could twiddle the delta nu knob skillfully. This was fixed by using an injection lock [12] to a quartz crystal-controlled signal. Ashley and Palka presented a paper and wrote a patent application. No one knew why the cavity-stabilized oscillator worked better. It is obvious to the most casual observer that the area under Figure 6 is much less than the area under Figure 7, and that the resulting line width of the cavity-stabilized klystron is much less. Pursuing this problem resulted in useful research in measuring laser frequency and transmitter noise measurements. Risley, et al [13] document the noise data on all the sources used in this research and show the excellent agreement of measured results using the cavity discriminator, electronic phase-locked comparison oscillator, and computing counters (sigma versus tau or Allan variance data). The results of the work at the National Bureau of Standards are one of the bases for emphasizing the use of frequency *discriminators in making transmitter noise measurements.*

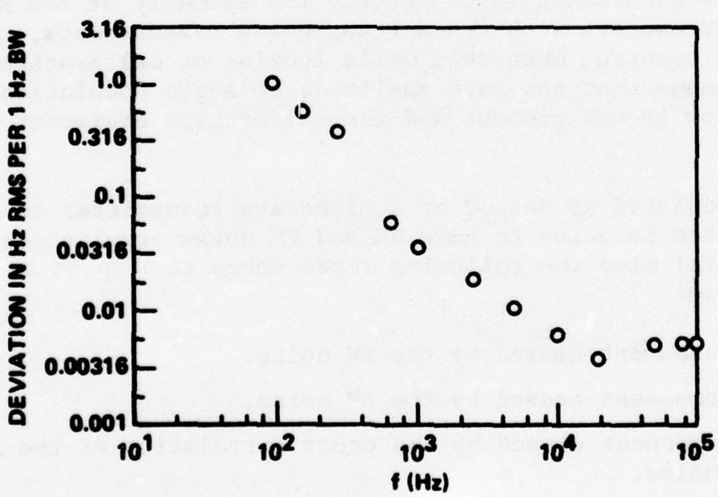

Figure 6. Measured FM noise of a cavity-stabilized X-band reflex klystron.

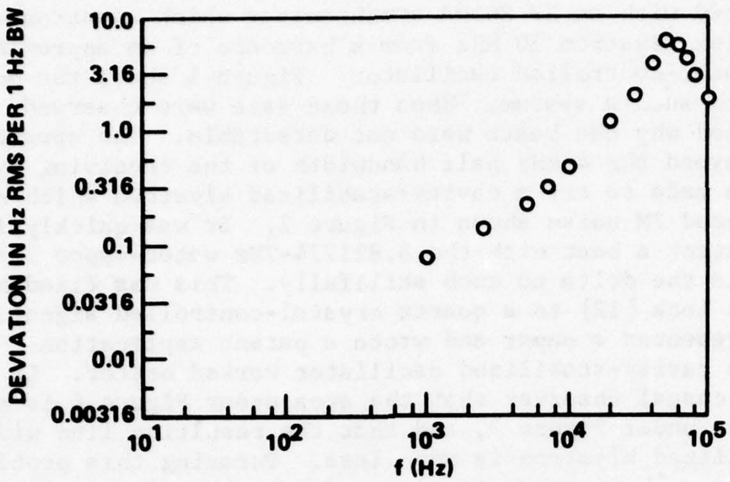

Figure 7. Measured FM noise of an X-band reflex klystron stabilized with the HP-2650 oscillator synchronizer.

The case when FM and AM are present simultaneously deserves brief comment. For the sinusoidal case, the result of simultaneous amplitude and angle modulation is to destroy the symmetry of the RF spectrum. For our concern with transmitter noise measurements, this assymmetrical spectrum when seen while looking at calibration side frequencies means that the pure amplitude or angle modulation needed for calibration is not present and the calibration equipment must be realigned.

The unmodulated RF output of a microwave transmitter or of any RF signal source is going to have AM and FM noise simultaneously present. This will give the following three terms to compute to find the RF spectrum:

a) A component caused by the FM noise.

b) A component caused by the AM noise.

c) A component caused by the cross correlation of the AM and FM noise.

For frequencies within a few kilohertz of the carrier region, the FM noise term dominates. For frequencies more than a few tens of kilohertz from the carrier region, the AM term dominates. The cross-correlation term would be found by applying the Wiener-Khintchine theorem to the cross-correlation function. Middleton [7] (Sec 14.3-1) shows that the computed effect is to skew the spectrum. Thus, the result is of diagnostic value for transmitters (such as countermeasures jammers) which use large index noise modulation to produce wideband noise as the primary output. If the spectrum is assymmetrical, then the modulation is not pure amplitude or angle. For the low residual level of the AM and FM noise in an unmodulated transmitter, a measurement of the cross-correlation function would be required before anything definitive could be said about the importance of this term in the RF spectrum computation.

What conclusions can be reached from this diversion into communications theory? In relationship to the topic of transmitter noise measurements, these conclusions must include the following:

a) The nature of noise in transmitters and the measurement equipment available dictate the measurement of AM noise and ΦM or FM noise.

b) The theory of amplitude or angle modulation by sinusoids gives the basis for alignment and calibration of noise measurement equipment.

c) A microwave spectrum analyzer is useful for detecting spurious outputs several megahertz from the carrier region in a transmitter, but cannot yield sufficient information about the noise modulation of the output.

d) A microwave spectrum analyzer is needed for alignment and calibration of AM and angle modulation noise measurement equipment.

e) Estimation of the contribution of AM noise to the RF spectrum is straigntforward [Equation (20)] and has no traps.

f) Estimation of the contribution of angle modulation noise to the RF spectrum is not straightforward [Equation (34)] and many people unknowingly fall into a serious trap [using Equation (32)].

g) FM or ΦM noise specifications based on a few numbers such as carrier to noise ratio at specified frequency(s) from the carrier are not sufficiently precise. A curve showing maximum limits of the angle modulation noise must be specified for completeness.

h) If one is asked to find the RF spectrum in the region near the carrier, one should ask why.

i) If there is a good reason for determining the spectrum in the region near the carrier, carefully measure AM noise and ΦM or FM noise and then compute. This paper does not present enough detail to make the computations; Middleton [7], however, does go into detail.

j) One must be careful about extrapolating theory based on sinusoids only to a situation where random noise is involved.

More will be presented on measuring transmitter noise in the following sections.

IV. MEASUREMENT OF AM NOISE

The first measurement to consider is determining $N_A(t)$ in Equation (5). In practice, the measurement yields the spectral description $\underset{\sim}{S}_{AM}(f)$. There has been no significant advance in this technique since the reports by Ashley, et al [3] and Ondria [4] in 1968. As reported by Ashley and Ondria, a direct detector diode is more simple than a superheterodyne demodulator and has a lower threshold. For measuring noise and determining the threshold or noise floor of the measurement, a double diode circuit such as the one shown in Figure 8 is convenient. The two detector mounts are insulated for dc and baseband frequencies up to several megahertz by plastic shims between the mounts and the main portion of the waveguide. This allows the diodes to be connected in series aiding or opposing as indicated in the low-frequency equivalent circuit of Figure 8. The dc blocks account for the two secondary windings on the equivalent circuit. The two capacitors labeled C are actually cable capacity and the resistors R are 1 to 10 kΩ each. At this impedance level, there has been no trouble with interference at the power line frequency either in or out of shielded rooms.

Silicon point contact diodes can be operated at approximately 1 V dc without exceeding the peak inverse voltage limit. Because this is a high-level detection application and the noise from the diodes sets the power limit, the threshold can be improved by increasing the power applied to the diodes. The higher peak inverse voltage that can be applied to a Schottky barrier diode makes it a superior device for AM measurements. (Note that detector mounts designed for point contact diodes will not function properly with Schottky barrier diodes because of the difference in V-I characteristics of the two kinds of diodes. This can be remedied over a limited but adequate microwave bandwidth either with a slide screw tuner or by inserting an adjustable capacitive post in the diode mount at a point $0.3\lambda_g$ in front of the diode.)

Depending on the type of detector used, a signal level of 1 to 5 dBm is applied to the detector input and the two mounts are tuned for maximum dc voltage output. Both detectors should be adjusted as closely as possible for identical operation. Data are then taken to determine the detector characteristics of Figure 9. The abscissa of this curve is obtained by normalizing the RF voltage with respect to the voltage at the operating point. The RF input to the detector is controlled by a precision waveguide attenuator. The ratio of the slope at the operating point to the slope of a line from the origin to the operating point gives a correction factor, C, which is used in Equation (37). The correction factor for the diode characteristic of Figure 9 amounts to 3.5 dB at an operating level of 2.45 V.

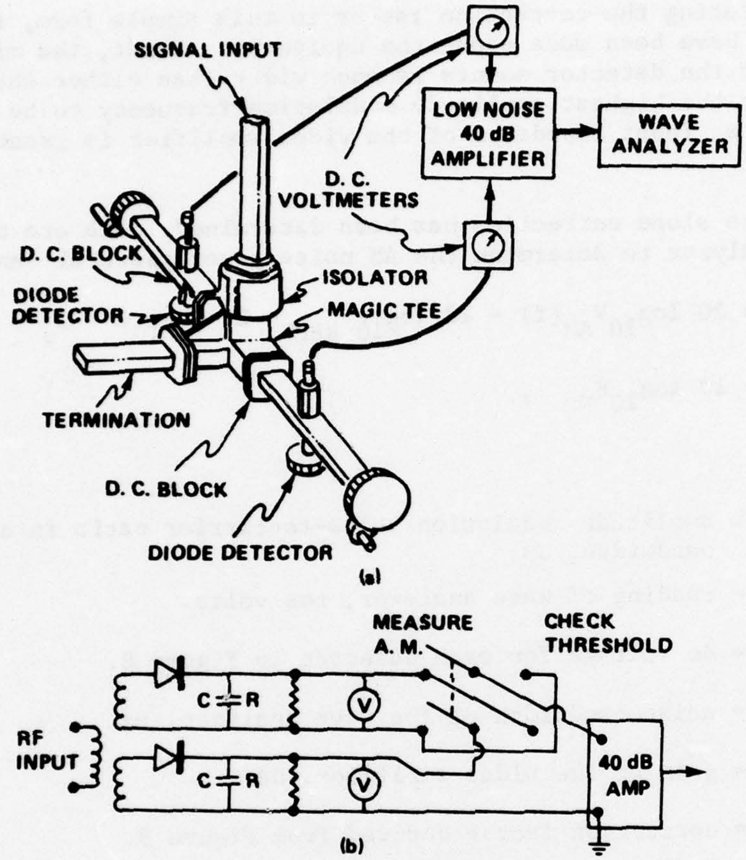

Figure 8. Balanced AM detector and low-frequency equivalent circuit.

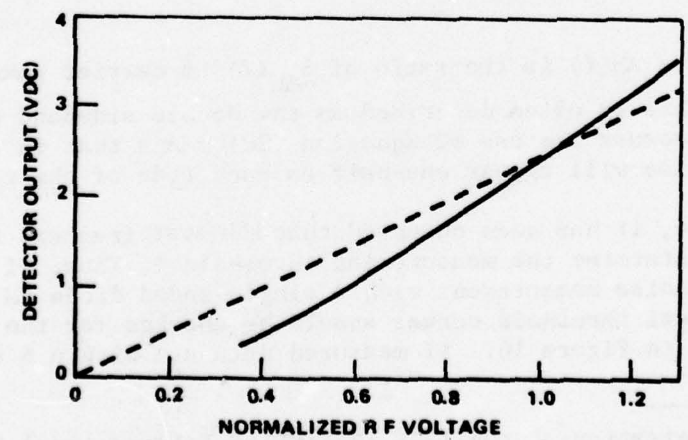

Figure 9. Schottky barrier diode detection characteristics (HPA2627 diodes at 10 GHz).

When stating the correction factor in this simple form, two assumptions have been made about the equipment. First, the microwave bandwidth of the detector mounts is much wider than either the maximum deviation or the highest amplitude modulation frequency to be measured. Second, the ac input impedance of the video amplifier is greater than 20 R.

Once the slope correction has been determined, data are taken from the wave analyzer to determine the AM noise power spectral density:

$$AM(f) = 20 \log_{10} V_{AM}(f) - 20 \log_{10} V_{REF} - 20 \log_{10} C - G_v - 6$$

$$- 10 \log_{10} B_N \quad , \tag{37}$$

where

AM(f) = amplitude modulation noise-to-carrier ratio in a 1-Hz bandwidth, dB.

$V_{AM}(f)$ = reading of wave analyzer, rms volts.

V_{REF} = dc voltage for each detector in Figure 8.

B_N = noise bandwidth of the wave analyzer, Hz.

G_v = gain of the video amplifier, dB.

C = correction factor derived from Figure 9.

f = modulation frequency, read from wave analyzer, Hz.

The 6-dB factor is used to account for the two detectors operating in series. If a single diode detector is used, this 6-dB factor is omitted.

The quantity AM(f) is the ratio of $\underset{\sim}{S}_{AM}(f)$ to carrier power expressed in decibels. This is often described as the double sideband noise to carrier ratio because the use of Equation (20) means that in the RF spectrum this AM noise will appear one-half on each side of the carrier.

In practice, it has been observed that the most frequent failing is neglecting to determine the measurement threshold.* Thus, if one attempts an AM noise measurement with a single-ended diode AM demodulator, the typical threshold curves should be checked for the demodulator device as shown in Figure 10. If measured data are within 6 dB of the

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*In a conversation at the 1976 IEEE-MTT-S International Symposium, Dr. John Ondria made a similar observation.

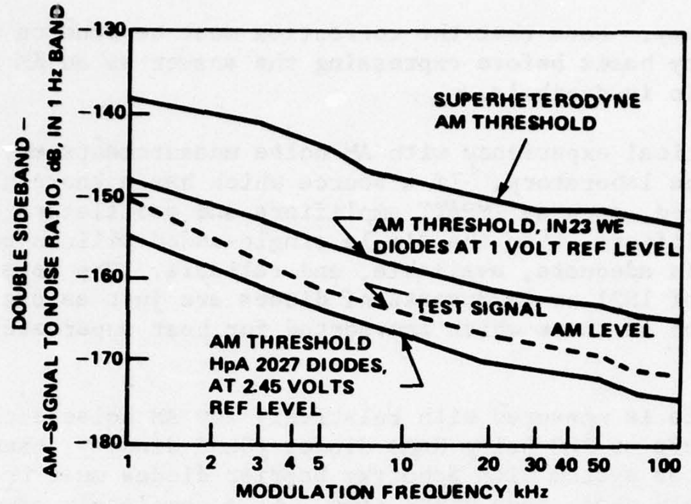

Figure 10. The AM threshold for a superheterodyne receiver, silicon point contact diodes (IN23WE), and Schottky barrier diodes (HpA 2627).

expected threshold, then a balanced detector such as the one illustrated for waveguide in Figure 8 needs to be constructed. The test for the threshold shown in Figure 8(b) consists of operating the individual diodes in series opposition so that the AM on the test signal is mostly cancelled. The balance of the diodes is very important and must be tested under RF excitation. One easy method is to apply intentional sinusoidal AM on the test signal and match the detectors to yield minimum output at the AM frequency. The first intuition would be to balance the detectors by tuning adjustment. This tends to cause unbalance for modulation frequencies above 100 kHz. A better way is to adjust the sizes of the load resistors R to balance the detector system. A plot of noise spectrum from the detector is a good indication of threshold. Reconnecting the diodes to series aiding then gives an output which is the sum of threshold noise and AM noise. If the output is within 2 to 10 dB of threshold, then a correction

$$V_{AM}(f) = \sqrt{V^2_{AMM}(f) - V^2_{ATH}(f)} \quad , \tag{38}$$

where

$V_{AM}(f)$ = corrected AM noise

$V_{AMM}(f)$ = uncorrected AM noise data

$V_{ATH}(f)$ = threshold data

improves accuracy. Note that the correction must be done on a power spectral density basis before expressing the answer as an AM noise to carrier ratio in decibels.

Some practical experience with AM noise measurements may help save time in the laboratory. If a source which has a known high AM noise is measured, such as IMPATT amplifiers and oscillators or transmitter TWT amplifiers, then the simple single-ended silicon point contact diode is adequate, available, and reliable. The least expensive versions of 1N21 or 1N23 sorts of diodes are just as useful in this application as those which are sorted for best superheterodyne noise figure.

If a source is measured with relatively low AM noise such as klystron oscillators or TEO using GaAs diodes (Gunn diodes), then the full balanced detector system with Schottky barrier diodes must be employed and the threshold check made every time. If a completely new and unknown source is measured, then a careful and complete calibration and determination of measurement threshold must be done.

A degradation of Schottky barrier diodes which is not understood has been experienced.* The peak inverse voltage allowable decreases with operating life and some strange noises appear in the demodulator output. This requires replacement of the diodes before taking more data.

A simple operational check for an adequate measurement threshold is worth knowing. In the calibration procedure, a precision signal frequency attenuator is needed to take the slope data. This attenuator is left in the system for the balance of the measurement. Regardless of the kind of detection system, increasing attenuation on the input of the demodulator by 3 dB should make the wave analyzer output indicator decrease by the same 3 dB. If there is less than a decibel of change on the wave analyzer output, the system threshold is inadequate. If there is a 2-dB change, taking the threshold data and correcting it will give a reliable answer.

One method of reducing AM threshold [4,5] has been thoroughly evaluated by Fikart, et al. [14,15,16] and found to be of dubious use. Essentially, the cavity discriminator [3,4,5] can be used for AM noise measurements by inserting a 90-degree phase shift in the reference line going to the output phase detector. Component imperfections and drift of the source and or the discriminator cavity tuning cause sufficient errors to make this a method of last resort. Since only the two-resonator klystron oscillators have approached the threshold of the dual diode system with Schottky barrier diodes, the method of last resort is not often needed.

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*In a conversation with Dr. John Ondria at the 1976 IEEE-MTT-S International symposium, he reported similar experiences. The diode manufacturers are aware of the problem, and diodes recently produced are better in this regard.

When measuring an amplifier or frequency multiplier, AM noise at input and output is measured separately. Because these components usually operate in a highly saturated mode, there is little correlation between input and output AM noise. Furthermore, the AM noise at the output is quite dependent on relative carrier power output. A tendency to neglect making the AM noise measurement, which is not justified by transmitter device performance, has been observed.

V. MICROWAVE FM NOISE MEASUREMENTS

The historical development of accepted noise measurement circuits will be followed by describing FM or phase noise measurements. The discussion in this section will concentrate on the microwave region above 5 GHz. The following section will discuss FM noise measurements for carrier frequencies below 5 GHz.

The history is difficult to document because much of the work was classified for security reasons in the era following World War II and during the Korean War. This work has now been declassified, but the archival journal papers were not written at the time, and now one would have to study company and military reports to adequately give credit to those who made the first contributions. With regard to equipments available in the USA, one can trace it back to the Royal Signals and Radar Establishment (RSRE), Great Malvern, Worchester, England. During the 1950's the RSRE supported the development of klystron oscillators and amplifiers for Doppler radar use. These klystrons quickly presented a measurement challenge, and the apparent result was the development of the cavity discriminator which Whitwell and Williams [17] attribute to "S. B. Marsh, J. D. Clare, and S. A. Drage of RSRE who evolved the frequency discriminator and constructed the first experimental models,..." This equipment was marketed in the US by James Scott & Co., Ltd., Glasgow, Scotland, as the Allscott Model 123 Noise Measuring Equipment.*

Although it did represent a completely designed equipment with a set of operating instructions, the RSRE discriminator bridge had two superiorities over the prior art: first, it would reject a large amount of incidental AM noise on the test signal; and, second, input power could be increased (with proper adjustment of the equipment) to override the threshold noise in the equipment. Thus, the 10-W type of two-resonator klystron oscillator, which still holds the world record for low FM noise, could be measured.

Users of the RSRE cavity discriminator quickly tire of tuning the microwave bridge based on a magic tee or 3-dB four-port directional coupler, as well as tuning the klystron used as a local oscillator for the superheterodyne equipment. Many microwave sources put out 10 mW

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*Rast made his first believable FM noise measurement with Allscott equipment in the early 1960's.

or more, and the input power sensitivity is not needed. Also, the 3-dB hybrid in the bridge costs a total of 6 dB in threshold or detection slope because the signal goes through the hybrid twice. These factors and the need to measure lower noise sources first motivated Ondria [4] and then, Ashley et al. [3] (working independently) to make significant improvements in this equipment. As shown in Figure 11, a circulator has replaced the magic tee and the cavity has been modified to include an adjusting screw on the coupling iris. This simplifies tuning and improves the threshold by nearly 6 dB. Because the intermediate frequency amplification is not often needed, the local oscillator and IF equipment has been left out in the interest of simplicity. Although the final phase detector could be the magic tee circuit of Figure 8, it was found more convenient to use a standard balanced mixer* in the discriminator as illustrated in Figure 12.

There are two approaches to deriving a theory for demodulator circuits such as this frequency discriminator: (a) the quasi-stationary frequency assumption [3] or (b) the sideband phasor method [4]. The quasi-stationary frequency assumption has the advantage of yielding a direct answer for the discriminator slope and showing that the answer is good for modulation frequency approaching zero; thus, one can use the discriminator for short-term and even medium-term stability indications for comparison with other measurement techniques. The sideband phasor method yields the upper modulation frequency correction factors more directly. The first method is the one used in communication theory study of noise problems in FM systems [7]. Either method will yield an equation for the slope, S, of the form

$$S = \frac{\Delta v}{\Delta \nu} = \frac{K\sqrt{P_i}\ 10^{-(G_H/20)}\ Q_o}{\nu_c}, \tag{39}$$

where

$\Delta\nu$ = small deviation from the carrier frequency, pk Hz

Δv = small dc change in output, V

P_i = input power, W

G_H = loss in the system between input and phase detector, dB

Q_o = internal Q of the microwave resonator

ν_c = carrier frequency, Hz

K = a factor depending mainly on the conversion loss in the phase detector.

