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Semiconductor Measurement Technology:

Modulation Measurements for Microwave Mixers

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James M. Kenney *

Center for Electronics and Electrical Engineering
National Engineering Laboratory
National Bureau of Standards
Washington, D.C. 20234

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Semiconductor Measurement Technology: Modulation Measurements for Microwave Mixers

James M. Kenney*
Center for Electronics and Electrical Engineering
National Bureau of Standards
Washington, D.C. 20234

The measurement of mixer conversion loss using periodic or incremental modulation of the local oscillator, and the evaluation and minimization of the associated systematic and random uncertainties, are discussed in terms of an X-band mixer measurement system constructed at NBS. It is shown that the systematic uncertainty in the incremental modulation method of measuring conversion loss results largely from the uncertainties in the calibration of microwave attenuation and power.

It is also shown that the "modulation" (periodic modulation) and "incremental" (incremental modulation) methods of measuring conversion loss are essentially identical, the only practical distinction being in the somewhat different instrumentation required by the different modulation rates.

Several improvements in the periodic and incremental modulation techniques are introduced. Novel circuits for measuring intermediate-frequency output conductance and local-oscillator return loss are described which may also be useful for other immittance measurements.

Key words: Conversion loss; diode measurement; intermediate-frequency (i-f) output conductance; measurement uncertainties; microwave mixer diodes; modulation; point-contact diodes; reflectometer; return loss; Schottky-barrier diodes; semiconductor diodes; standing-wave ratio (SWR).

1. Introduction

The manufacturers of microwave mixer diodes currently lack easily traceable standards with which to calibrate their production line measurements. There are no standards laboratories to provide diode calibrations, as were once maintained by the Department of Defense.

There are no specified measurement tolerances, and no allowance is made for measurement uncertainty at the specification limits as required by MIL-STD-750B (4.2.3), which also requires an annual calibration which is not being performed at present. As a result there are large uncertainties in the absolute values of the measured parameters, even though manufacturers attempt to maintain a uniform product by comparison with groups of diodes measured in previous years. These uncertainties have, in turn, frequently resulted in appreciable differences in the average quality of diodes made by different manufacturers to the same specification limits.

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* Now with Polymer Science and Standards Division, Center for Materials Science, National Measurement Laboratory, National Bureau of Standards.

As a consequence of these differences, competitive bidding based only on price cannot be conducted fairly, and inferior diodes are being purchased for replacement use.

Compounding these problems is a lack of detailed descriptions of modern measurement methods in the literature, and an influx of new diode producers unacquainted with traditional orally transmitted practices. In addition, very rarely is there any incoming inspection by the Department of Defense and most other users, who are even more handicapped by the lack of standards.

Microwave mixer measurements have traditionally been regarded as something of a "black art," presenting great difficulties in obtaining reproducible or even repeatable results, and with reputed systematic differences among competitive measurement techniques. Unfortunately, there has also been a tendency for the original diode manufacturers to regard their measurement techniques as proprietary, with no consideration made for the possibility of remeasurement by diode users, thus making it difficult to establish common standards. Manufacturers are understandably reluctant to accept new standards that may affect yield or force revisions of specification limits. The standard (MIL-spec) point-contact diodes, which still dominate the mixer field, have been reduced in price to the extent that there is little incentive for industry to improve measurements. Perhaps as a consequence of this experience, there has been little effort at standardizing Schottky-barrier types, although many are made with identical specifications.

The primary objectives of this program were to refine and evaluate existing and new microwave mixer measurement methods so that satisfactory accuracy and repeatability could be obtained. The program also sought to establish procedures for determining the accuracy and repeatability of these measurement methods.

1.1 Background

An important phase in the history of semiconductor electronics was the revival of the point-contact semiconductor "crystal detector" for microwave radar receivers long after it had been discarded in favor of the vacuum tube for lower frequency reception. Two types of microwave diodes were developed for radar use: one for use in the heterodyne conversion of microwave signals to a lower intermediate frequency, commonly called "mixing" (not to be confused with the linear addition of signals at audio frequencies which is also called "mixing"), and the other for use in direct demodulation of the microwave signals, commonly called "video detection." The physical construction of these two types is quite similar, the only major difference being in the different methods of testing for these applications. Mixer diodes have been by far the more commonly used and have therefore received the greatest attention in the development of improved types. Their high usage has also resulted in a relatively low unit cost. For these reasons it is common for mixer types to be used in detector applications even though they have not been evaluated for this use. Schottky-barrier diodes are now being made for use as microwave mixers and detectors, but the much lower prices of point-contact diodes and the large amount of existing equipment designed for their use have supported the continued large production of the latter. The measurement methods devised for point-contact diodes are equally applicable to Schottky diodes, since the measured parameters are those of the entire mixer: the diode in a holder (mount) with local oscillator drive. If the holder and local oscillator meet the appropriately chosen specifications, the mixer characteristics can be attributed to the diode.