*Such as a Varian Orthomode mixer designed as a down-converter for superheterodyne receivers.

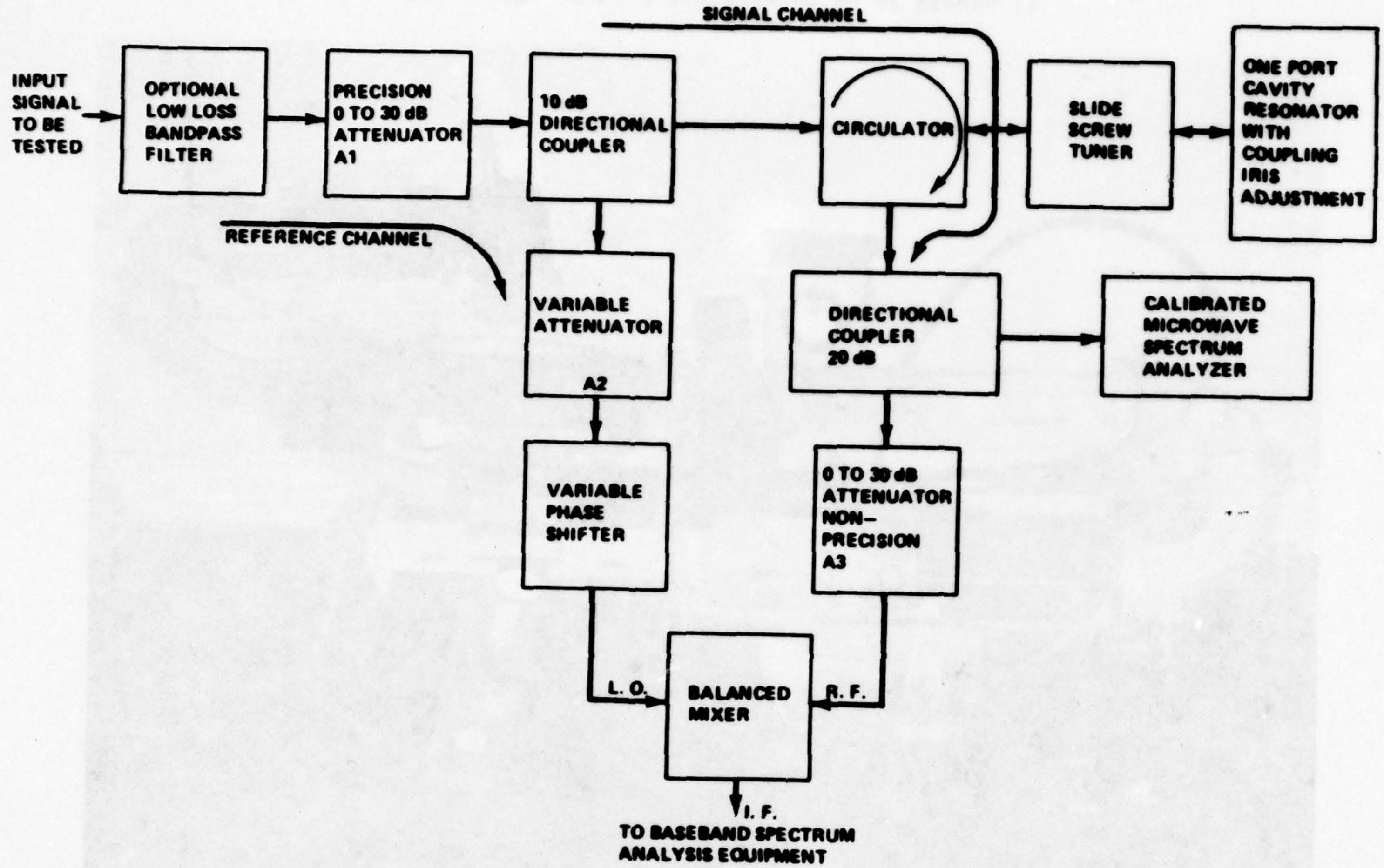

Figure 11. A microwave frequency discriminator for FM noise measurement.

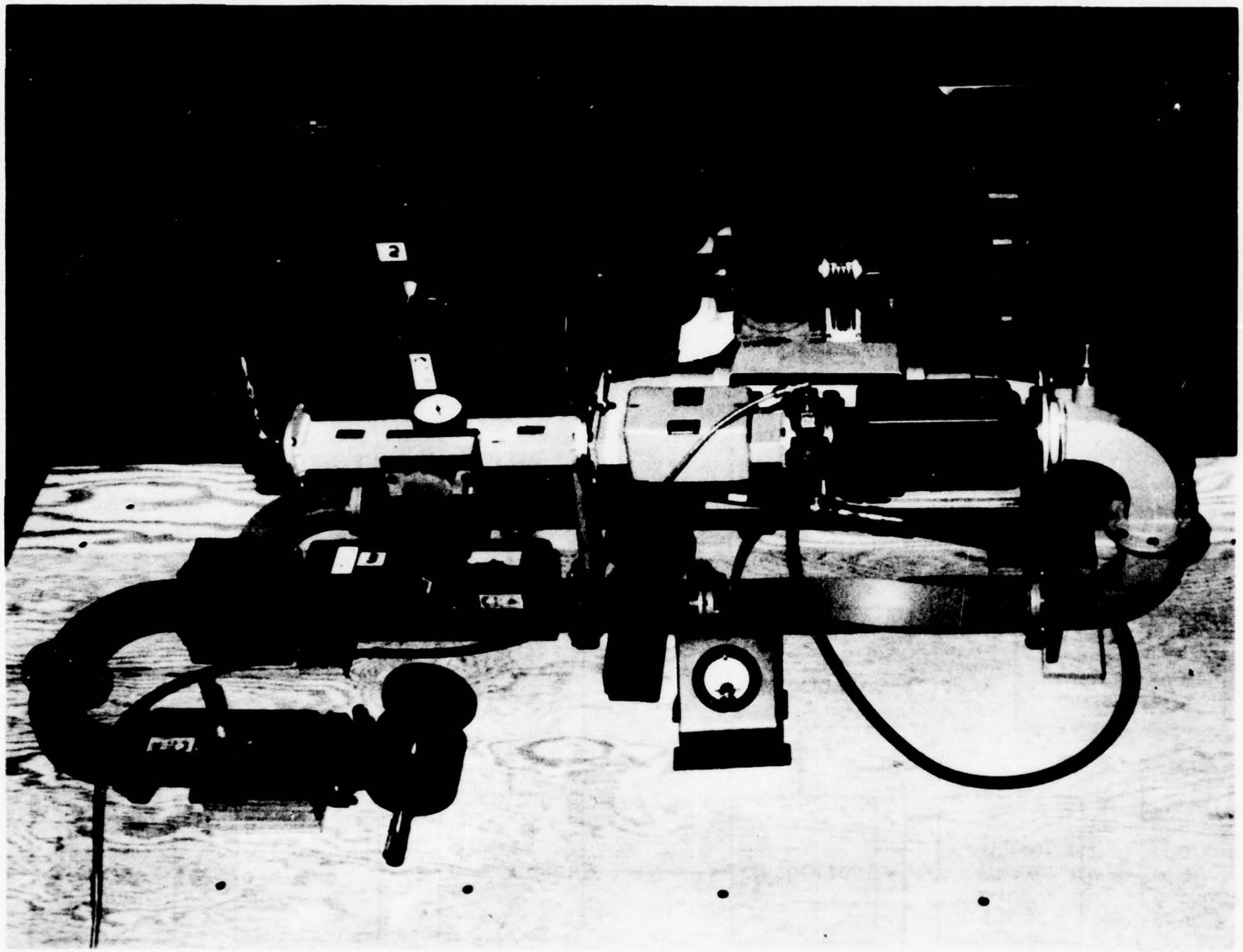

Figure 12. C-band implementation of Figure 11.

This equation is principally for study rather than for actual calculation. The measurement of the factor K has some conceptual difficulties. It is not necessary to know that measurement of a resonator Q_o, which is probably greater than 20,000, is a challenging problem. Thus, most laboratories use a technique based on applying a known angle modulation and measuring Δv to determine S. Such a technique will be described with the tuning and operating procedure. Presently, it is important to draw some conclusions from Equation (39) that influence operation.

First, note that the slope is proportional to $\sqrt{P_i}$. Therefore, if a calibration is performed to determine S, this result is valid only as long as P_i is not changed. All the other factors of Equation (39) are quite stable, and as long as the power is maintained constant, the slope is very stable with time. Therefore, it is reasonable to use one source (which might be easy to angle modulate) for calibration to measure another source; either hold the input power constant or measure the power ratio and make the appropriate correction.

Second, measurement of a small deviation, $\Delta\nu$, (corresponding to a quiet oscillator) requires the highest obtainable value of S to override the noise in the crystals of the phase detector. Because carrier frequency, ν_c, and K are fixed, P_i must be as high as possible, Q_o high, and G_H minimized. If the circulator shown in Figure 11 is used, G_H is approximately 1 dB; whereas the use of a magic tee or 3-dB coupler will make G_H approximately 7 dB. This 6-dB difference directly translates to a 6-dB lower measurement threshold. If these measures do not result in an adequately low threshold, then methods discussed in Section VII must be employed.

The determination of S is performed as part of the tuning and operating procedure. This will be the organization of this discussion. Notice that the evolution of signal sources and spectrum analyzers has caused appreciable change in calibration procedure from those reported earlier [3,4]. In the 1950's and early 1960's, most of the sources being measured were direct microwave oscillators (klystrons, TEO, IMPATT, etc.) which could be easily frequency-modulated via a bias port. Also, the deviation in these devices could be proved from theory and by measurement to be linear with respect to voltage applied. Microwave spectrum analyzers of that era were not stable and calibration accuracy was low; thus, the spectrum analyzer was used to determine the first carrier null and the corresponding modulation index, 2.405. Then the modulating voltage was attenuated by a measured amount and the linearity of the oscillator depended on to calculate the amount of angle modulation used in the calibration. A good fraction of sources to be tested are controlled by VHF or UHF crystal oscillators and usually do not have angle-modulation facility. Also, microwave spectrum analyzers have experienced continuous improvement. Several manufacturers offer equipment with excellent stability and reliable calibration. In particular, the intermediate frequency attenuator is a stable device that can be accurately calibrated and used as a secondary standard in making the slope calibration. Dynamic range and internal noise in the spectrum analyzer make it possible to measure calibration sidebands 50 to 60 dB below the carrier with adequate accuracy.

For the discussion of the discriminator tuning, calibration, and operation, assume that angle modulation side frequency can be applied to the source being measured 50 to 60 dB down from the carrier. If the source has internal phase or FM capability, then the microwave spectrum analyzer is used to measure the side frequency to carrier ratio. If the source cannot be modulated, then a circuit such as the one shown in Figure 13 can be connected between source and the discriminator. The idea is simple, i.e., AM side-frequencies with suppressed carrier are generated with a crystal modulator and reinserted with the needed 90-deg phase shift to appear as angle modulation side-frequencies. The difficulty is to know that the needed 90-deg phase shift has been obtained and if carrier leakage through the balanced modulator will confuse the calibration. The amplitude detector indicates that the phase shifter is set correctly because a simple phasor diagram indicates that minimizing AM on the recombined signal is an accurate indication of proper side-frequency phasing, even if the balanced modulator has carrier leakage of the order of the side-frequencies generated (as has proved the case in modulators we have tested).

During the process of setting up and measuring the calibration side frequencies, the precision input attenuator of Figure 11 should be set for maximum attenuation to protect the crystals in the balanced mixer of the discriminator.

The operation procedure for the circuit of Figure 13 is as follows:

a) Apply as much signal frequency power to the input of the balanced modulator as the device will stand without burnout.

b) Set the audio signal generator at the desired side-frequency spacing (e.g., 10 kHz) and adjust the output level while viewing the signal frequency spectrum analyzer. Do not exceed the level which causes second-order side-frequencies (e.g., ±20 kHz) to appear.

c) Tune the audio spectrum analyzer to the frequency of the audio signal generator (e.g. 10 kHz), then adjust the variable phase shifter to minimize the indication on the audio spectrum analyzer.

d) Check the level of the angle modulation side-frequencies on the calibrated spectrum analyzer.

In operating the circuit, be careful not to burn out the crystals in the balanced modulator with too much RF input or to cause generation of higher order sidebands by the application of excessive audio input power. Usually, it is possible to obtain calibration side-frequencies in the range 50 to 60 dB below the carrier. Side-frequencies more than 60 dB below the carrier are too close to the internal noise level of the microwave spectrum analyzer to measure accurately for calibration purposes. Side-frequencies less than 50 dB below the carrier requires a modification of Equation (40).

One idea which might occur is to use this calibration technique to do a swept calibration of the output indicator; e.g., putting a line of constant side-frequency level on the output graph. This idea has been tried with reasonable success. For side-frequency spacing greater than 5 kHz, the microwave spectrum analyzer can resolve the sideband structure and prove that the calibration side-frequency amplitude does remain constant. On several balanced modulators used in Figure 13, the calibration level line did not behave as expected below 5 kHz. The trouble was isolated to the balanced modulator. Perhaps a thermal time constant or an unsuspected filter inside the balanced modulator is the cause of the trouble. It is advised not to assume the side-frequency level is constant if one cannot prove it with the microwave spectrum analyzer.

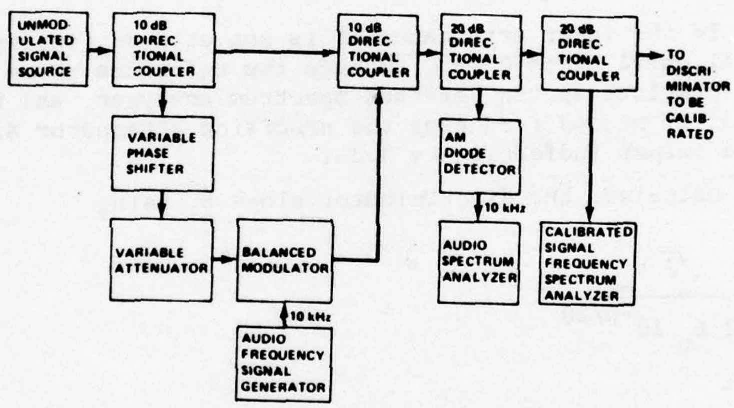

Figure 13. Generation of angle-modulated side-frequencies for discriminator calibration.

With the side-frequency level set and measured, the discriminator of Figure 11 is tuned and calibrated as follows:

a) With the reference channel attenuator (A2) set at 0 dB and the threshold check attenuator (A3) set at maximum, decrease the input attenuator (A1) until the reference power at the local oscillator (LO) input of the balanced mixer is as high as the mixer will stand without burnout. (A crystal current indicator is a useful way to make this determination.)

b) Replace the cavity resonator with a well matched termination and adjust the slide screw tuner for minimum carrier indication on the microwave spectrum analyzer. This cancels undesired leakage in the three-port circulator.

c) Reinstall the cavity resonator and tune to resonance. Check the mode chart and cavity tuning calibration to make certain that the desired mode is being used.* The indication of resonance is a dip in carrier level on the microwave spectrum analyzer. Adjust the coupling iris to minimize carrier amplitude, recheck the tuning, etc. Within a few iterations, the carrier level can be reduced 60 dB, which is more than sufficient. A more important goal to attain at this stage is to make certain that the resonator is tuned as precisely as possible to carrier frequency. With the calibration side-frequencies present, this is simple to determine because the side-frequencies amplitudes will be the same (the spectrum is symmetrical) if the resonator is properly tuned. At this stage, it is normal to see the side-frequencies' amplitudes within 10 dB of the residual carrier.

d) Reduce attenuator A3 to zero while observing mixer current which should not change. Tune the baseband spectrum analyzer to the modulation frequency of the calibration side-frequencies (e.g., 10 kHz) and adjust the reference channel phase shifter for maximum indication on the baseband spectrum analyzer. (Watch for spectrum analyzer overload.)

e) If the input attenuator A1 is not at 0 dB, increase A2, then decrease A1 until A1 is 0 dB. Recheck the reference phase adjustment. Check for overload in the baseband spectrum analyzer, and insert input attenuation if needed (reducing the precision attenuator A1 3 dB should reduce the output indication by 3 dB.

f) Calculate the discriminator slope S, using

$$S = \frac{\sqrt{2}\, V_{AC}}{2\, f_m\, 10^{-D/20}} \,, \tag{40}$$

where

V_{ac} = discriminator AC output voltage at f_m

f_m = modulation frequency, Hz

D = side-frequency amplitude with respect to carrier frequency amplitude, dB

and store this value for later use in data processing.

g) Insert the threshold check attenuator A3, turn off the calibration side-frequency, and operate the wave or spectrum analyzer at all frequencies where FM noise data are desired. (The input attenuator of

*These cavities using the TE_{01n} modes are an order of magnitude more difficult to protect from spurious mode troubles than the TE_{111} mode used in cavity wavemeters.

the wave or spectrum analyzer should be set for minimum at this step. An overload light on the input amplifier of the wave analyzer is most useful.) Record or store these threshold voltages, $V_{FTH}(f)$, and the measurement bandwidth, B_N.

h) Reduce the threshold check attenuator A3 to 0 dB, check overload of the baseband equipment, and take FM noise data at the desired baseband frequencies, $V_{FMM}(f)$. [Use the same values of B_N as in step g)].

i) Calculate corrected deviation at each modulation frequency using:

$$V_{FM}(f) = S\sqrt{V_{FMM}^2(f) - V_{FTH}^2(f)} \quad . \tag{41}$$

Note that the bandwidth for the measurement, B_N, was set in the baseband wave analyzer and must be stated along with the values for the FM noise.

j) Convert the FM noise data to other bandwidths, forms, etc.

There are several items that may be helpful when working on ones first or second FM noise measurements. As in the case of AM measurements, it was observed that the most consistent failure was neglecting determining measuring system threshold. The development of threshold checking routines is considered the most important contribution to the art of noise measurement. If one is attempting to measure a state-of-the-art source with only a few milliwatts output, it will probably be found that the threshold is too high (suggestions for this problem can be found in Section VII).

The usual problem is finding a cavity suitable for this use. Cavities intended for use as wavemeters do not have a sufficiently large coupling iris to achieve the critical coupling needed. Modification is usually not a trivial task. To achieve the high Q_o needed, the cavity mode is normally from the TE_{01n} family. The TE_{011} will have a $Q_o \approx 22{,}000$ at 10 GHz with a 10% tuning range, which is moderately free of spurious modes. As the higher order modes are used to obtain higher Q, the spurious mode problem becomes more serious. The vendors' most familiar with this kind of cavity resonator are those who develop transmission cavities for oscillator stabilization [18].

Most of the mysteries are found in the baseband equipment. All the good wave and spectrum analyzers have input and intermediate frequency attenuators to allow optimization of the level at the first up-converter. These attenuators must be set correctly [as noted in Steps V.d), V.e.), V.f.), and V.g.)]. Some older wave analyzers which were not intended for noise spectrum analysis do not have enough gain in the input stages and a preamplifier is needed.

Another baseband problem is ground loops. When the spectrum analyzer indicates at 60 Hz and harmonics thereof, the question is whether the indication means trouble in the signal source or the measurement apparatus. The threshold check partially indicates if the indication is in the measurement apparatus because FM from the source will be attenuated by the amount of the threshold check attenuation inserted. Eliminating ground loops seems to be more a matter of experience than anything that is amendable to recording. One source of trouble is the safety ground in the power cords of electronic instruments. All the baseband equipment should be plugged into one box of outlets and the cords run as a bundle as far as possible. The RF input to the final amplitude or phase detector is best supplied through dc blocks to break up the worst ground loop. Remember that it is the magnetic field that couples into a loop; thus, minimizing loop area and placing low frequency breaks in the loops will minimize current flow.

Minimizing external interference with the measurements is sometimes accomplished by using a shielded room. Experience has been that this is not as effective as one might think. In general, data have been taken in research and development laboratory environments and no significant difference found if the whole apparatus is moved inside a shielded room. In fact, a degradation has sometimes been seen because the shielded rooms often have ventilation blowers with appreciable acoustic noise output. This acoustic noise vibrates waveguide walls and cavity resonators (as well as the signal source) and causes a messy spectrum below 2 kHz. The threshold check does not distinguish whether this is in the signal source or in the measurement apparatus. If a messy spectrum in the FM noise below 2 kHz is seen, turn off all possible blowers and acoustic noise generators to estimate how much of the mess is acoustic. Unfortunately, shielded rooms get hot a few minutes after the blower is turned off. This acoustic noise must be tolerated. This experience shows that the only instance where outside interference could not be eliminated without a shielded room was in trying a measurement on the production floor near a test station for high-power pulsed klystrons. The magnetic radiation from the pulse transformers cluttered the noise measurement on a CW klystron at a nearby test station.

There does seem to be a learning curve associated with making transmitter noise measurements. Approximately a man-week is required to understand the specification and concept of transmitter noise. After equipment is assembled, another man-week is required to learn the ritual of all the tuning, calibrating, and data taking that must be done. At the end of this second week, one can expect to have some faith in eventually having adequate confidence in the measurements. At the end of 3- to 4-man-weeks, one can expect enough confidence to tell one's superior that the answers are right and that the art of transmitter noise measurements has been mastered. Results such as the typical measurement of Figure 14 are a pleasure to report.

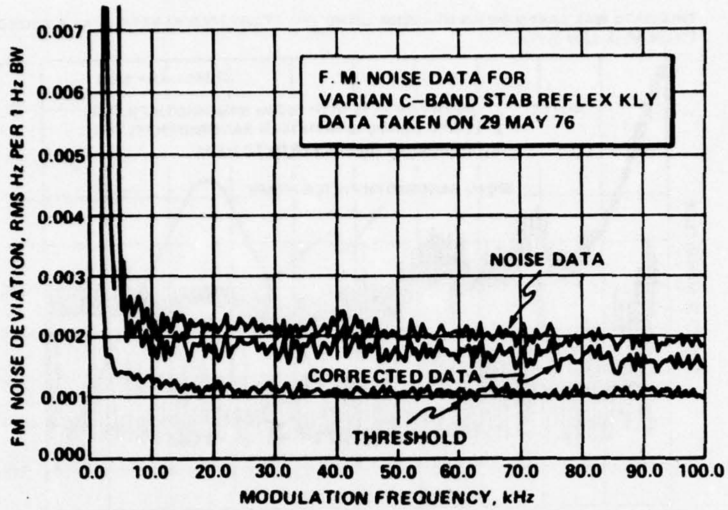

Figure 14. Typical C-band klystron data plot using the HP-3045A automated spectrum analysis system.

The use of a discriminator is logical for measuring an oscillator output and has been found equally useful for studying amplifiers.

Using a frequency discriminator for the measurement of amplifier phase noise might seem inappropriate at first glance. A phase bridge has been used to compare output and input of amplifiers to determine the noise added by the amplifiers [19]. There are several problems with this latter idea. First, an adequately low noise drive source must be used. The simple phase bridge will not determine adequacy. Second, the threshold will be inadequate if carrier suppression is not added. Third, after the measurement results are available, the amplifier noise and driving signal noise have to be combined to find output noise.

The frequency discriminator solves all of these problems. Since it is needed for the study of drive source noise, the discriminator is available for studying an appropriate sample of the amplifier output.

The measurement of AM and FM noise at the output of a 10-W or higher power TWT or klystron will normally show power supply ripple and grounding problems as shown in Figure 15. It has been reported [20] that the output noise cannot be predicted on the basis of noise figure measurements made at low levels and using an IF amplifier at 30 MHz. In the range from 3 dB below to saturated power output, the output noise significantly increases, probably because of RF defocusing of the electron beam and interception on the RF structure.

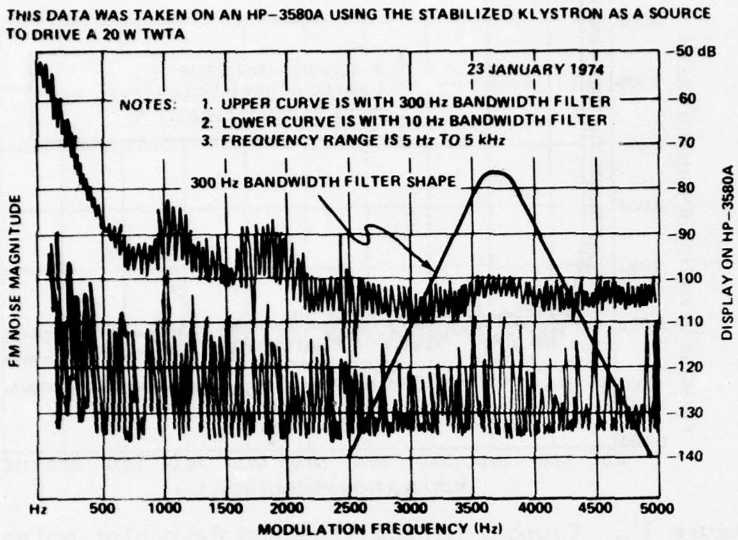

Figure 15. Power supply ripple and gounding problems.

The complete noise study of a microwave transmitter requires measurement of AM and FM noise at all significant intermediate and output ports. The cavity resonator discriminator discussed will make the FM noise measurement at the output ports, but is not practical for some of the intermediate stages. Section VI presents a description of a newly developed technique that is practical for the submicrowave portions of a transmitter.

VI. SUBMICROWAVE FM NOISE MEASUREMENTS

The need to make transmitter noise measurements in the lower microwave frequency range, VHF, and UHF ranges has grown because of the evolution of frequency synthesis techniques for microwave transmitters. When a microwave transmitter noise measurement of a synthesized signal yields an unacceptable noise, then one must go back and measure the components used in the synthesis. This was shown to be the case with the radio frequency simulation system developed and installed at the McMorrow Laboratories of the US Army Missile Command, Redstone Arsenal, Alabama. It was necessary to make many of the noise measurements below 2 GHz.

The method recommended by Shoaf, et al. [21] for this frequency range was to electronically phase lock the oscillator under test to a second identical unit. A spectral analysis of the correction voltage in the phase lock loop is the basis for determinnng the phase noise. To achieve a minimal proof that the oscillators are identical (and that the measured noise can be attributed to each oscillator on a 50-50 basis), a third oscillator is needed so that the three oscillators can be compared two at a time. The need to develop the equipment for the electronic phase locking is a strong disadvantage of this method. A method similar to the cavity discriminator methods described in the previous section is needed. As frequency gets lower, the TE_{011} cavities get larger, and somewhere below 2 GHz of unreasonable size and weight.* Such a new method must share the following advantages with the RSRE cavity discriminator if the method is to attain bench-mark status:

a) It must be relatively easy to adjust and calibrate.

b) It must be adequately stable in frequency and time.

c) An operational method for determining the measurement threshold must be included.

d) If additional signal power is available, the threshold must be reducible.

e) When properly tuned, the discriminator must reject incidental AM noise on the signal under test.

This need for a new method led to the study of transmission line discriminators. These have been known for a long time, but the conventional wisdom for dismissing them was well expressed by Campbell [22] in 1964. "The simplest detector -- with the exception of the slope detector -- is the interferometer, where the dispersive element is a long line. This is usually used for wide-band applications, with its accompanying low sensitivity. More sensitive versions require an excessively long line and are characterized by multiple responses." The more important faults of prior art transmission line discriminators were that additional input signal could not be used to reduce threshold and that AM on the test signal was not rejected.

After all the years of operating the microwave discriminator of Figure 11, it suddenly became obvious that replacing the cavity with a long, shorted transmission line and slide screw tuner as shown in Figure 16 gives a discriminator which meets all of the requirements previously listed. In the VHF range, circulators and slide screw tuners may be hard to obtain while 3-dB four-port hybrid junctions are readily available; thus, the adaptation shown in Figure 17 is useful, although the threshold and slope are 6 dB poorer than the circulator version.

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*We have not been able to purchase cavities for the region below 5 GHz.

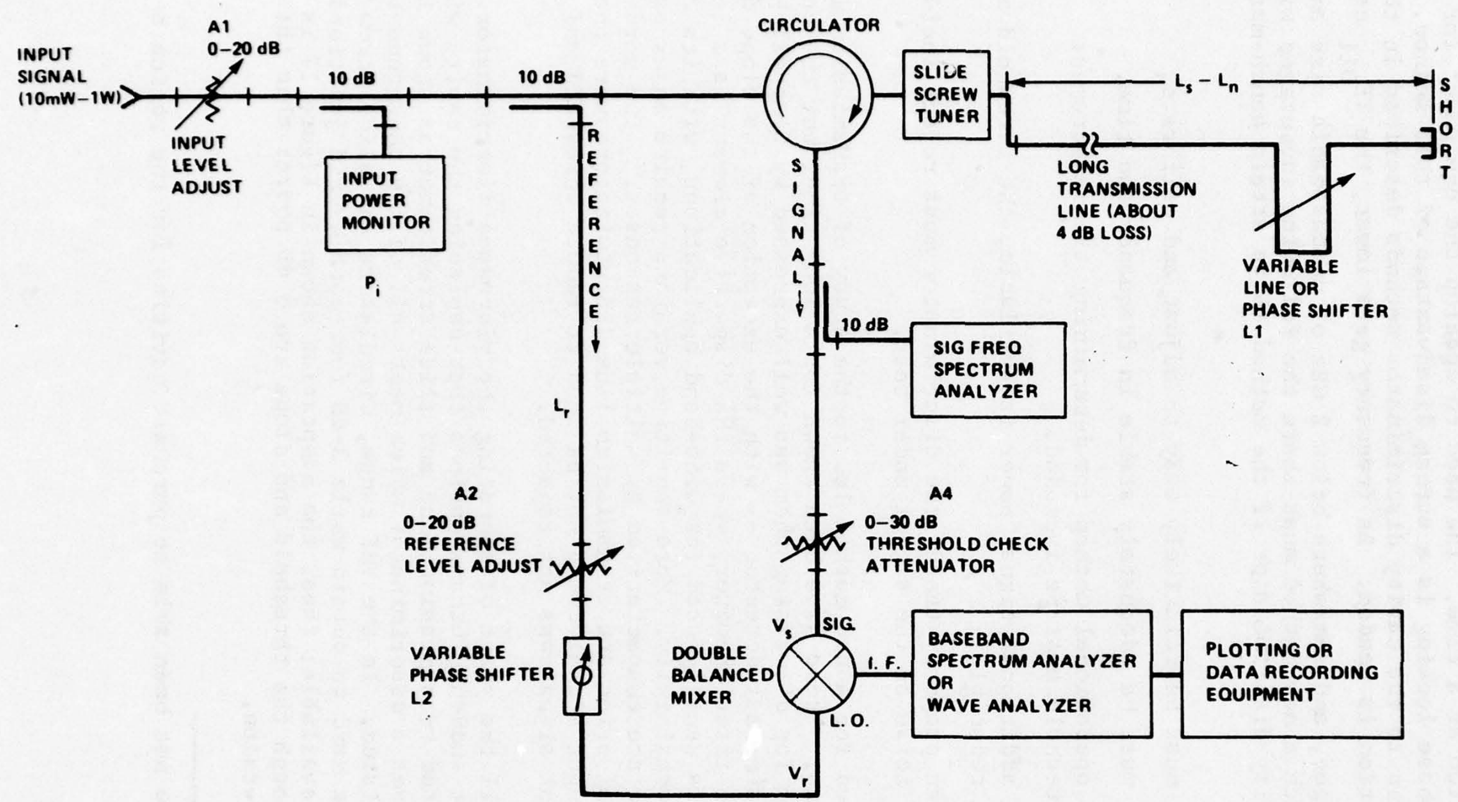

Figure 16. An optimum sensitivity transmission line discriminator.

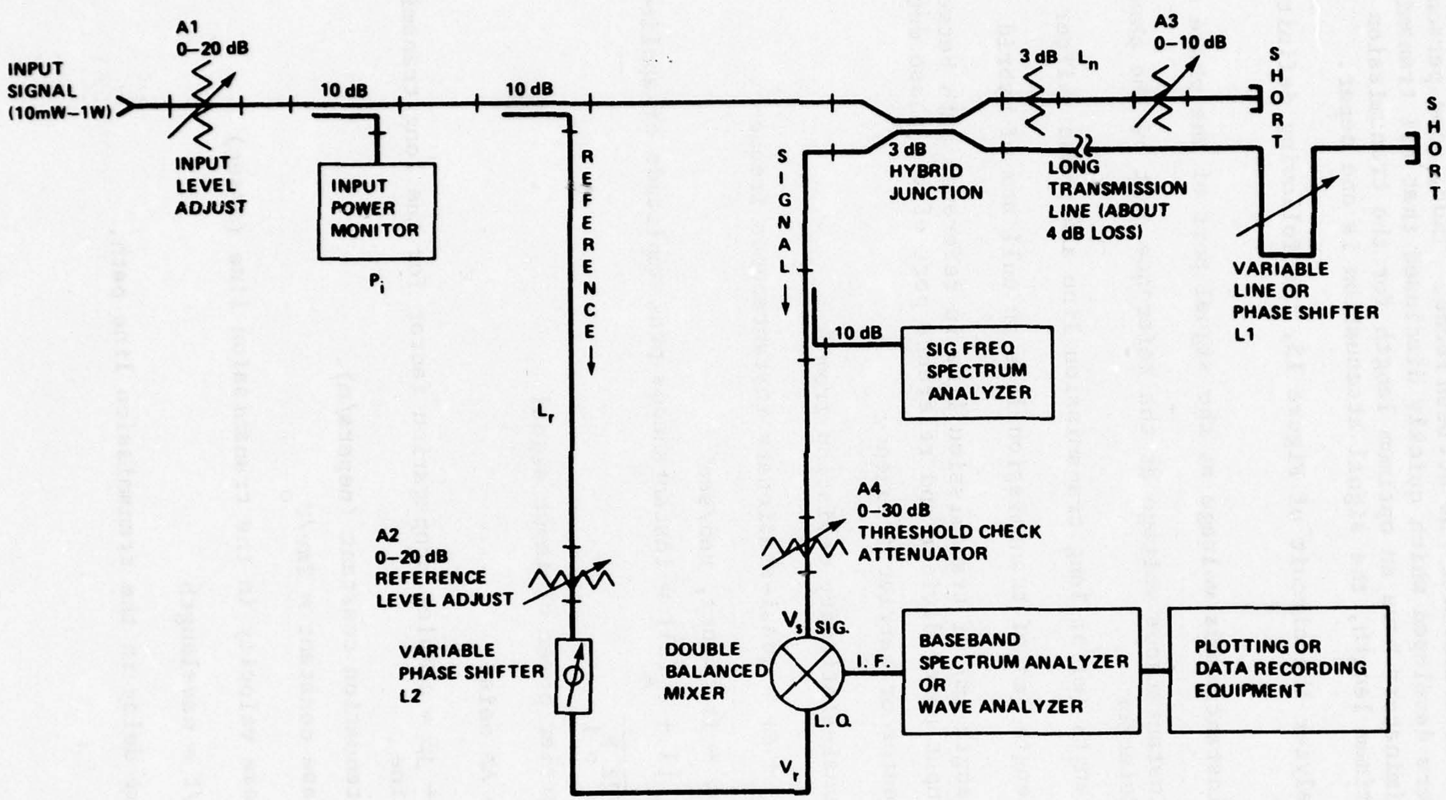

Figure 17. A wide operating frequency range transmission line discriminator.

A published theory to adequately illuminate the use of the transmission line discriminator has notbeen found. Thus, some personal theories were developed which quickly disclosed that all transmission line discriminators have an optimum length for the transmission line. At this optimum length, the signal attenuation is one neper.

To analyzer the circuit of Figure 15, the following definitions are needed:

v_s = instantaneous voltage at the signal port of the phase detector

v_r = instantaneous voltage at the reference port of the phase detector

L_s = length (m) of long transmission line and phase shifter

L_n = length (m) of transmission line on null arm of hybrid

L_r = length (m) of transmission line in reference path between input power divider and reference port of the phase detector

ν_c = center or carrier frequency

$\Delta\nu$ = quasi-stationary deviation from ν_c

$\nu = \nu_c + \Delta\nu$ = quasi-stationary instantaneous frequency

$\omega = 2\pi\nu$ = frequency, rad/sec

$A = A_c\,[1 + N_A(t)]$ = instantaneous peak amplitude of applied signal

$A_c = \sqrt{2Z_oP_i}$

P_i = carrier power of input signal

$N_A(t)$ = AM noise

$\Gamma = \alpha + j\beta$ = complex propagation factor for the long transmission line

α = attenuation constant (nepers/m).

β = phase constant = $2\pi\nu/U_o$

U_a = phase velocity in the transmission line (m/sec)

$\lambda = U_o/f$ = wavelength

T_d = time delay in the transmission line path.

The start of the derivation is determining the amplitude and phase of the signal applied to the SIG input of the phase detector, v_s. Since the input signal is A cos ωt, it is convenient to use phasors. (Re[ ] means "real part of". A carat over a variable means it is complex. $j \triangleq \sqrt{-1}$.)

$$v_s = R_e\left(Ae^{-2\Gamma L_s} e^{j\omega t} + AK_N e^{-j2\beta L_N} e^{j\omega t}\right) \tag{42}$$

$$R_e\left(\hat{V}_s e^{j\omega t}\right)$$

$$\hat{V}_s = A\left(e^{-2\alpha L_s} e^{j\beta L_s} + K_N e^{-j2\beta L_N}\right) \quad . \tag{43}$$

In these equations, the first term represents the transmission line path and the second term represents the carrier nulling path with K_N a measure of the magnitude. The factors of two in the preceding exponents occur because the signals traverse the lines in both directions. Defining

$$K_s = e^{-2\alpha L_s} \tag{44}$$

as the line loss ratio and using $\beta = 2\pi(\nu_c + \Delta\nu_c)/U_o$,

$$\hat{V}_s = A\left(K_s e^{-j4\pi(\nu_c+\Delta\nu)/U_o} + K_N e^{-j4\pi(\nu_c+\Delta\nu)\,L_N/U_o}\right). \tag{45}$$

The tuning procedure makes $K_s \approx K_N$ at null. To account for the inability to obtain a perfect null, let

$$K_N = K_s/\zeta \quad , \tag{46}$$

where

$$\zeta \approx 1 \quad . \tag{47}$$

Now

$$\hat{V}_s = A\,K_N e^{-j2\beta L_n}\left(\zeta e^{-j4\pi(\nu_c+\Delta\nu)(L_s-L_n)/U_o} + 1\right) \tag{48}$$

using $K_s \approx K_N$

$$\hat{V}_s = AK_N e^{-j4\pi(\nu_c+\Delta\nu)L_N/U_o} \left(\zeta e^{-j4\pi(\nu_c+\Delta\nu)(L_s-L_N)/U_o} + 1 \right) . \quad (49)$$

If $(L_s - L_n)$ is adjusted so that

$$e^{-j4\pi\nu_c(L_s-L_N)/U_o} = -1 \quad , \quad (50)$$

then

$$\hat{V}_s = AK_s e^{-j4\pi(\nu_c+\Delta\nu)L_N/U_o} \cdot \left(1 - \zeta e^{-j4\pi(\Delta\nu)(L_s-L_N)/U_o} \right) \quad (51)$$

for the case $\zeta = 1$ and small $\Delta\nu$

$$\hat{V}_s = AK_s e^{-j(\pi/2+(\nu_c+\Delta\nu)L_N/U_o)} \cdot \sin(4\pi(\Delta\nu)(L_s - L_N)/U_o) \quad (52)$$

which can be written

$$\hat{V}_s = |V_s| e^{j(\theta+\pi/2+(\nu_c+\Delta\nu)L_N/U_o)} \quad , \quad (53)$$

where

$$|V_s| = AK_s \, 4\pi(L_s - L_N)(\Delta\nu)/L_N \quad (54)$$

$$\theta = \mathrm{sgn}(\Delta\nu) \quad . \quad (55)$$

The phase detector output can be made equal to

$$v = v_o + \Delta v = 2(10^{-G_M/20} \, |V_s| \sin\theta \quad , \quad (56)$$

where G_M is the conversion loss in the mixer used as a phase detector and $v_o = 0$ for proper adjustment of the reference phase. For the total discriminator

$$\frac{\Delta v}{\Delta\nu} = 8\pi\sqrt{2Z_oP_i}\left(10^{-(G_M+G_H)/20}\right)(L_s - L_N)e^{-2\alpha L_s}/U_o \quad . \quad (57)$$

This simple derivation was made possible by the assumption $\zeta = 1$. In practice, $0.99 < \zeta < 1.01$ might be expected. To show that the exact value of ζ is not important, the Hewlett Packard 9830 calculator has been used to compute and plot the magnitude and phase of $\hat{V}_s$ with ζ set to give 40, 60, 80, and 100 dB carrier suppression (40 dB corresponds to $\zeta = 0.99$ or 1.01). These plots are shown in Figures 18 and 19. Setting these values of amplitude and phase in the equation for the phase detector output yielded output voltage curves which fell on top of each other. An artificial offset was added in making the plot of Figure 20. The conclusion drawn from these plots is that the depth of the null is not critical as far as slope is concerned. This has been proven by laboratory measurements.

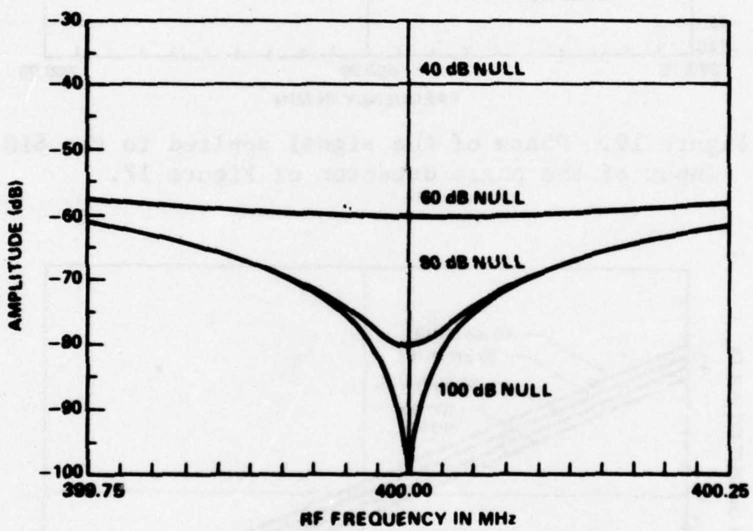

Figure 18. Amplitude of the signal applied to the SIG input of the phase detector of Figure 17.

Notice the analogy between the equations for the slope of the cavity resonator discriminator, Equation (39), and the slope of the transmission line discriminator, Equation (57). Both equations are proportional to $\sqrt{P_i}$ and energy stored in the system and both show that AM on the input signal is rejected in the output. Perhaps the only surprise is that Equation (57) does not have the factor ν_c. Therefore, the transmission line descriminator can be calibrated at one carrier frequency and then used over a ±5% carrier frequency band (with retuning at each value of ν_c within the band) without recalibration if P_i is held constant or corrected.

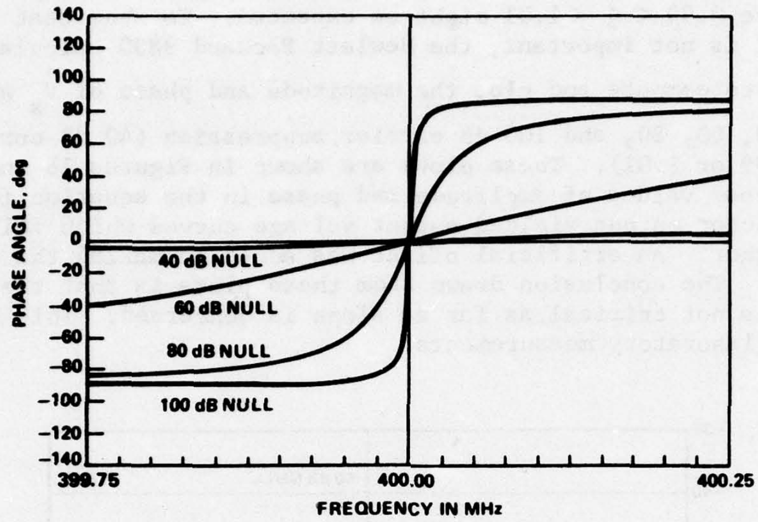

Figure 19. Phase of the signal applied to the SIG input of the phase detector of Figure 17.

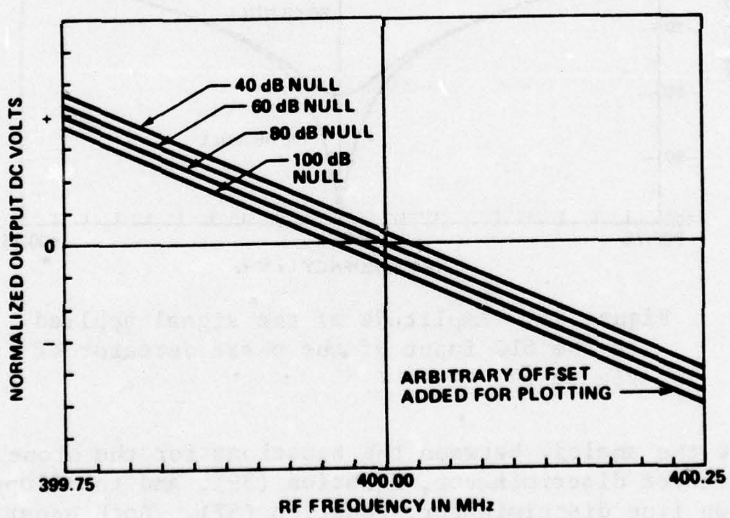

Figure 20. Output voltage versus input RF frequency for the discriminator of Figure 16.

The slope equation can be normalized and plotted as shown in Figure 21. Notice that this curve has a maximum at $\alpha L_s = 0.5$ corresponding to a round trip attenuation of one neper (8.68 dB) in the transmission line. Thus, it is known there is an optimum length for the transmission line which depends only on α, the loss per unit length. The lower the line loss, the longer the line and higher the slope. Intuition that indicated that the lower loss line should be better is indeed the truth.

The maximum shown in Figure 21 is broad enough that a transmission line cut to be optimum at one frequency can be used over a surprisingly wide carrier frequency range. For coaxial cable below 5 GHz, α is solely dependent on skin depth and changes as $\sqrt{\nu_c}$. Thus, for a 3-dB limit on sensitivity (and threshold) a line optimum at ν_1 can be used from two octaves below ν_1 to three octaves above ν_1.

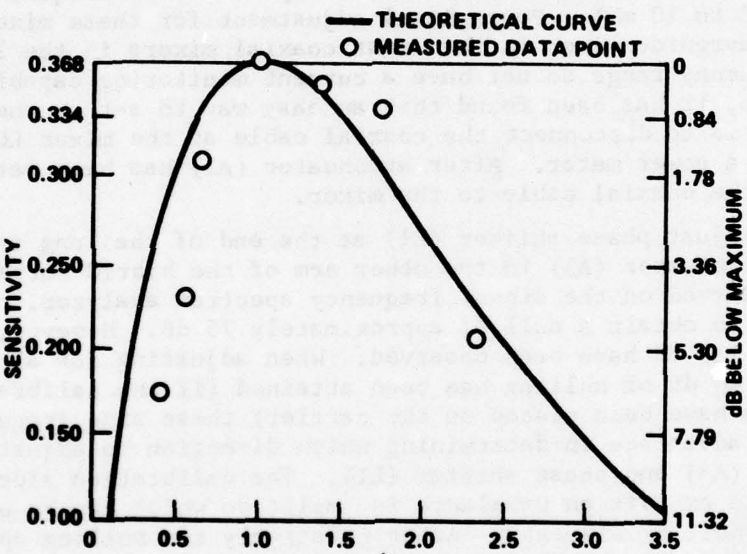

Figure 21. Sensitivity versus total transmission line length.

To verify the theory previously presented, detailed measurements have been made in the MICOM laboratories at Redstone Arsenal to confirm Figures 18, 19, and 20. It has been an even more gratifying experience using these transmission line discriminators to make FM noise measurements on sources from 10 MHz to 5 GHz. At 5 GHz, the slope is approximately 5 dB less than a cavity resonator discriminator $Q_o \approx 33{,}000$ corresponding to a TE_{013} mode.

The theory and operating procedure with this transmission line discriminator can be summarized as an alignment, calibration, and operating procedure. Obtaining calibration side-frequencies is essentially the same as discussed in paragraphs V.a) through V.d) and need not be repeated except to indicate it is a required operation. The typical transmission line discriminator shown in Figure 18 is normally implemented with coaxial components and can be operated over several octaves of input signal frequency range without changing mixer, directional coupler, and the hybrid junction. The alignment and operating procedure can be described in the following steps.

a) Adjust the input attenuator [A1] for approximately 20 mW going into the discriminator, then adjust the threshold attenuator (A4) for maximum attenuation.

b) Adjust the reference level attenuator (A2) for the proper power level at the input LO port of the mixer. (Normal LO power requirements for mixers of the coaxial type in these frequency ranges vary from 2 to 10 mW). Power level adjustment for these mixers differ from the waveguide mixer in that most coaxial mixers in the 2-GHz and below frequency range do not have a current monitoring capability. For this reason, it has been found that an easy way to set up the mixer LO port power is to disconnect the coaxial cable at the mixer LO port and connect in a power meter. After attenuator (A2) has been set, then reconnect the coaxial cable to the mixer.

c) Adjust phase shifter (L1) at the end of the long transmission line and attenuator (A3) in the other arm of the hybrid for a carrier null as observed on the signal frequency spectrum analyzer. Normally, it is easy to obtain a null of approximately 75 dB. However, nulls as good as 110 dB have been observed. When adjusting for a carrier null after 40 dB of nulling has been attained (if the calibration side frequencies have been placed on the carrier) these side frequencies can be used to advantage in determining which direction to adjust the attenuator (A3) and phase shifter (L1). The calibration side-frequencies will tend to exhibit an unbalance in amplitude which can be used to adjust the nulling elements. After relatively few nulling operations it becomes very easy to watch the side-frequency amplitude changes and then be able to tell which element should be adjusted and in which direction (i.e., increase or decrease attenuation and more or less phase shift). A properly tuned null will exhibit a carrier with a side-frequency amplitude not more than 10 dB below the carrier (70-dB null) Both side-frequencies will have the same amplitude. The display will be completely symmetrical on the spectrum analyzer used to monitor the nulling operation.

d) Reduce threshold check attenuator (A4) to zero. Reduce input level attenuator (A1) to zero while readjusting reference level adjust attenuator (A2) (remember to use the power meter) to maintain proper mixer LO port power level. Apply angle modulation (if this was not done in the nulling procedure) with an amplitude of 50 to 60 dB below the incoming carrier and offset from 10 to 20 kHz from the carrier.

e) Locate the modulation frequency on the baseband spectrum analyzer. Adjust variable phase shifter (L2) for a maximum signal output on the baseband spectrum analyzer. Remove the angle modulation calibration signal. (The system is now aligned, calibrated, and ready to record data.)

f) Calibrate the discriminator slope, S, using Equation (40) presented in the slope calculation in paragraph V.f) and store this value for later use in data processing.

g) Insert maximum threshold check attenuation (A4) and operate the wave analyzer to record the equipment threshold or noise floor.

h) Reduce threshold check attenuator (A4) to zero, check for baseband overload, and operate the wave analyzer by recording the FM noise as the system is operating.

i) If the FM noise measurement is less than 10 dB above the equipment threshold, calculate the corrected deviation at each modulation frequency, using Equation (41), paragraph V.i). Note that the bandwidth for the measurement was set in the baseband wave analyzer and must be stated along with the values for the FM noise.

j) Convert the FM noise data to other bandwidths, forms, etc., as required for the format of presentation desired.

As in the case of the cavity discriminator, the most vexing problem is inadequate threshold. A typical example would be measuring a crystal oscillator directly at 100 MHz or lower frequency. After finding the lowest loss coaxial line that one can afford, the threshold is still too high and the only remedy is to use some form of amplification, or use the two-oscillator method, being equally careful to determine threshold. Approximately 30-dB of amplification with a power output of approximately a watt is needed to overcome the threshold problem. This presents the question "will the amplifier noise confuse the measurement?" This is the topic of the Section VII, but this experience indicates that such an oscillator can be measured with an amplifier and transmission line discriminator.

For the total microwave transmitter, this issue is evaded by noting that the amplifier is needed to drive the frequency multiplier. It is the output of this amplifier which is important, and this can be measured with the discriminator. Once the power is increased and the deviation increased by the frequency multipler, one can measure with ease and confidence in the submicrowave and lower microwave regions.

VII. REDUCTION OF FM NOISE MEASUREMENT THRESHOLD

The problem which has caused the most trouble in the laboratory is the measurement of FM noise on low power signal sources. At microwave frequencies, noise reduction by transmission cavity stabilization reduces output power by typically 6 dB as well as significantly reducing the FM noise. At submicrowave frequencies, the basic crystal oscillators are operated at relatively low output power and even after amplification the following frequency multiplier loses power in the process. However, these sources for microwave signals represent the state-of-the-art in low noise and must be measured.

The discriminator slope equations for the cavity resonator discriminator, Equation (39), and the transmission line discriminator, Equation (57), show that the slope is proportional to $\sqrt{P_i}$ and energy storage (Q of the cavity, length of the line) in the system. The noise generated in the crystals of the phase detector is constant because the local oscillator port input power is held constant. Thus, the threshold depends on $\sqrt{P_i}$ and energy storage. This is illustrated in Figure 22 for the cavity discriminator. After one has used the highest possible cavity Q_o, or the optimum length of lowest loss transmission line, the only way to reduce the measurement threshold is to add amplification to the system. This amplification must be added with care so that the verification of the system threshold is not lost. One useful method for adding this amplification was developed by Ashley and Palka [23].

The usual noise properties of an amplifier (expressed as the noise figure) are not important because this amplifier must operate at a high carrier level. The important factor is that the amplifier not add FM noise to the signal test. Also, since the discriminator rejects AM, the amplifier can add some AM noise without degrading the measurement. The data published by Ashley and Palka [24] show that for a locking gain of approximately 20 dB an injection-locked oscillator will reproduce (below 100 kHz) the FM noise of the locking signal for any but the most sophisticated of signal sources. These data are shown in Figure 23.

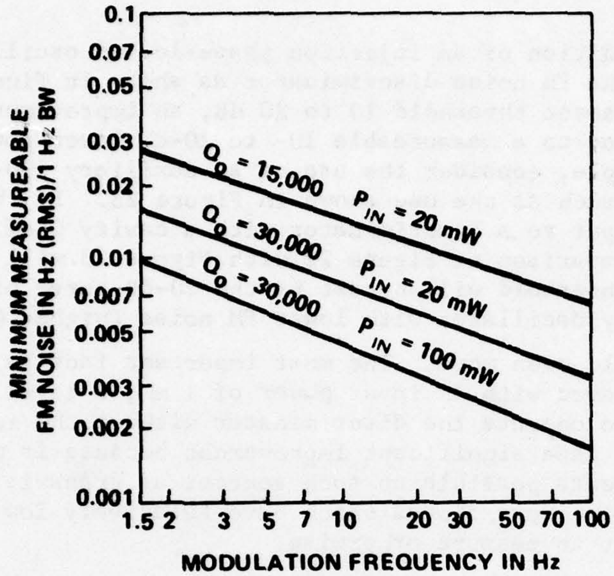

Figure 22. Threshold level of a 16-GHz FM discriminator. (This threshold is set by noise in the phase detector crystals).

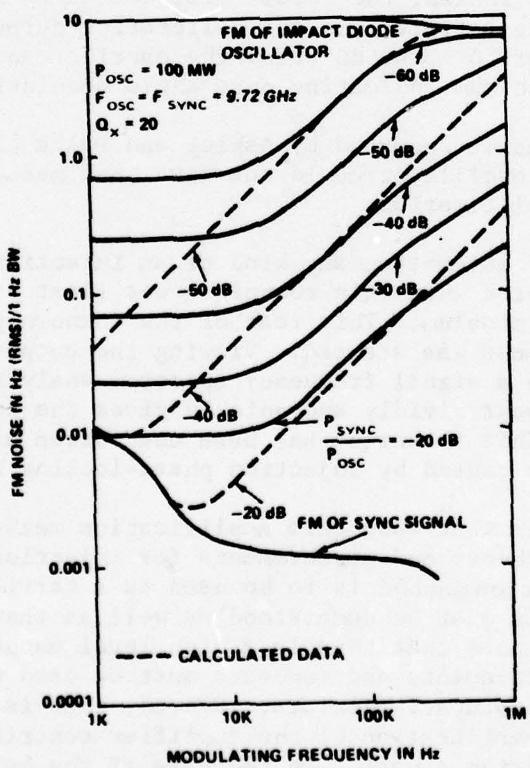

Figure 23. FM noise in an injection-synchronized phase avalanche diode oscillator.

Thus, the addition of an injection phase-locked oscillator as an input stage for the FM noise discriminator as shown in Figure 24 can improve the measurement threshold 10 to 20 dB, an improvement which is directly translated to a measureable 10- to 20-dB lower power signal. As a typical example, consider the use of an auxiliary 100-mW avalanche-diode oscillator such as the one shown in Figure 23. If this oscillator is used as the input to a discriminator with a cavity Q of 20,000 to 30,000, then a comparison of Figure 22 with Figure 23 will show that the measurement threshold will be set by the 20-dB curve of Figure 23. Using an auxiliary oscillator with lower FM noise (higher Q_{ext}) would lower the threshold even more. The most important fact is that this threshold is achieved with an input power of 1 mW, a level that would be insufficient to operate the discriminator without the auxiliary oscillator. This is a significant improvement because it makes good FM noise measurements possible on such sources as transmission cavity-stabilized IMPATT or Gunn diodes which have relatively low power and are most difficult to measure otherwise.

In addition to improving the measurement threshold, the auxiliary oscillator also makes it much easier to calibrate the discriminator. The usual oscillator can be frequency-modulated by applying AC to the bias port. While locked, the result will not be FM but ΦM; however, this will be quite satisfactory for calibration purposes because a side-frequency level of 50 to 60 dB below the carrier can be achieved with a symmetrical spectrum indicating pure angle modulation.

A typical result reported by Ashley and Palka [23] is shown in Figure 25. This oscillator could not have been measured without some type of added amplification.

Some fear of attempting any kind of an injection phase-locking experiment was noted when this technique was first proposed to someone with a threshold problem. This fear of the unknown promptly receded after the experiment was started. Viewing the output of the circulator of Figure 24 with a signal frequency spectrum analyzer will show the locking process most vividly and quickly gives the needed confidence in the method. This technique has been used often (also by Mr. Palka) and no difficulty caused by injection phase-locking has been experienced.

The main idea that made this amplification method work was the availability of theory and measurements for injection phase-locking. If another amplification method is to be used as a carrier frequency pre-amplifier, then it must be understood as well as the injection phase-locking. First, note that this is a high level output and the usual noise figure measurements and concepts must be used with caution if any degree of output saturation exists. Second, this is a chicken and egg problem because verification of the amplifier contribution requires a known low noise drive signal. In the case of the injection lock noise

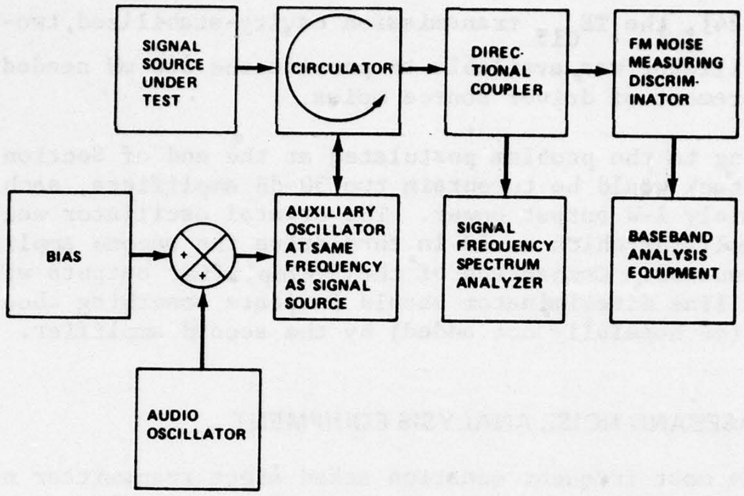

Figure 24. Use of an injection-locked oscillator to lower the threshold of an FM measurement.

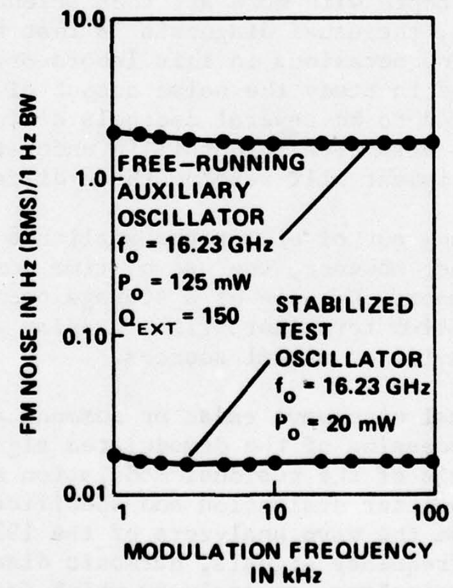

Figure 25. Typical measurement of a low noise, low power oscillator.

experiment [24], the TE_{015} transmission cavity-stabilized, two-resonator klystron oscillator was available to provide the 600 mW needed to make a good measurement of driver source noise.

Reverting to the problem postulated at the end of Section VI, the method of attack would be to obtain two 30-dB amplifiers, each capable of approximately 1-W output power. The crystal oscillator would then drive one amplifier which would in turn drive the second amplifier via a 30-dB attenuator. Comparison of the two amplifier outputs with the transmission line discriminator should indicate something about the FM noise added (or hopefully not added) by the second amplifier.

VIII. BASEBAND NOISE ANALYSIS EQUIPMENT

The most frequent question asked about transmitter noise measurements is: "Are these measurements repeatable from one laboratory to another?" The answer to this can be only a qualified "yes", with the qualification being that the personnel making the measurements to be compared have at least several man-months of experience in what is often regarded as a topic with more art than science. When differences are enumerated, the usual diagnosis is that the baseband equipment is different. On many occasions in this laboratory, two different analyzers have been used to study the noise output of a common discriminator. The answers were found to be several decibels different on first evaluation. Only careful attention to detail in understanding and calibrating the baseband equipment will resolve these differences.

The signal coming out of either the amplitude or angle demodulator is a function of time; however, the use of time domain analysis equipment is not very common. The use of a storage oscilloscope [3] has proved useful for medium term instability display and the display of pathological misbehavior of signal sources.

Because the usual microwave radar or communications system employs frequency domain processing of the demodulated signals, it is the frequency domain analysis of the residual modulation noise which is most significant to transmitter evaluation and specification. The equipment used has evolved from the wave analyzers of the 1930's developed for the study of audio frequency signals, harmonic distortion in amplifiers, etc. In current terminology, an analyzer which is manually tuned (or perhaps sweep-tuned with a motor drive) is called a wave analyzer, while an analyzer which has an electronic sweep of frequency is called a spectrum analyzer. The obvious conflict of terminology is that a sweeping analyzer for displaying the spectrum RF or microwave frequencies is also called a spectrum analyzer. An idea of the frequency range of the input signal (baseband, RF) is attempted whenever using the name spectrum analyzer.

Baseband analyzers operating in the frequency domain can be further classified as either constant bandwidth (e.g. 100 Hz) or constant percentage bandwidth (e.g. 1/3 octave.) Because the theory of noise is rooted in the idea of constant measuring bandwidth and because system noise performance depends on constant incremental bandwidth (i.e., multiplexing a number of 3-kHz channels into a broad baseband signal for transmission) the constant bandwidth analyzers are more frequently used, although some operating time convenience could be achieved by correct use of constant percentage bandwidth equipment. In this last category, measurement time can be tremendously reduced by operating parallel banks of overlapping constant bandwidth filters. This equipment is usually called a real-time analyzer.

Digital filtering and processing of noise data to obtain frequency domain information (by use of correlation computation and fast fourier transform methods) is gaining wide acceptance in environmental studies but has not been used for analyzing modulation noise.

The constant bandwidth wave or spectrum analyzers are superheterodyne receivers such as the one shown in Figure 26. This particular equipment has been selected for discussion because it is typical of all constant bandwidth analyzers and illustrates the basic concepts.* Other equipments may use higher IF to accommodate wider input frequency range, but the block diagrams will be remarkably similar.

The input circuits box contains three essential items:

a) Input attenuator system.

b) Amplification.

c) A low pass filter.

The input attenuator is needed when the analyzer is used for relatively high level input such as during the calibrate or alignment stage of a noise measurement. To minimize the noise contribution of the internal stages of the analyzer, this attenuation must be adjusted to the minimum possible loss for a measurement.

The amplifier is needed to allow measurement of lower level signals such as the noise output from a demodulator. Because the noise figure for an amplifier can be made several decibels better than the noise figure for the mixer, the noise performance of the analyzer can be improved with at least 10 dB of gain in the input amplifier. The amplifier must be used with care because it can be over driven (as can the following mixer) to make a measurement invalid. An input stage overload indicator is a tremendous operational convenience.

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*This discussion is not intended as an endorsement of any particular commercially-available equipment.

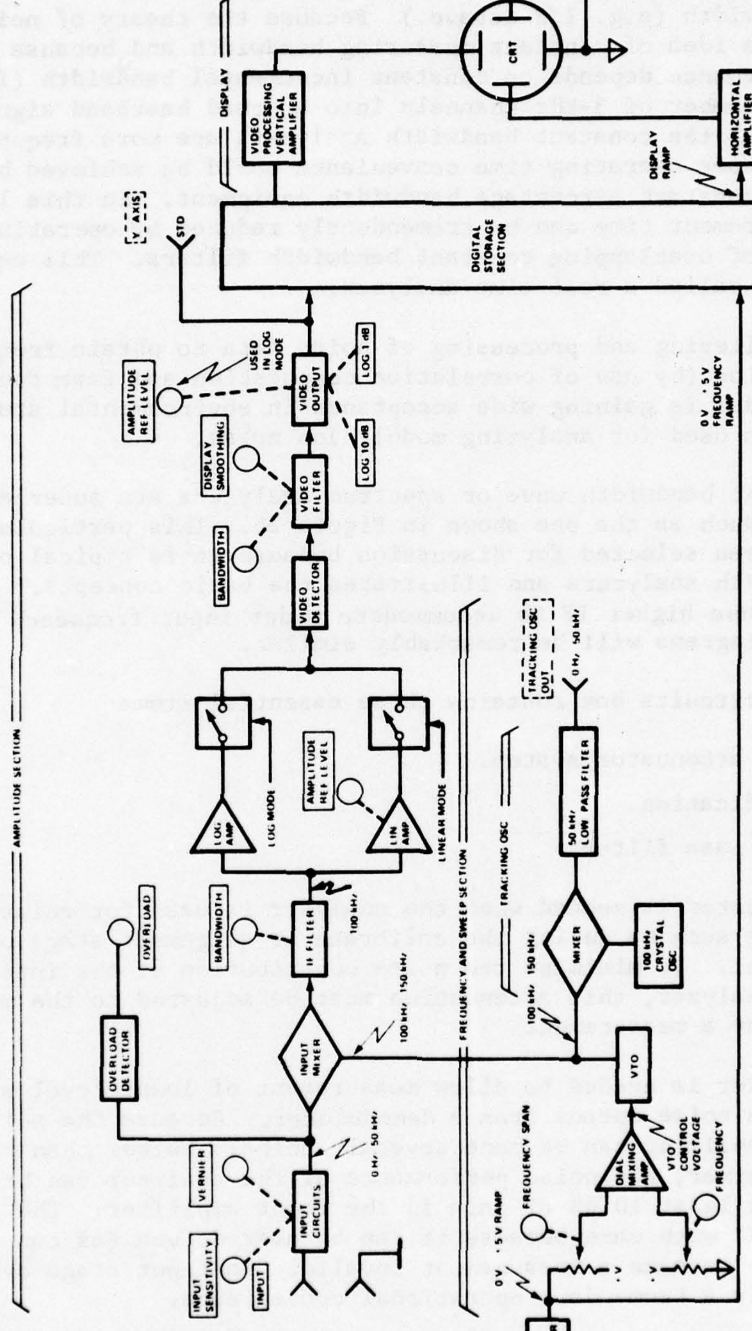

Figure 26. Simplified block diagram of an HP-3580A spectrum analyzer.

The low-pass filter is needed to reject signal components at IF (100 kHz) and in the image band (200 to 250 kHz).

The mixer converts the input spectrum (0 to 50 kHz) to sidebands above and below the local oscillator frequency (100 to 150 kHz). The mixer is balanced to minimize the local oscillator component in the output. The objective is to move the input spectrum past the fixed frequency and relatively narrow bandpass of the IF amplifer and filter. The dynamic range of the analyzer is set in the mixer. The mixer normally works so well that one is not appreciative of the design skill required.

The stability and operational features designed into the local oscillator system (a swept voltage tuned oscillator in this example) are the biggest factors in determining the stability and operational features of the analyzer. These features can be appreciated without detailed study of the circuits.

The IF filter stage sets the measurement bandwidth which is crucial to the accurate analysis of the noise signal. In this analyzer a five-stage synchronously-tuned crystal filter has a bandwidth (between 3 dB points) of 1 Hz. By adding resistive losses and compensating the interstage amplifer gains, the bandwidth can be swtiched in a 1-3-10 sequence to a maximum of 300 Hz. For this synchronous tuning, the frequency response has the familiar bell shape of the Gaussian function.

The choice of filter shape is usually made on the basis of measuring a spectrum consisting of a collection of discrete frequencies (rather than noise.) The first decision is the choice of gaussian or rectangular shape. In sweeping analyzers, the gaussian shape is preferred for nearly optimum buildup as a sinusoidal signal is swept past the filter. For manually-tuned analyzers, the rectangular shape is preferred to allow a minor amount of frequency drift between the source generating the spectrum and the local oscillator in the analyzer.

For noise analysis, the choice is not important. It is important to know the noise bandwidth of the response shape - a number not normally in the instruction manual. If the frequency response of the amplifier, H(jf), is centered at f_o and is down 40 dB at f_{LO} and f_{HI}, then the noise bandwidth is

$$B_N = \frac{1}{|H(jf_o)|^2} \int_{f_{LO}}^{f_{HI}} |H(jf)|^2 \, df \quad . \tag{58}$$

It has been found [25] that approximately 30 data points defining H(jf) are sufficient for a very simple rectangular rule integration to determine B_N. This must be done for each bandwidth of the analyzer.

Typically, the noise bandwidth is wider than the 3-dB bandwidth. For this analyzer, it is typically 12% wider than the measured 3-dB bandwidth.

Another specification of the IF filter shape relates the frequencies at which the filter attenuation is 60 dB and 3 dB. The ratio of a 60-dB bandwidth to a 3-dB bandwidth is called the shape factor and is approximately 10 to 1 for gaussian shape filters. This factor determines the spacing between discrete frequencies which can be resolved by the analyzer and the lowest frequency which can be measured without interference from the zero response. As a good rule, discrete frequencies spaced 15 times the bandwidth can be accurately measured. The minimum frequency which can be measured is approximately eight times the IF bandwidth. The cause of this limitation is the leakage of the local oscillator output through the balanced mixer. As the analyzer is tuned to low input frequency, the leaked local oscillator signal rides up the skirts of the IF filter response and registers as an output of the filter which must be classed as an error.

The output of the IF filter is amplified by a bandpass amplifier to increase the voltage enough to adequately override dc drift in the detector circuitry. To increase the dynamic range of the analyzer, attenuators are again used to optimize signal levels in the amplifer.

The logarithmic compression of a decibel vertical scale is most useful for first looks at a signal. The most stable and smooth way to accomplish the logarithmic compression is to perform this operation in the IF amplifier. Thus, the spectrum analyzer that has been described in Figure 26 has a second IF amplifer, the LOG AMP, which has a dynamic compression range of 120 dB. This amplifier causes no extra thinking for sinusoidal signal measurement, but does exercise the brain cells for noise measurement.

If the input to the spectrum analyzer has a gaussian probability density function, the output of the narrowband IF filter also has a gaussian density function. The nonlinearity of the logarithmic compression changes this density function and thereby changes the power relationship. The result of the compression of the peaks is that a true root mean square (RMS) detection circuit will read 1.45 dB less than the correct value. If the LOG AMP is used, then noise readings must be corrected by adding 1.45 dB while sinusoidal readings do not require correction.

After IF amplification, the detector stage produces a dc output proportional to the IF output voltage (or the logarithm for the decibel output mode.)

In most analyzers, the detector responds to the time average of the IF output voltage. This is solved in the calibration by the well known ratio of average to RMS for a sinusoid. However, when used as a noise analyzer this is in error by 1.05 dB, which must be added to the output indication to obtain the effective time average (RMS) of the noise. A true RMS detector would be better for a noise analyzer.

For the measurement of sinusoidal signals the dc from the detector is a stable value; however, for noise measurement it bounces around with the bigger bounces on the narrower IF bandwidths. Post-detection filtering (the video filter of Figure 26) reduces the bounce of the output at the expense of measurement time.

The detected output is used to drive some form of output indicator. In the usual wave analyzer the indicator is a simple dc voltmeter calibrated to read RMS volts and possibly decibels with respect to a stated reference. For the spectrum analyzer the output indicator is often an oscilloscope with some form of memory. For the analyzer of Figure 26 the memory is digital with two sections; thus, two different frequency sweeps of data can be compared.

All of this digression into nonmicrowave instrumentation must be understood to make effective microwave transmitter noise measurements. Before using a baseband spectrum analyzer, spend a few hours reading the instruction book - it will be time well invested.

IX. EQUIPMENT OPERATION

Once one has decided to make a transmitter noise measurement, decided on the methodology, acquired the necessary equipment, and located a working space, a very significant question arises: What is the sequence of operations? Intuition might guide one through the maze, but it is doubtful that the path will be close to optimum; furthermore, if it is hoped to automate the measurement, the sequence of operations must be thought out in advance and worked through on a manual basis to understand the development of automatic routines such as those presented in the next section.

This experience is outlined in the operations flow chart of Figure 27. A similar operations sequence can be developed for AM noise measurements. (This development is left as an exercise for the student.) Discussion of this flow chart will be the basis for sharing laboratory experience with the reader.

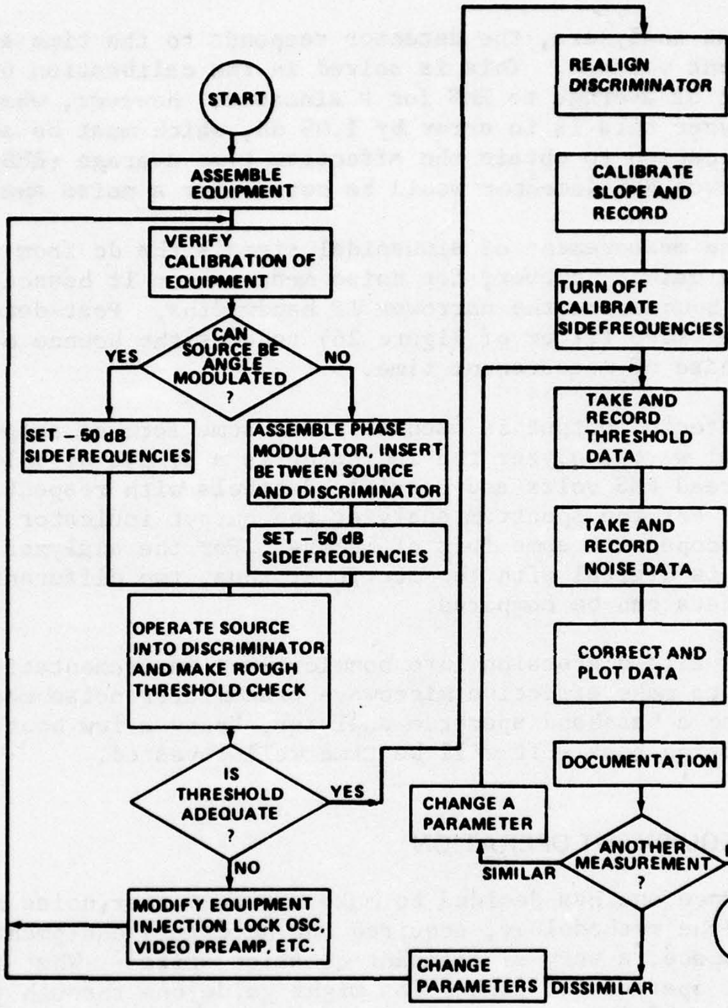

Figure 27. Operations flow chart for making FM noise measurements.

When the equipment is assembled and operating, all of the defective equipment has probably been found and repaired. Next, take the time to verify the calibration of the RF spectrum analyzer, power meters, audio oscillators, and precision attenuators. For the RF spectrum analyzer, determine if the IF attenuator calibration is correct. The calibration of the baseband spectrum analyzer will not normally include the noise bandwidth, so a separate project needs to be initiated to take the data and do the numeric integration to determine this bandwidth for each position of the bandwidth selector switch. While considering the baseband analyzer, check to see if the IF amplifier is linear or logarithmic and if the detector is average or RMS responding. Make note of the necessary correction factors.

Once the source is working and the discriminator has been aligned, a few checks will indicate if it is worthwhile to take detailed data. With the calibrate side-frequencies on, tune the baseband analyzer to this modulation frequency and check for overload. Operate the input attenuator for maximum sensitivity without overload. If the analyzer does not have an overload indicator, take 10 dB out of the IF attenuator and put in 10 dB with the input attenuator. The output indicator should remain unchanged. If a preamplifier is used ahead of the baseband analyzer, use an oscilloscope to look for clipping at the output. Insertion of 3 dB at the input of the discriminator should make the output indicator fall 3 dB if nothing is overloading.

Turn off the calibration signal and reduce the input attenuator of the baseband analyzer as much as possible without overload. If the output indication is not useful, then remove IF attenuation. If one cannot achieve at least midscale indication at the analyzer output with the minimum bandwidth to be used, then a preamplifier is needed for the input of the baseband analyzer. After this check is passed, drop the input power by 3 dB and note whether the output indicator also drops 3 dB. If it does, the threshold check routine to be performed later will yield a negligible correction. If it drops at least 1 dB, there is hope that the threshold check routine will allow correction of the final data to reasonable accuracy. If it drops less than 1 dB, then the threshold is not adequate and either a video preamplifier or the injection-locked oscillator of Section VII must be added.

After it is known that the threshold is adequate, then it is worthwhile to take detailed data. If the threshold is more than adequate and data are being manually taken and recorded, the slope calibration can be manipulated to make the full scale of the output indicator a convenient deviation. For example, the 10-μV full scale might be made equivalent to 10-Hz deviation by clever manipulation of input power and the variable gain knob on the analyzer. At each point where noise data are to be taken the threshold equivalent deviation or actual voltage, the analyzer bandwidth, and the modulation frequency must be recorded. The same data are taken to determine the noise. Several important factors were written into the step by step alignment and operation instructions for the discriminators; steps V.d) through V.i) and V.d) through V.i) of Sections V and VI.

The preceding recording of data is made most convenient if an X-Y recorder can be attached to the wave or spectrum analyzer. This raises the question of how fast the frequency can be swept in taking the data. For noise spectra, which are dominated by discrete frequencies (often caused by power supply ripple), the rule is

$$\text{Sweep rate} = 0.5\ (\text{Bandwidth})^2\ \text{Hz/sec} \quad ,$$

where the bandwidth is also in hertz. The reason for this rule is that time is required for the high-Q filter in the IF section to build up to the steady state. If the noise spectrum is random and white or pink in shape, then this sweep rate can be multiplied by a factor of five which is an appreciable saving in data taking time.

Notice that the sweep rate goes as the square of bandwidth so the largest possible bandwidth should be used. This is where judgement in taking the data is required. If modulation frequencies below 1000 Hz must be studied, a 10-Hz or smaller bandwidth is needed to resolve power supply ripple components spaced at 50 or 60 Hz and to approach zero frequency as closely as the baseband analyzer will allow. For higher modulation frequency ranges, make a first sweep with a wider bandwidth, e.g., 100 or 300 Hz. If the result is a white or pink spectrum with no significant humps, the data can be given the final processing and used. If a hump is seen, make a second sweep with the narrowest possible bandwidth and at a sweep rate appropriate for sinusoids. The reason for this advice is shown in Figure 15. The lump in the curve taken with the 300-Hz bandwidth at 3700 Hz might normally be disregarded as something mysterious in the TWT. Use of the 10-Hz bandwidth and a much slower sweep rate shows that a fine structure related to the 60-Hz power supply is the cause of this hump. Additional data show that the same spectrum analyzer can detect the ripple components in the TWT amplifier power supply. A look in the time domain showed switching spikes from the silicon rectifiers were getting past the regulator stage.

Figure 15 also illustrates several other important facts concerning spectrum analyzer operation and data interpretation. For the 300-Hz bandwidth data, notice the similarity of the curve below 500 Hz and the shape of the IF filter response. This is the zero response (helped with a big spectral component at 120 Hz) of the analyzer. In the range above 1500 Hz a pure random noise spectrum can be measured in one bandwidth and mathematically transformed to another. For these two bandwidths the curve should be separated by 30 dB, which is not true in Figure 15. Thus, before mathematically transforming bandwidth, take a narrowband sweep to insure that the spectrum is that of random noise only.

The rules for mathematically transforming noise data to different bandwidths are based on the fact that noise power through a linear filter is proportional to the bandwidth. Thus, if data such as AM noise are taken in decibels with a measurement bandwidth, B_N, transformation to another band, B_A, is accomplished by adding $10 \log_{10} (B_A/B_N)$. An RMS noise deviation is transformed by multiplying by $\sqrt{B_A/B_N}$. Note, these transformations are valid only when the noise spectrum is flat over the larger bandwidth.

One block of Figure 27 which is often neglected, and later regretted, is documentation. This is especially true when taking data with a sweeping analyzer, record bandwidths, sweep rates, sweep ranges, and something to indicate if the noise measurement threshold is low or must be corrected. It is important that the curves for the deviation scale be marked.

The decision block after completing one data run shows a path around the data taking loop which does not include the calibration block. When taking this shortcut, remember that the calibration remains constant for 5% carrier frequency changes and for constant input power. The preceding information applies to well-behaved portions of a transmitter. While working with klystrons, backward wave oscillators (BWO), IMPATT oscillators, TEO, and crystal-oscillator controlled sources, pathological behavior typically in the form of big jumps in the output of the baseband analyzer has been observed. The jumps seem to occur at randomly-related bands of frequency, but are occuring over a wide band of baseband frequencies and coming and going with time. A better indicator is to use a preamplifier and storage oscilloscope to show the time of occurrence of these noise bursts.

X. AUTOMATION OF FM NOISE MEASUREMENTS

After one has traversed the major and minor loops of Figure 27 it will be found that most of the time is spent taking, documenting, and processing data. It is difficult to visualize the complete automation of the noise measurement process, but not so difficult to visualize computer control of the noise data taking and processing to significantly reduce operating time.

The first decision is the choice of calculator, minicomputer, microprocessor, or time-shared computer as the controller for the automatic system. For this automation, only a few instruments are controlled and the measurement time is set in the equipment; therefore, blinding speed and large size are not needed in the controller. The calculator has enough size and speed for this job. The task is similar in size and speed to many other measurement systems. This class of systems brought about the development of the IEEE standard #488-1975 interface bus for calculator or computer control of measurements and data processing. Using the IEEE standard #488-1975, the Hewlett-Packard Loveland Instrument Division has developed a spectrum analyzer system for the 10-Hz to 13-MHz range, the HP 3045A system.* The controller is the HP 9830A calculator which is programmed in the computing language BASIC. The software has been written to use this system in partially automating transmitter noise measurements.

The organization first divides the tasks into manual or calculator control as shown in Figure 28. Next, the programming is subdivided on the basis of the operations sequence (Figure 27). In this sequence, there are branching paths and even the possibility of branching within blocks. Several of the blocks have enough work to be subprograms. One

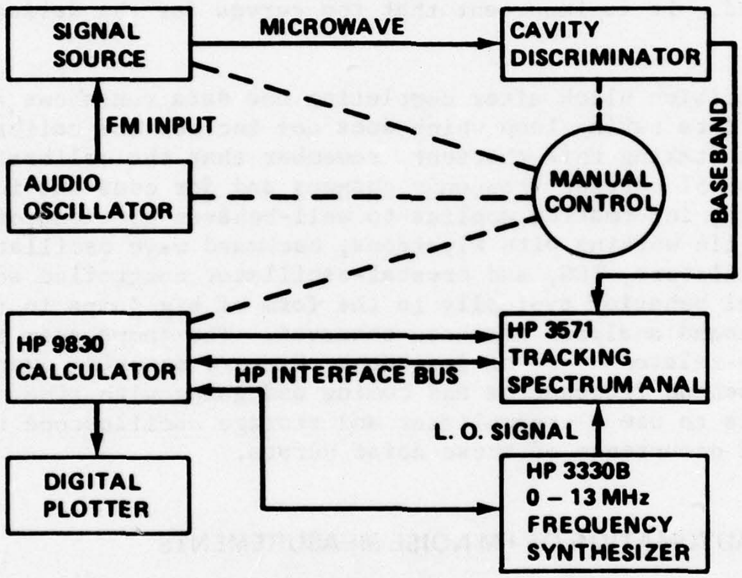

Figure 28. Automated measurement of near carrier noise.

simple way to implement the calculator control is to use the 20 function keys to control the operation of various portions of the program; then, much of the branching is accomplished by pressing the appropriate function key. The following subprograms were written and stored on the function keys.

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*This discussion is not intended as an endorsement of any commercially available equipment.

Name	Function
CAL	Calibrates all equipment
THRESH	Stores threshold data
NOISE	Takes and stores noise data
PLOT	Processes and plots noise data
LINEAR GRID	Draws a linear grid for plotted data
LOG GRID	Draws a logarithmic grid for plotted data

The function keys with the appropriate overlay are shown in Figure 29. Each of these subprograms can be operated independently as far as the calculator is concerned; this makes possible several noise data runs for one calibrate and threshold run.

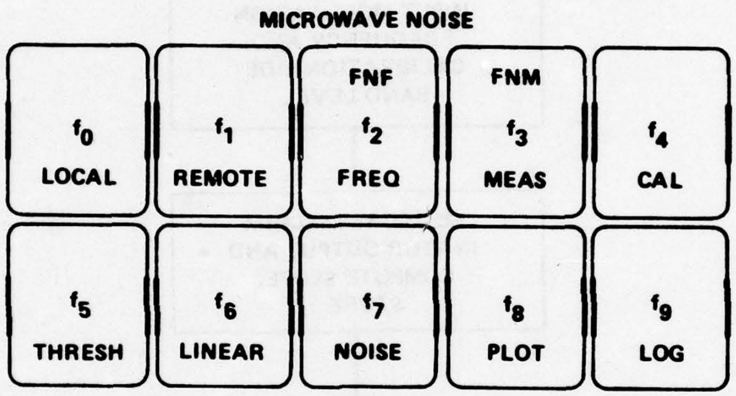

Figure 29. HP-9830 function keys used to sequence the automated noise measurement.

Usually, the first step is operation of CAL. The calculator issues control statements and takes data to calibrate the slope (hertz deviation per volt of output) of the microwave discriminator. This is done by applying a small angle modulation to the source under test and adjusting for a known deviation. (The equations to use were given with the discriminator theory.) The result of this calibration is stored in the calculator memory and used in later data processing. The flow chart for subprogram CAL is shown in Figure 30, and the BASIC program used by the HP-9830A calculator is presented in List A-13, page 88.

The second step in the measurement sequence is operation of THRESH to determine the noise floor or threshold of the discriminator. The theory is presented in Sections V and VI, but the practice is very time consuming because data for the threshold must be taken at each frequency where noise is to be measured. The calculator is given the frequency limits for which measurements are desired and the number of data points. The threshold data are then taken and the results stored in an

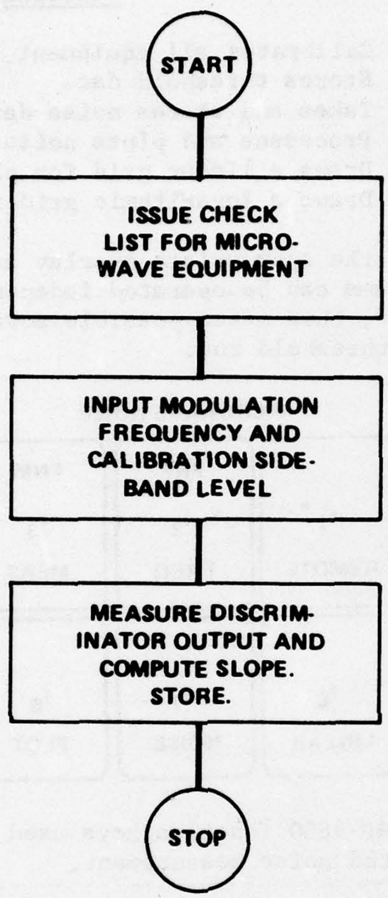

Figure 30. Flowchart, subprogram CAL.

array variable which will be used later to correct actual noise data. Thus, all of the measurements can be corrected for threshold before plotting.

The third step is to use NOISE to take data at the same frequencies where the threshold is known from the previous step. This is completely under the control of the calculator which again stores the results in its memory (as an array variable). During this step, the operator can observe the operation of the source under test and the microwave discriminator tuning. Those who have spent many hours of tuning a wave analyzer and recording data with pencil and paper appreciate the ease of this step.

Probably the most enjoyable step is pushing the calculator key labeled PLOT. This subprogram takes the arrays for threshold and noise, calibration factors, and other corrections to compute threshold-corrected noise deviation while the calculator-driven plotter plots complete results. Documentation is made painless by answering "yes" to a displayed question, "Do you want to label the axes?" The result is a plot such as the one shown in Figure 14 where everything but the caption was done by the plotter. The linear grid portion was plotted by activating the proper function key. After years of doing noise measurements the hard way, it is pleasure to watch the plotter deliver the results of automated noise measurements.

Far more significant than the pleasure of the engineers is the time savings and improved documentation that result from automation. It has been proved by personal example that someone familiar with making FM noise measurements manually can learn to operate the automated system in less than 1 hour. For production line testing, test personnel can be taught to operate this automated equipment much easier than one can be taught to make manual controlled noise measurements and to perform the data processing calculations. If significant numbers of noise measurements need to be made, then the extra equipment to automate the system will quickly pay for itself.

At the time of this writing, we are at perhaps the 90% point on the automation learning curve. It is premature to formally document all of the subprograms and listings; however, this work has been written as an appendix which includes all listings. Workers at the Hewlett-Packard Loveland Instrument Division have prepared an applications note on the same topic [26].

XI. CONCLUDING REMARKS

Ten conclusions have been listed at the end of the theoretical part of this report in Section III. The following conclusions regarding the experimental technique of making transmitter noise measurements are a continuation of those listed in Section III.

a) The measurement of AM noise in all stages of a microwave transmitter is best done with direct diode detectors. The worst pitfall is failing to determine the measurement threshold.

b) The measurement of FM noise for carrier frequencies above 5 GHz is now a repeatable and reliable procedure using the discriminator of Section VI. Be careful in calibrating and determining threshold.

c) The measurement of FM noise for carrier frequencies below 5 GHz can now be accomplished with the improved transmission line discriminator of Section VII. After a few months of operational experience at frequencies between 30 MHz and 5 GHz, it is believed that the technique is as repeatable and reliable as the methods above 5 GHz.

d) The most difficult FM noise measurement problem is a low-power, low-noise source. The use of an injection phase-locked oscillator as a discriminator preamplifier (described in Section VII), has proved a viable solution for the problem.

e) Most of the problems concerning lack of repeatability from one laboratory to another can be resolved by understanding the baseband analysis equipment.

f) The art of a transmitter noise measurement is in the optimum use of various bandwidths in the wave analyzer.

g) Automation of the measurement is fun (and time consuming) for those doing the work.

h) Not many transmitter noise measurements per month are required to justify the cost of automation.

Two topics have not been discussed: first, the use of two oscillator techniques for phase modulation noise measurement, and second, pulsed-noise measurements. Two oscillator methods are well described by Shoaf, et al., [27] and apply more to time and frequency applications than to microwaves. It is believed that for submicrowave portions of transmitters, the transmission line discriminator is a more useful tool. No personal implementations have been made with regard to pulsed-transmitter noise measurements and nothing can be added to the work of Sann [19].

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Appendix. PROGRAM LISTINGS FOR THE HP-3045 AUTOMATED SPECTRUM ANALYZER

The information contained in this appendix provides a complete listing of all programs and subprograms required to make and completely document FM noise measurements in three different formats using the HP-3045A automated spectrum analyzer. The program listed in Table A-1 under the column designated LIN-LIN tabulates all the subroutines, including the calculator key that contains that subroutine, required to acquire and document an FM noise measurement. The vertical scale is calibrated in FM noise deviation (rms Hz per 1-Hz bandwidth)and the horizontal scale is calibrated in modulation frequency (kHz)with both scales plotted on a linear grid as shown in Figure A-1.

The program listed in Table A-1 under the column designated LOG-LIN tabulates all the subroutines required to acquire and document an FM noise measurement with the vertical scale calibrated in a linear format which is used to plot noise either in hertz deviation per 1-Hz bandwidth or in phase noise (dB WRT 1 mrad/1-Hz bandwidth)and the frequency or horizontal scale is calibrated logarithmically in hertz. The LOG-LIN grid plot is shown in Figure A-2.

The program listed in Table A-1 under the column designated LOG-LOG tabulates all the subroutines required to acquire and document FM noise measurements in logarithmic format for the FM noise magnitude and the modulation frequency ranges in kilohertz. The LOG-LOG grid plot is shown in Figure A-3.

Figure A-4 provides a sample of data plot format in the LIN-LIN mode. Notice that the program includes all the data conversion calculations for scaling measurement to an equivalent per 1-Hz bandwidth. The printed data, numbers and alphabet, are also put on the plot. These programs are all arranged in a question-answer format allowing scale selection, the equipment nomeclature, and the date the test was run. In addition, three types of data can be plotted:

1) The threshold level.
2) The raw noise measured data.
3) The calculated true noise corrected for threshold.

TABLE A-1. CALCULATOR PROGRAM STRUCTURES FOR MAKING FM NOISE MEASUREMENTS

Function Key Number*	LIN-LIN(1)	LOG-LIN(2)	LOG-LOG(3)
f_0	A-1	A-1	A-1
f_1	A-2	A-2	A-2
f_2	A-3	A-3	A-3
f_3	A-4	A-4	A-4
f_4	A-8	A-13	A-18
f_5	A-9	A-14	A-19
f_6	A-10	X	X
f_7	A-11	A-15	A-15
f_8	A-12	A-16	A-20
f_9	X	A-17	A-21
f_{10}	A-5	A-5	A-5
f_{12}	A-6	A-6	A-6
f_{13}	A-7	A-7	A-7

*See Figure 29

NOTES:

LIN-LIN(1) designates a program for FM noise measurement using linear parameters for voltage in the vertical dimension and frequency in the horizontal dimension.

LOG-LIN(2) designates a program for FM noise measurement using linear parameters for voltage or dBC in the vertical dimension and logarithmic parameters for frequency in the horizontal dimension.

LOG-LOG(3) designates a program for FM noise measurement using logarithmic parameters in the vertical dimension (voltage) and logarithmic parameters for frequency in the horizontal dimension.

X designates an unused program key for a particular program.

The quantities tabulated, i.e., A-1, A-2,...A-21 are identified in the appendix as List A-1 LOCAL (f_0), List A-2 REMOTE (f_1), etc. The list identifies the program on a step by step basis. A complete measurement program is contained in each column.

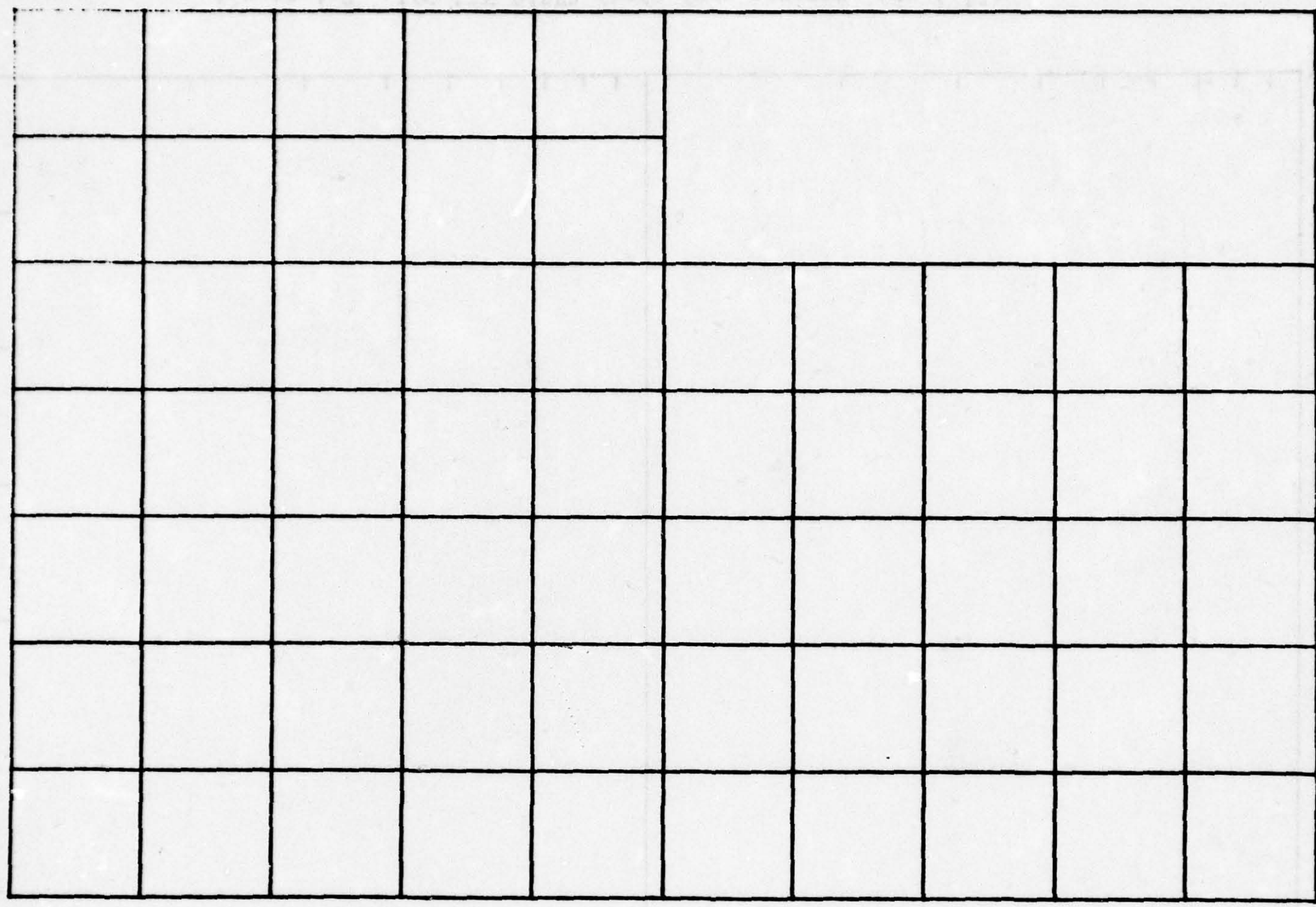

Figure A-1. LIN-LIN PAPER (made from program list A-10).

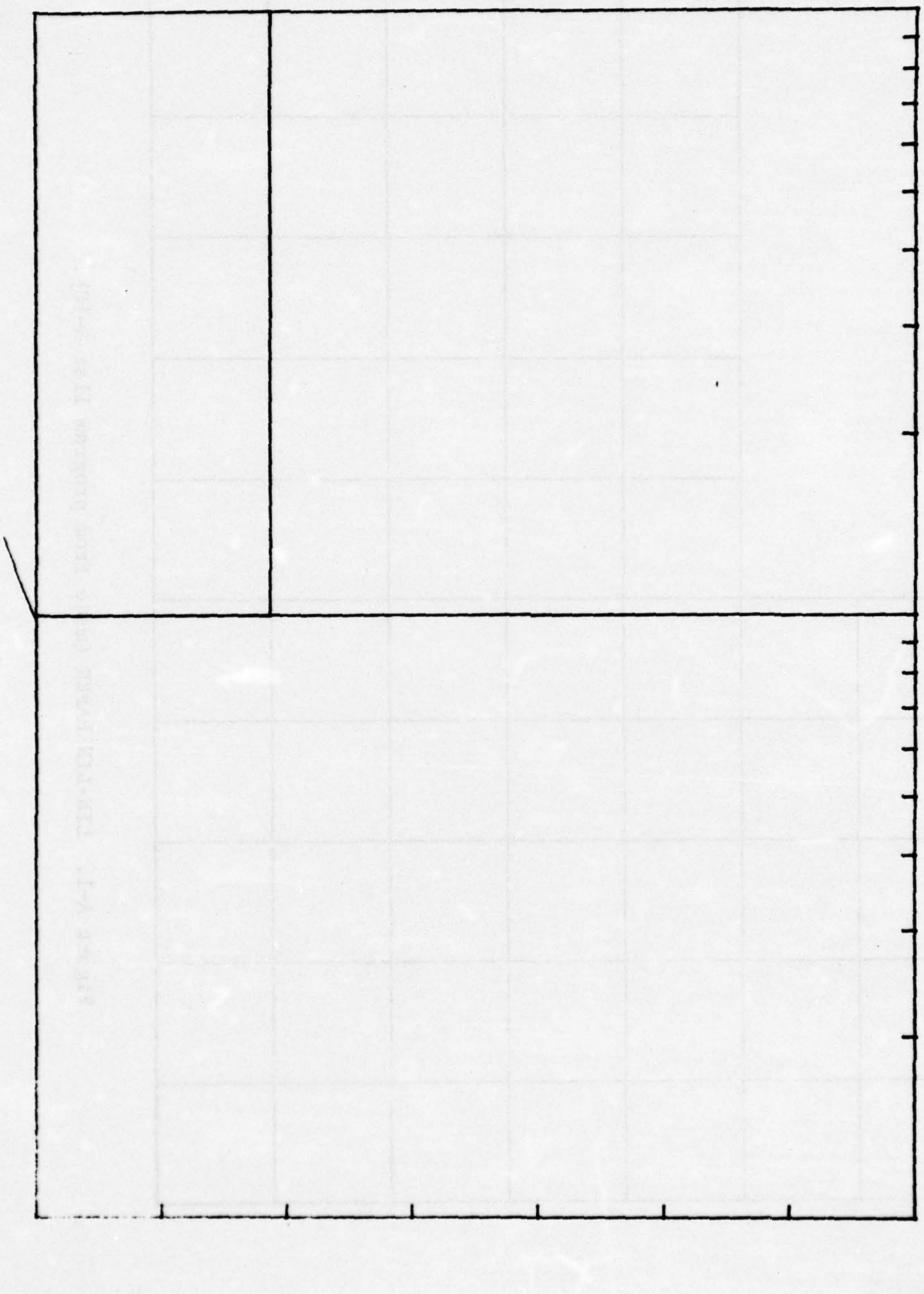

Figure A-2. LOG-LIN PAPER (made from program list A-17).

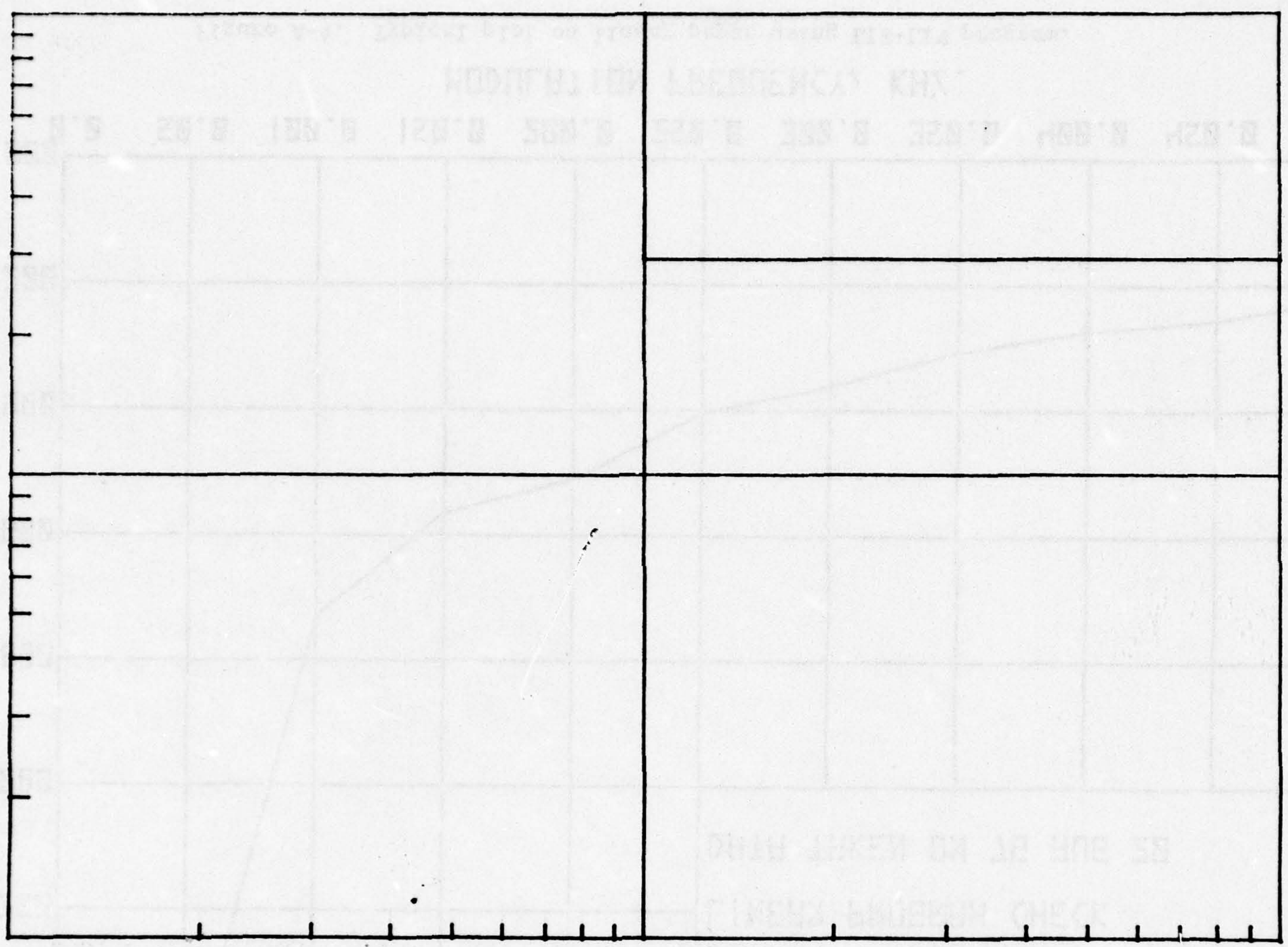

Figure A-3. LOG-LOG PAPER (made from program list A-21).

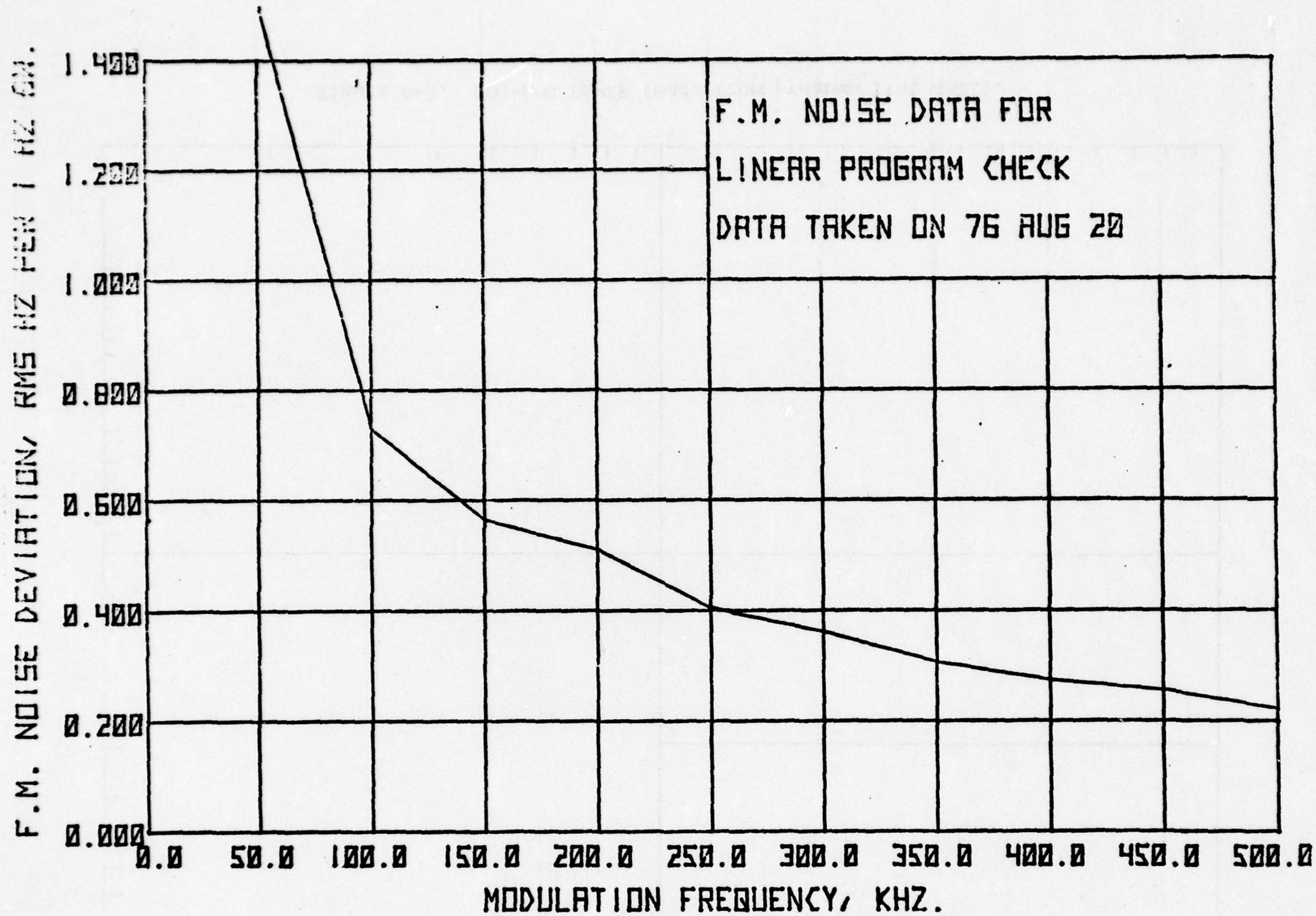

Figure A-4. Typical plot on linear paper using LIN-LIN program.

```
10 OUTPUT (13,20)1024;
20 FORMAT B
30 END
```

List A-1 LOCAL (f_0)

```
10 OUTPUT (13,20)768;
20 FORMAT B
30 END
```

List A-2 REMOTE (f_1)

```
10 DISP "FREQUENCY,BANDWIDTH";
20 INPUT F,B
30 GOSUB FNF(F) OF 30
60 END
```

List A-3 FREQ (f_2)

```
10 GOSUB FNM0 OF 10
20 FLOAT 2
30 DISP V
40 END
```

List A-4 MEAS (f_3)

```
10 FOR J=1 TO M+1
29 PRINT F[J],N[J]*1E+06,N[J]*,E+06/B[J]
50 NEXT J
190 END
```

List A-5 DIAGNOSTIC (f_{10})

```
10 DEF FNF(F)
20 CMD "?U#"
30 OUTPUT (13,40)F
40 FORMAT 'L',F1000.1,'='
50 CMD "?U1"
60 OUTPUT (13,70)"B",C
70 FORMAT F1000.0
80 RETURN 0
90 END
```

List A-6 DEFINED FUNCTION FNF(F) (f_{12})

```
10 DEF FNM(Z)
20 Z2=4
30 V=0
40 CMD "?U1","M1R1S1"
50 FOR Z1=1 TO Z2
55 WAIT 1E+03/B
60 CMD "?U1","T","?50"
70 ENTER (13,80)Z
80 FORMAT 1X,F7.2
90 V=V+10↑(Z/20)
100 NEXT Z1
110 V=V/Z2
120 RETURN V
130 END
```

List A-7 DEFINED FUNCTION FNM(Z) (f_{13})

```
10 OUTPUT (13,20)1024;
12 L0=0
20 FORMAT 8
30 DIM A$[3],B$[36],C$[36],TS[251],NS[251],FS[251],BS[251]
32 DIM D$[10],I$[30],CS[251]
50 DISP "WHAT IS TODAY'S DATE";
60 INPUT D$
70 DISP "WHAT SOURCE ARE U MEASURING";
80 INPUT I$
90 DISP "HAS THE SIGNAL SOURCE WARMED UP";
100 INPUT A$
110 IF A$="NO" THEN 90
120 DISP "IS THE REFERENCE POWER CORRECT";
130 INPUT A$
140 IF A$="Y" THEN 180
150 DISP "SET THE REFERENCE POWER, DIMBULB"
160 WAIT 5000
170 GOTO 120
180 DISP "IS THE REFERENCE PHASE CORRECT";
190 INPUT A$
200 IF A$="Y" THEN 230
210 DISP "TWIDDLE THE PHASE SHIFTER"
220 GOTO 180
230 DISP "HAVE YOU CHECKED INPUT OVERLOAD";
240 INPUT A$
250 IF A$="NO" THEN 90
260 OUTPUT (13,20)768;
270 DISP "WHAT IS CALIBRATE MOD FREQ";
280 INPUT A
290 C=2
300 B=30
310 F=A
320 GOSUB FNF(F) OF 320
330 GOSUB FNM(B) OF 330
340 DISP "HOW MANY DB DOWN ARE SIDEBANDS";
350 INPUT D
360 S=A*2*(10↑(-D/20))/(SQR(2)*V)
370 B$="THE DISCRIMINATOR SLOPE IS"
380 C$="PK HZ / MICROVOLT"
390 S7=INT(S*1E-03)/1000
400 PRINT B$,S7,C$
410 PRINT LIN(3)
420 DISP B$
430 WAIT 2000
440 DISP S7;C$
450 WAIT 5000
460 K1=8
470 DISP "CALIBRATION IS COMPLETE
472 END
```

List A-8 LIN-LIN CALIBRATE (f_4)

```
10 OUTPUT (13,30)1024;
20 L0=0
30 FORMAT B
40 IF K1=0 THEN 820
50 DISP "HAVE YOU CALIBRATED";
60 INPUT A$
70 IF A$="NO" THEN 820
80 DISP "IS THE CALIBRATE SIGNAL OFF";
90 INPUT A$
100 IF A$[1]="N" THEN 80
110 DISP "IS THE DISCRIMINATOR TUNED";
120 INPUT A$
130 IF A$="NO" THEN 110
140 DISP "IS THE THRESHOLD ATTENUATOR IN";
150 INPUT A$
160 IF A$="NO" THEN 140
170 DISP "HAVE U CHECKED INPUT OVERLOAD";
180 INPUT A$
190 IF A$[1]="N" THEN 170
200 OUTPUT (13,30)768;
210 L0=2
220 DISP "WHAT IS SWEEP START FREQUENCY";
230 INPUT F1
240 DISP "WHAT IS SWEEP STOP FREQUENCY";
250 INPUT F2
260 DISP "WHAT IS PLOT START FREQUENCY";
270 INPUT F9
280 P0=(F2-F9)/10
290 DISP "YOU CAN USE UP TO 250 STEPS."
300 WAIT 2000
310 DISP "HOW MANY FREQUENCY STEPS";
320 INPUT M
330 F=F1
340 F3=(F2-F1)/M
350 PRINT "FREQ","BW","VOLTS",LIN(1)
360 FOR J=1 TO M+1
370 GOSUB 480
380 F[J]=F
390 B[J]=B
400 C[J]=C
410 GOSUB FNF(7) OF 410
420 GOSUB FNM(0) OF 410
430 T[J]=V
440 F=F+F3
450 NEXT J
460 GOTO 840
470 GOTO 840
480 IF F<500 THEN 600
490 IF F<1250 THEN 630
500 IF F<2500 THEN 660
510 IF F<5000 THEN 690
```

```
520 IF F<25000 THEN 720
530 IF F<100000 THEN 750
540 IF F<510000 THEN 780
550 DISP "YOU HAVE ASKED FOR A FREQUENCY"
560 WAIT 1000
570 DISP "BEYOND THE RANGE OF THE DISCRIM."
580 WAIT 1000
590 GOTO 850
600 B=SQR(3.189)
610 C=0
620 GOTO 800
630 B=SQR(10.82)
640 C=1
650 GOTO 800
660 B=SQR(32.2331)
670 C=2
680 GOTO 800
690 B=SQR(107.7)
700 C=3
710 GOTO 800
720 B=SQR(288.96)
730 C=4
740 GOTO 800
750 B=SQR(1129.57)
760 C=5
770 GOTO 800
780 B=SQR(3456.9)
790 C=6
800 RETURN
810 GOTO 840
820 DISP "PRESS CALIBRATE"
830 GOTO 850
840 DISP "THRESHOLD DATA ARE COMPLETE."
850 END
```

List A-9 LIN-LIN THRESHOLD (f_5)

```
10 SCALE -1.2,10.3,-1.5,7.5
20 X=Y=0
30 PLOT X,Y,-2
40 Y=7
50 PLOT X,Y,-1
60 X=1
70 PLOT X,Y,-2
80 Y=0
90 PLOT X,Y,-1
100 X=2
110 PLOT X,Y,-2
120 Y=7
130 PLOT X,Y,-1
140 X=3
150 PLOT X,Y,-2
160 Y=0
170 PLOT X,Y,-1
180 X=4
190 PLOT X,Y,-2
200 Y=7
210 PLOT X,Y,-1
220 X=5
230 PLOT X,Y,-2
240 Y=0
250 PLOT X,Y,-1
260 X=6
270 PLOT X,Y,-2
280 Y=5
290 PLOT X,Y,-1
300 X=7
310 PLOT X,Y,-2
320 Y=0
330 PLOT X,Y,-1
340 X=8
350 PLOT X,Y,-2
360 Y=5
370 PLOT X,Y,-1
380 X=9
390 PLOT X,Y,-2
400 Y=0
410 PLOT X,Y,-1
420 X=10
430 PLOT X,Y,-2
440 Y=7
450 PLOT X,Y,-1
460 X=0
470 PLOT X,Y,2
480 PEN
490 Y=6
```

```
500 PLOT X,Y,-2
510 X=5
520 PLOT X,Y,-1
530 X=10
540 Y=5
550 PLOT X,Y,-2
560 X=0
570 PLOT X,Y,-1
580 Y=4
590 PLOT X,Y,-2
600 X=10
610 PLOT X,Y,-1
620 Y=3
630 PLOT X,Y,-2
640 X=0
650 PLOT X,Y,-1
660 Y=2
670 PLOT X,Y,-2
680 X=10
690 PLOT X,Y,-1
700 Y=1
710 PLOT X,Y,-2
720 X=0
730 PLOT X,Y,-1
740 Y=0
750 PLOT X,Y,-2
760 X=10
770 PLOT X,Y,-1
780 PEN
790 END
```

List A-10 LIN-LIN PAPER (f_6)

```
10 OUTPUT (13,20)1024;
12 IF L0=0 THEN 260
20 FORMAT B
30 PRINT LIN(3),"M = ";M
40 DISP "IS THRESHOLD ATTENUATOR OUT";
50 INPUT A$
60 IF A$="NO" THEN 30
70 DISP "HAVE U CHECKED INPUT OVERLOAD";
80 INPUT A$
90 IF A$[1]="N" THEN 70
100 OUTPUT (13,20)768;
120 M9=0
130 FOR J=1 TO M+1
140 F=F[J]
150 B=B[J]
160 C=C[J]
170 GOSUB FNF(F) OF 170
180 GOSUB FNM(0) OF 180
190 N[J]=V
200 K2=V/B
210 IF M9>K2 THEN 240
220 M9=K2
230 K=J
240 NEXT J
250 GOTO 310
260 FOR P=1 TO 3
270 DISP "YOU DIDN'T RUN THRESHOLD DATA."
280 WAIT 2000
290 NEXT P
300 GOTO 330
310 DISP " "
320 DISP "NOISE DATA ARE READY FOR PLOT."
330 END
```

List A-11 LIN-LIN NOISE (f_7)

```
10 V8=10↑(2.5/20)
20 PRINT LIN(3),"MAXIMUM NOISE INFORMATION",LIN(1)
30 PRINT "MODULATION","MEASUREMENT","F.M.NOISE","RANDOM OR"
40 PRINT "FREQUENCY","BANDWIDTH","RMS HZ/1 HZ BW"," COHERENT",LIN(1)
50 IF 1.2*N[K+1]/B[K+1]>N[K]/B[K] THEN 100
60 A$="COH"
70 N7=N[K]*K1
80 GOTO 120
90 PRINT
100 N7=N[K]*K1*V8/B[K]
110 A$="RDM"
120 PRINT F[K],B[K]↑2,N7,A$
130 PRINT LIN(2)
140 SCALE 0,(F2-F9)*11.5/10,0,9
150 DISP "IS LINEAR PAPER IN THE PLOTTER";
160 INPUT A$
170 IF A$[1]="N" THEN 150
180 P0=(F2-F9)/10
190 PEN
200 DISP "DO YOU WANT TO LABEL SCALES";
210 INPUT A$
220 IF A$[1]="N" THEN 560
230 DISP "VERTICAL SCALE MAXIMUM SHOULD BE"
240 WAIT 2000
250 DISP ".7, 1.4,0R 3.5 * A POWER OF 10."
260 WAIT 2000
270 DISP "WHAT IS VERTICAL SCALE MAXIMUM";
280 INPUT V7
290 OFFSET P0*1.2,0
300 PLOT 3*P0,0.8,1
310 LABEL (*,2,1.7,0,7/10)"MODULATION FREQUENCY, KHZ."
320 FOR J1=0 TO 10
330 PLOT (J1*P0-P0/1.9),1.2,1
340 IF F2<1001 THEN 420
350 IF F2<10001 THEN 390
360 LABEL (370,1.8,1.7,0,7/10)(J1*P0/1000+F9/1000);
370 FORMAT F6.1
380 GOTO 440
390 LABEL (400,1.8,1.7,0,7/10)(J1*P0/1000+F9/1000);
400 FORMAT F6.2
410 GOTO 440
420 LABEL (430,1.8,1.7,0,7/10)(J1*P0/1000+F9/1000);
430 FORMAT F6.3
440 NEXT J1
450 OFFSET +0,1.5
460 PLOT +0.2*P0,0,0
470 LABEL (*,2,1.7,PI/2,7/10)"F.M. NOISE DEVIATION, RMS HZ PER 1 HZ BW."
480 FOR J1=0 TO 7
490 PLOT 0.19*P0,(J1-3.1),1
```

```
500 IF V7>100 THEN 530
510 LABEL (610,1.8,1.7,0,7/10)V7*J1/7;
520 GOTO 540
530 LABEL (620,1.8,1.7,0,7/10)V7*J1/7;
540 NEXT J1
550 GOSUB 940
560 SCALE 0,(F2-F9)*11.5/10,0,V7*9/7
570 DISP "DO YOU WANT TO PLOT"
580 WAIT 1000
590 DISP "THRESHOLD(T),NOISE(N),COR'CTD(C)";
600 INPUT A$
610 FORMAT F7.3
620 FORMAT F7.1
630 F=F[1]
640 OFFSET P0*1.2,V7*1.5/7
650 IF A$="T" THEN 680
660 IF A$="N" THEN 700
670 GOTO 720
680 N7=T[1]*K1/B[1]
690 GOTO 730
700 N7=N[1]*K1/B[1]
710 GOTO 730
720 N7=SQR(ABS(N[1]↑2-T[1]↑2))*K1/(B[1])
730 PLOT F-F9,N7*V8,-2
740 WAIT 100
750 PEN
760 V5=V8
770 FOR J=2 TO M+1
780 IF A$="T" THEN 830
790 IF N[J]/B[J]<1.3*N[J-1]/B[J-1] THEN 810
800 V5=B[J]
810 IF A$="N" THEN 850
820 GOTO 870
830 N7=T[J]*K1/B[J]+V8
840 GOTO 890
850 N7=N[J]
860 GOTO 880
870 N7=SQR(ABS(N[J]↑2-T[J]↑2))
880 N7=N7*K1/B[J]+V5
890 PLOT F[J]-F9,N7,+2
900 V5=V8
910 NEXT J
920 PEN
930 GOTO 1020
940 SCALE -1.2,10.3,-1,7.5
950 PLOT 5.1,6.5,1
960 LABEL (*,2,1.7,0,7/10)"F.M. NOISE DATA FOR"
970 PLOT 5.1,6,1
980 LABEL (*,2,1.7,0,7/10)I$
```

```
990 PLOT 5.1,5.5,1
1000 LABEL (*,2,1.7,0,7/10)"DATA TAKEN ON ";D$
1010 RETURN
1020 DISP "FETCH PLOT,CONT560 TO PLOT AGAIN"
1030 GOTO 1050
1040 DISP "PRESS THRESHOLD"
1050 GOTO 1060
1060 END
```

List A-12 LIN-LIN PLOT (f_8)

```
10 COM F$[251],C$[251],B$[251],T$[251],N$[251],D$[14],I$[36],L0,K,K1,F5,D5,M
12 OUTPUT (13,30)1024;
20 L0=0
30 FORMAT B
40 DIM A$[3],B$[36],C$[36],T$[251],N$[251],F$[251],B$[251]
50 DIM D$[10],I$[30],C$[251]
60 DISP "WHAT IS TODAY'S DATE";
70 INPUT D$
80 DISP "WHAT SOURCE ARE U MEASURING";
90 INPUT I$
100 DISP "HAS THE SIGNAL SOURCE WARMED UP";
110 INPUT A$
120 IF A$="NO" THEN 100
130 DISP "IS THE REFERENCE POWER CORRECT";
140 INPUT A$
150 IF A$="Y" THEN 190
160 DISP "SET THE REFERENCE POWER, DIMBULB"
170 WAIT 5000
180 GOTO 130
190 DISP "IS THE REFERENCE PHASE CORRECT";
200 INPUT A$
210 IF A$="Y" THEN 240
220 DISP "TWIDDLE THE PHASE SHIFTER"
230 GOTO 190
240 DISP "HAVE YOU CHECKED INPUT OVERLOAD";
250 INPUT A$
260 IF A$="NO" THEN 100
270 OUTPUT (13,30)768;
280 DISP "WHAT IS CALIBRATE MOD FREQ";
290 INPUT A
300 C=2
310 B=30
320 F=A
330 GOSUB FNF(F) OF 330
340 GOSUB FNM(0) OF 340
350 DISP "HOW MANY DB DOWN ARE SIDEBANDS";
360 INPUT D
370 S=A*2*(10↑(-D/20))/(SQR(2)*V)
380 B$="THE DISCRIMINATOR SLOPE IS"
390 C$="PK HZ / MICROVOLT"
400 S7=INT(S*1E-03)/1000
410 PRINT B$,S7,C$
420 PRINT LIN(3)
430 DISP B$
440 WAIT 2000
450 DISP S7;C$
460 WAIT 5000
470 K1=S
480 DISP "CALIBRATION IS COMPLETE"
490 END
```

List A-13 LOG-LIN CALIBRATE (f_4)

```
10 COM FS[251],CS[251],BS[251],TS[251],NS[251],D$[14],I$[36],L0,K,K1,F5,D5,M
12 OUTPUT (13,30)1024;
20 L0=0
30 FORMAT B
40 IF K1=0 THEN 750
50 DISP "HAVE YOU CALIBRATED";
60 INPUT A$
70 IF A$="NO" THEN 750
80 DISP "IS THE CALIBRATE SIGNAL OFF";
90 INPUT A$
100 IF A$[1]="N" THEN 80
110 DISP "IS THE DISCRIMINATOR TUNED";
120 INPUT A$
130 IF A$="NO" THEN 110
140 DISP "IS THE THRESHOLD ATTENUATOR IN";
150 INPUT A$
160 IF A$="NO" THEN 140
170 DISP "HAVE U CHECKED INPUT OVERLOAD";
180 INPUT A$
190 IF A$[1]="N" THEN 170
200 OUTPUT (13,30)768;
210 L0=1
220 DISP "WHAT IS THE START FREQUENCY";
230 INPUT F5
240 DISP "HOW MANY DECADES DO YOU WANT";
250 INPUT D5
260 DISP "HOW MANY POINTS DO YOU WANT";
270 INPUT M
280 FOR J=1 TO M+1
290 F=F5*10↑((J-1)*D5/M)
300 GOSUB 410
320 C[J]=C
330 F[J]=F
340 B[J]=B
350 GOSUB FNF(F) OF 350
360 GOSUB FNM(0) OF 360
370 T[J]=V
390 NEXT J
400 GOTO 770
410 IF F<500 THEN 530
420 IF F<1250 THEN 560
430 IF F<2500 THEN 590
440 IF F<5000 THEN 620
450 IF F<25000 THEN 650
460 IF F<100000 THEN 680
470 IF F<510000 THEN 710
480 DISP "YOU HAVE ASKED FOR A FREQUENCY"
490 WAIT 1000
500 DISP "BEYOND THE RANGE OF THE DISCRIM."
```

```
510 WAIT 1003
520 GOTO 780
530 B=SQR(3.189)
540 C=0
550 GOTO 730
560 B=SQR(10.82)
570 C=1
580 GOTO 730
590 B=SQR(32.2331)
600 C=2
610 GOTO 730
620 B=SQR(107.7)
630 C=3
640 GOTO 730
650 B=SQR(288.96)
660 C=4
670 GOTO 730
680 B=SQR(1129.57)
690 C=5
700 GOTO 730
710 B=SQR(3456.9)
720 C=6
730 RETURN
740 GOTO 770
750 DISP "PRESS CALIBRATE"
760 GOTO 780
770 DISP "THRESHOLD DATA ARE COMPLETE."
780 END
```

List A-14 LOG-LIN THRESHOLD (f_5)

```
10 COM FS[251],CS[251],BS[251],TS[251],NS[251],D$[14],I$[36],L0,K,K1,F5,D5,M
11 OUTPUT (13,20)1024;
12 IF L0=0 THEN 260
20 FORMAT B
30 PRINT LIN(3),"M = ";M
40 DISP "IS THRESHOLD ATTENUATOR OUT";
50 INPUT A$
60 IF A$="NO" THEN 30
70 DISP "HAVE U CHECKED INPUT OVERLOAD";
80 INPUT A$
90 IF A$[1]="N" THEN 70
100 OUTPUT (13,20)768;
120 M9=0
130 FOR J=1 TO M+1
140 F=F[J]
150 B=B[J]
160 C=C[J]
170 GOSUB FNF(F) OF 170
180 GOSUB FNM(0) OF 180
190 N[J]=V
200 K2=V/B
210 IF M9>K2 THEN 240
220 M9=K2
230 K=J
240 NEXT J
250 GOTO 310
260 FOR P=1 TO 3
270 DISP "YOU DIDN'T RUN THRESHOLD DATA."
280 WAIT 2000
290 NEXT P
300 GOTO 330
310 DISP " "
320 DISP "NOISE DATA ARE READY FOR PLOT."
330 END
```

List A-15 LOG NOISE (f_7)

```
16 COM F$[25],C$[25],E$[5.],T$[25],H$[25],D$[14],I$[36],L9,K,K1,F5,I5,M
17 V8=10↑(2.5/20)
310 FORMAT F7.1
900 PRINT LIN(3),"  START"," DB W.R.T."
904 PRINT "FREQUENCY","ONE MILLIRADIAN",LIN(1)
909 P9=20*LGT(N[1]*K1*V8*1000/F[1]/B[1])
910 PRINT F[1],P9,LIN(3)
920 DISP "DO YOU WANT TO LABEL SCALES";
930 INPUT A$
940 IF A$[1]="N" THEN 1140
950 DISP "WHAT IS VERT SCALE MAX DB";
970 INPUT P7
980 GOSUB 2000
998 SCALE -1.25,10.25,-1.25,7.75
1000 PLOT -0.85,0,0
1010 LABEL (*,2,1.7,PI/2,7/10)"PHASE NOISE, DB W.R.T. 1 MILLIRAD/1 HZ BW."
1012 SCALE -1.25,10.25,-8.75,0.25
1020 X=-0.85
1030 Z=0
1050 FOR Y=0 TO -7.5 STEP -7.5/7
1060 PLOT X,(Y-0.085),1
1070 LABEL (1080,1.8,1.7,0,7/10)P7+10*Z
1080 FORMAT F5.0
1090 Z=Z-1
1100 NEXT Y
1120 GOSUB 1800
1130 SCALE -1.25,10.25,-8.75,0.25
1140 DISP "DO YOU WANT TO PLOT"
1142 WAIT 1000
1144 DISP "THRESHOLD(T),NOISE(N),COR'CTD(C)";
1150 INPUT A$
1160 J=1
1180 GOSUB 2200
1182 P9=(-P7+20*LGT(N7*1000/F[J]))*7.5/70
1184 IF P9>-7.5 THEN 1190
1186 P9=-7.5
1190 PLOT (J-1)/M*10,P9,-2
1200 FOR J=2 TO M+1
1210 GOSUB 2200
1212 P9=(-P7+20*LGT(N7*1000/F[J]))*7.5/70
1214 IF P9>-7.5 THEN 1220
1216 P9=-7.5
1220 PLOT (J-1)/M*10,P9,2
1230 NEXT J
1240 PEN
1249 DISP "FETCH PLOT  CONT 1130 TO PLOT AGAIN."
1250 GOTO 5000
1790 GOTO 1800
1800 SCALE -1.2,10.3,-.. 5
```

```
1810 PLOT 5.3,6.8,1
1820 LABEL (*,2,1.7,0,7/10)"F.M. NOISE DATA FOR"
1830 PLOT 5.3,6.3,1
1840 LABEL (*,2,1.7,0,7/10)I$
1850 PLOT 5.3,5.8,1
1860 LABEL (*,2,1.7,0,7/10)"DATA TAKEN ON ";D$
1870 RETURN
1880 DISP "FETCH PLOT,CONT410 TO PLOT AGAIN"
1890 GOTO 1920
1910 DISP "PRESS THRESHOLD"
1920 GOTO 5000
2000 SCALE -1.25,10.25,0,9
2010 PLOT 3,0.65,1
2030 LABEL (*,2,1.7,0,7/10)"MODULATION FREQUENCY, KHZ."
2040 Y=0.95
2042 Z=0
2050 FOR X=0 TO 10 STEP 10/D5
2060 PLOT (X-0.73),Y,1
2070 LABEL (310,1.8,1.7,0,7/10)F[1]*(10↑Z)/1000;
2072 Z=Z+1
2080 NEXT X
2100 RETURN
2198 GOTO 5000
2200 V5=V8
2208 IF A$="T" THEN 2260
2210 IF J=1 THEN 2260
2220 IF N[J]/B[J]<1.3*N[J-1]/B[J-1] THEN 2260
2230 V5=B[J]
2260 IF A$="T" THEN 2300
2270 IF A$="N" THEN 2340
2276 N7=K1*V5*SQR(ABS(N[J]↑2-T[J]↑2))/B[J]
2280 GOTO 2400
2300 N7=K1*V8*T[J]/B[J]
2310 GOTO 2400
2340 N7=K1*V5*N[J]/B[J]
2400 RETURN
5000 END
```

List A-16 LOG-LIN PLOT (f_8)

AD-A036 389 ARMY MISSILE RESEARCH DEVELOPMENT AND ENGINEERING LAB--ETC F/G 9/3
MEASUREMENT OF TRANSMITTER NOISE AT HIGH FREQUENCY THROUGH MICR--ETC(U)
SEP 76 T A BARLEY, G J RAST, J R ASHLEY
UNCLASSIFIED RE-7T-8 NL

2 OF 2
ADA036389

END
DATE
FILMED
3-77

```
100 SCALE -1.25,10.25,-1.25,7.75
120 Y=0
130 DISP "HOW MANY DECADES HORIZONTALLY";
140 INPUT H1
144 K5=10/(LOG(10)*H1)
200 FOR I=0 TO H1
202 Y9=7.5
210 C=10↑I
220 X=K5*LOG(C)
230 PLOT X,Y,0
232 IF X>9.999 THEN 250
234 IF X<5 THEN 250
236 Y9=5.5
250 FOR Y=0 TO Y9 STEP Y9
260 PLOT X,Y,2
280 NEXT Y
290 PEN
302 Y=0
306 IF I=H1 THEN 480
310 FOR M1=2 TO 9
320 M2=M1*C
330 X=K5*LOG(M2)
340 FOR Y=0 TO 0.1 STEP 0.1
350 PLOT X,Y,-2
370 NEXT Y
380 Y=0
390 PEN
410 NEXT M1
460 NEXT I
480 Y=0
490 PLOT X,Y,-2
492 X=0
494 PLOT X,Y,-1
500 PEN
510 FOR Y=7.5/7 TO 6*7.5/7 STEP 7.5/7
512 X=0
520 PLOT X,Y,-2
530 X=0.1
540 PLOT X,Y,-1
550 NEXT Y
570 Y=7.5
580 X=0
590 PLOT X,Y,-1
600 X=10
610 PLOT X,Y,2
620 PEN
630 Y=5.5
640 PLOT X,Y,0
650 X=5
```

```
670 PLOT X,Y,2
680 Y=7.5
690 PLOT X,Y,-1
700 PEN
750 END
```

List A-17 LOG-LIN PAPER (f_9)

```
10 COM FS[251],CS[251],BS[251],TS[251],NS[251],D$[14],I$[36],L0,K,K1,F5,D5,M
20 DIM A$[3],B$[36],C$[36]
30 FORMAT B
40 OUTPUT (13,30)1024;-
50 L0=0
60 DISP "WHAT IS TODAY'S DATE";
70 INPUT D$
80 DISP "WHAT SOURCE ARE U MEASURING";
90 INPUT I$
100 DISP "HAS THE SIGNAL SOURCE WARMED UP";
110 INPUT A$
120 IF A$="NO" THEN 100
130 DISP "IS THE REFERENCE POWER CORRECT";
140 INPUT A$
150 IF A$="Y" THEN 190
160 DISP "SET THE REFERENCE POWER, DIMBULB"
170 WAIT 5000
180 GOTO 130
190 DISP "IS THE REFERENCE PHASE CORRECT";
200 INPUT A$
210 IF A$="Y" THEN 240
220 DISP "TWIDDLE THE PHASE SHIFTER"
230 GOTC 190
240 DISP "HAVE YOU CHECKED INPUT OVERLOAD";
250 INPUT A$
260 IF A$="NO" THEN 100
270 OUTPUT (13,30)768;
280 DISP "WHAT IS CALIBRATE MOD FREQ";
290 INPUT A
300 C=2
310 B=30
320 F=A
330 GOSUB FNF(F) OF 330
340 GOSUB FNM(0) OF 340
350 DISP "HOW MANY DB DOWN ARE SIDEBANDS";
360 INPUT D
370 S=A*2*(10↑(-D/20))/(SQR(2)*V)
380 B$="THE DISCRIMINATOR SLOPE IS"
390 C$="PK HZ / MICROVOLT"
400 S7=INT(S*1E-03)/1000
410 PRINT B$,S7,C$
420 PRINT LIN(3)
430 DISP B$
440 WAIT 2000
450 DISP S7;C$
460 WAIT 5000
470 K1=S
480 DISP "CALIBRATION IS COMPLETE"
490 END
```

List A-18 LOG-LOG CALIBRATE (f_4)

```
10 COM FS[251],CS[251],BS[251],TS[251],NS[251],D$[14],I$[36],L0,K,K1,F5,D5,M
12 OUTPUT (13,30)1024;
20 L0=0
30 FORMAT B
40 IF K1=0 THEN 750
50 DISP "HAVE YOU CALIBRATED";
60 INPUT A$
70 IF A$="NO" THEN 750
80 DISP "IS THE CALIBRATE SIGNAL OFF";
90 INPUT A$
100 IF A$[1]="N" THEN 80
110 DISP "IS THE DISCRIMINATOR TUNED";
120 INPUT A$
130 IF A$="NO" THEN 110
140 DISP "IS THE THRESHOLD ATTENUATOR IN";
150 INPUT A$
160 IF A$="NO" THEN 140
170 DISP "HAVE U CHECKED INPUT OVERLOAD";
180 INPUT A$
190 IF A$[1]="N" THEN 170
200 OUTPUT (13,30)768;
210 L0=1
220 DISP "WHAT IS THE START FREQUENCY";
230 INPUT F5
240 DISP "HOW MANY DECADES DO YOU WANT";
250 INPUT D5
260 DISP "HOW MANY POINTS DO YOU WANT";
270 INPUT M
280 FOR J=1 TO M+1
290 F=F5*10↑((J-1)*D5/M)
300 GOSUB 410
320 C[J]=C
330 F[J]=F
340 B[J]=B
350 GOSUB FNF(F) OF 350
360 GOSUB FNM(0) OF 360
370 T[J]=V
390 NEXT J
400 GOTO 770
410 IF F<500 THEN 530
420 IF F<1250 THEN 560
430 IF F<2500 THEN 590
440 IF F<5000 THEN 620
450 IF F<25000 THEN 650
460 IF F<100000 THEN 680
470 IF F<1.1E+06 THEN 710
480 DISP "YOU HAVE ASKED FOR A FREQUENCY"
490 WAIT 1000
500 DISP "BEYOND THE RANGE OF THE DISCRIM."
```

```
510 WAIT 1000
520 GOTO 780
530 B=SQR(3.189)
540 C=0
550 GOTO 730
560 B=SQR(10.82)
570 C=1
580 GOTO 730
590 B=SQR(32.2331)
600 C=2
610 GOTO 730
620 B=SQR(107.7)
630 C=3
640 GOTO 730
650 B=SQR(288.96)
660 C=4
670 GOTO 730
680 B=SQR(1129.57)
690 C=5
700 GOTO 730
710 B=SQR(3456.9)
720 C=6
730 RETURN
740 GOTO 770
750 DISP "PRESS CALIBRATE'
760 GOTO 780
770 DISP "THRESHOLD DATA ARE COMPLETE."
780 END
```

List A-19 LOG-LOG THRESHOLD (f_5)

```
10 COM FS[251],CS[251],BS[251],TS[251],NS[251],D$[14],I$[36],L0,K,K1,F5,D5,M
14 PRINT LIN(3),"MAXIMUM NOISE INFORMATION",LIN(1)
17 V8=10↑(2.5/20)
20 PRINT "MODULATION","MEASUREMENT","F.M.NOISE","RANDOM OR"
30 PRINT "FREQUENCY","BANDWIDTH","RMS HZ/1 HZ BW"," COHERENT",LIN(1)
32 IF 1.2*N[K+1]/B[K+1]>N[K]/B[K] THEN 48
34 A$="COH"
36 N7=N[K]*K1
38 GOTO 50
40 PRINT
48 N7=N[K]*K1*V8/B[K]
49 A$="RDM"
50 PRINT F[K],B[K]↑2,N7,A$
60 PRINT LIN(2)
310 FORMAT F6.1
800 DISP "HOW MANY VERTICAL DECADES";
810 INPUT H2
840 DISP "DO YOU WANT FREQ(F) OR PHASE(P)";
850 INPUT A$
870 IF A$="F" THEN 1300
880 IF A$="P" THEN 900
890 GOTO 840
900 PRINT LIN(3),"START FREQUENCY","MILLIRADIANS PHASE DEV.",LIN(1)
910 PRINT F[1],N[1]*K1*10↑(2.5/20)*1000/(F[1]*B[1])
920 DISP "DO YOU WANT TO LABEL SCALES";
930 INPUT A$
940 IF A$[1]="N" THEN 1140
950 DISP "WHAT IS VERT SCALE MAX MILLIRAD";
970 INPUT P7
980 GOSUB 2000
998 SCALE -1.25,10.25,-1.25,7.75
1000 PLOT -0.85,0,0
1010 LABEL (*,2,1.7,PI/2,7/10)"PHASE NOISE, RMS MILLIRADIANS PER 1 HZ BW."
1020 X=-1.01
1030 Z=0
1050 FOR Y=7.5 TO 0 STEP -7.5/H2
1060 PLOT X,(Y-0.085),1
1070 LABEL (1080,1.8,1.7,0,7/10)P7*10↑Z
1080 FORMAT F7.3
1090 Z=Z-1
1100 NEXT Y
1104 P7=P7/1000
1120 GOSUB 1300
1130 SCALE -1.25,10.25,-8.75,0.25
1132 L6=7.5/LOG(10)/H2
1140 DISP "DO YOU WANT TO PLOT"
1142 WAIT 1000
1144 DISP "THRESHOLD(T),NOISE(N),COR'CTD(C)";
1150 INPUT A$
1160 J=1
1180 GOSUB 2200
```

```
1182 P9=K6*LOG(N7/P7/F[J])
1184 IF P9>-7.5 THEN 1190
1186 P9=-7.5
1190 PLOT (J-1)/M*10,P9,-2
1200 FOR J=2 TO M+1
1210 GOSUB 2200
1212 P9=K6*LOG(N7/P7/F[J])
1214 IF P9>-7.5 THEN 1220
1216 P9=-7.5
1220 PLOT (J-1)/M*10,P9,2
1230 NEXT J
1240 PEN
1248 PRINT "FETCH PLOT, CONT1130 TO PLOT AGAIN.",LIN(2)
1250 GOTO 5000
1300 REM FM NOISE PLOTTING WILL START HERE.
1310 DISP "DO YOU WANT TO LABEL SCALES";
1320 INPUT A$
1340 IF A$[1]="N" THEN 1540
1350 DISP "WHAT IS VERT SCALE MAX HERTZ";
1370 INPUT P7
1380 GOSUB 2000
1398 SCALE -1.25,10.25,-1.25,7.75
1400 PLOT -0.85,0,0
1410 LABEL (*,2,1.7,PI/2,7/10)"F.M. NOISE DEVIATION RMS HZ PER 1 HZ BW"
1420 X=-1.01
1430 Z=0
1450 FOR Y=7.5 TO 0 STEP -7.5/H2
1460 PLOT X,(Y-0.085),1
1470 LABEL (1080,1.8,1.7,0,7/10)P7*10^Z
1480 FORMAT F7.3
1490 Z=Z-1
1500 NEXT Y
1520 GOSUB 1800
1530 SCALE -1.25,10.25,-8.75,0.25
1532 K6=7.5/LOG(10)/H2
1540 DISP "DO YOU WANT TO PLOT"
1542 WAIT 1000
1544 DISP "THRESHOLD(T),NOISE(N),COR'CTD(C)";
1550 INPUT A$
1560 J=1
1580 GOSUB 2200
1582 P9=K6*LOG(N7/P7)
1584 IF P9>-7.5 THEN 1590
1586 P9=-7.5
1590 PLOT (J-1)/M*10,P9,-2
1600 FOR J=2 TO M+1
1610 GOSUB 2200
1612 P9=K6*LOG(N7/P7)
1614 IF P9>-7.5 THEN 1620
1616 P9=-7.5
```

```
1620 PLOT [illegible]-10,F9,2
1630 [illegible]
1640 PEN
1643 DISP "FETCH PLOT, CONT1530 TO PLOT AGAIN."
1650 GOTO 5000
1790 GOTO 1890
1800 SCALE -1.2,10.3,-1,7.5
1810 PLOT 5.3,6.8,1
1820 LABEL (*,2,1.7,0,7/10)"F.M. NOISE DATA FOR"
1830 PLOT 5.3,6.3,1
1840 LABEL (*,2,1.7,0,7/10)I$
1850 PLOT 5.3,5.8,1
1860 LABEL (*,2,1.7,0,7/10)"DATA TAKEN ON ";D$
1870 RETURN
1880 DISP "FETCH PLOT,CONT410 TO PLOT AGAIN"
1890 GOTO 1920
1910 DISP "PRESS THRESHOLD"
1920 GOTO 5000
2000 SCALE -1.25,10.25,0,9
2010 PLOT 3,0.65,1
2030 LABEL (*,2,1.7,0,7/10)"MODULATION FREQUENCY, KHZ."
2040 Y=0.95
2042 Z=0
2050 FOR X=0 TO 10 STEP 10/D5
2060 PLOT (X-0.6),Y,1
2070 LABEL (310,1.8,1.7,0,7/10)F[1]*(10↑Z)/1000;
2072 Z=Z+1
2080 NEXT X
2100 RETURN
2199 GOTO 5000
2200 V5=V9
2203 IF A$="T" THEN 2260
2210 IF J=1 THEN 2260
2220 IF N[J]/B[J]<1.3*N[J-1]/B[J-1] THEN 2260
2230 V5=B[J]
2260 IF A$="T" THEN 2300
2270 IF A$="N" THEN 2340
2275 N7=[illegible]*V5*SQR(ABS(N[J]↑2-T[J]↑2))/B[J]
2280 GOTO 2400
2300 N7=[illegible]*V8*T[J]/B[J]
2310 GOTO 2400
2340 N7=[illegible]*V5*N[J]/B[J]
2400 RETURN
5000 END
```

List A-20 LOG-LOG PLOT (f_8)

```
10 SCALE -1.25,10.25,-1.25,7.75
20 Y=0
30 DISP "HOW MANY DECADES HORIZONTALLY";
40 INPUT H1
50 K5=10/(LOG(10)*H1)
60 DISP "HOW MANY DECADES VERTICALLY";
70 INPUT H2
80 K6=7.5/(LOG(10)*H2)
90 FOR I=0 TO H1
100 C=10↑I
110 X=K5*LOG(C)
120 PLOT X,Y,0
130 FOR Y=0 TO 7.5 STEP 7.5
140 IF X<5 THEN 170
150 IF Y<5 THEN 170
160 Y=5.5
170 PLOT X,Y,2
180 NEXT Y
190 PEN
200 Y=0
210 IF I=H1 THEN 320
220 FOR M0=2 TO 9
230 M2=M0*C
240 X=K5*LOG(M2)
250 FOR Y=0 TO 0.1 STEP 0.1
260 PLOT X,Y,-2
270 NEXT Y
280 Y=0
290 PEN
300 NEXT M0
310 NEXT I
320 X=5
330 FOR Y=5.5 TO 7.5 STEP 2
340 PLOT X,Y,2
350 NEXT Y
360 PEN
370 X=10
380 Y=5.5
390 PLOT X,Y,-2
400 Y=7.5
410 PLOT X,Y,2
420 PEN
430 X=Y=0
440 FOR I=0 TO H2
450 C=10↑I
460 Y=K6*LOG(C)
470 PLOT X,Y,0
480 FOR X=0 TO 10 STEP 10
490 IF Y<5.5 THEN 520
```

```
500 IF X .5 THEN 520
510 Y=5
520 PLOT X,Y,-2
530 NEXT X
540 PEN
550 X=0
560 IF I=H2 THEN 680
570 FOR M0=2 TO 9
580 M2=M0*C
590 Y=K6*LOG(M2)
600 FOR X=0 TO 0.15 STEP 0.15
610 PLOT X,Y,-2
620 NEXT X
630 PEN
640 X=0
650 PLOT X,Y,0
660 NEXT M0
670 NEXT I
680 PEN
690 X=5
700 Y=7.5
710 PLOT X,Y,-2
720 X=10
730 PLOT X,Y,2
740 PEN
750 END
```

List A-21 LOG-LOG PAPER (f_9)

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