1.2 Mixer Parameters

Although the performance of a mixer at high signal levels (e.g., in terms of its intermodulation distortion) is of some interest, the most important characteristics of a mixer are generally those concerned with its ability to heterodyne very low-level signals with a minimum degradation of signal-to-noise ratio. As with any low-gain multiport network, a mixer can degrade signal-to-noise ratio in three ways:

(1) By attenuating the signal (or merely by failing to amplify it) so that the noise added by following stages (the intermediate-frequency amplifier) is significantly large by comparison with the noise accompanying the signal. The parameter expressing this gain factor is known as "conversion loss." (Note: Precise definitions of mixer parameters are given in Appendix D.)

(2) By adding noise originated by the mixer. This added noise has two basic components: (a) the "essential" (thermal) noise that would be added by any completely passive network having the same loss while in thermal equilibrium at the same ambient temperature, and (b) the "excess" (shot) noise added by the active device (diode). The essential noise is determined by the conversion loss and the physical temperature. The excess noise is commonly characterized by "output noise ratio." Modern high-quality mixer diodes, particularly of the Schottky-barrier type, generally contribute very little excess noise.

(3) By adding noise which enters via unused "spurious" frequency conversions ("spurious ports"). The primary spurious conversion for a mixer is its image response. "Signal" and "image" are arbitrary designations for the two frequencies on opposite sides of the local-oscillator frequency which differ from it by the magnitude of the intermediate frequency. Due to the difficulty in selectively and reproducibly blocking the image by reactive termination, the mixers used in diode testing are broadly tuned to the local-oscillator frequency and are terminated in a line-match at both signal and image frequencies. Thermal noise from the image termination (at a "standard reference temperature," see Appendix D for definitions of terms) is accepted as a part of the mixer (redefined as a two-port which includes the termination of the spurious ports). Noise into the image port which is in excess of this standard noise temperature must be accounted for in the measurement. (The signal and image responses are identical in the broad-band mixer.) Other spurious responses may occur at frequencies in the vicinity of local oscillator harmonics, but this complication has not been investigated. In general, the

termination at the radio-frequency (r-f) port of the mixer should provide a good line-match at the signal, image, and local oscillator frequencies, and at all higher frequencies up to several times the local-oscillator frequency. Such a match not only insures reproducibility of the termination thermal noise, but also of the conversion loss, which depends upon the distribution of signal power among the various terminations and upon the match to the local-oscillator power and its harmonics. (Return of those harmonics to the mixer generating them could affect the conversion loss to an unknown degree depending upon the amplitude and phase of the reflection. The phase dependence would make a reflecting termination difficult to reproduce. The importance of the harmonic termination has not been investigated.)

An additional important way in which signal-to-noise ratio can be degraded is by the presence of noise at the signal and image frequencies contributed by the local oscillator. A single-ended (one-diode) mixer is used for diode testing, and the local-oscillator noise adds directly to the excess mixer noise. A narrow-band bandpass filter must therefore be used between the local oscillator and the mixer. Balanced mixers having two or more diodes are used in receivers; these provide local-oscillator noise rejection by an opposing-phase relationship between diode outputs at the intermediate frequency for noise accompanying the local oscillator, while maintaining an in-phase relationship for conversion from the signal port. Some mixers, of still greater complexity, also provide rejection of the image frequency, thereby making it possible to tune the mixer over a wide range by varying the local-oscillator frequency without also having to vary a preselector (image-rejection) filter.

A detailed analysis of the way in which the signal-to-noise ratio is degraded by the various factors mentioned above is not given. It can be seen, however, that the most important single factor associated with the diode is conversion loss. It can be shown that conversion loss degrades the signal-to-noise ratio by a factor equal to itself. The signal-to-noise ratio degradation by a network for a standard input noise is its noise figure; for finite bandwidths, it is referred to as "average noise figure," $\overline{F}$, which is the gain-weighted average of the (spot) noise figure [1]. The overall average noise figure, $\overline{F}_o$, of the mixer and the intermediate-frequency amplifier, in terms of mixer conversion loss, L, mixer output noise ratio, N, and intermediate-frequency average noise figure, $\overline{F}_i$, is $\overline{F}_o = L\,(N + \overline{F}_i - 1)$, with all quantities expressed as power ratios (not in decibels). Since L always appears as a multiplying factor, the overall average noise figure in decibels is always the sum of the loss in decibels and a noise term in decibels.

Mixer diodes are marketed with overall average noise figure limits in 0.5-dB steps. Suffix letters are added to the generic designation to denote the diode grade. The better diodes have a small excess noise contribution ($N \simeq 1$) which is not greatly different for the various diode grades. The 0.5-dB steps in overall average noise figure therefore result largely from 0.5-dB steps in conversion loss. However, this situation has not been reflected in the specification limits for conversion loss itself, which is still specified at the 6-dB level established for most earlier high-noise-figure types. This came about because of a decision by the diode industry, concurred in by its military customers, to require 100-percent inspection based only on overall average noise figure. This decision followed the development of gas-discharge r-f noise sources which permitted overall average noise figure to be determined by a single measurement (in contrast with a piece-wise determination using the equation given above). Unfortunately, the usefulness of these r-f noise sources as absolute standards has been limited by their apparent lack of stability (possibly resulting from cold-cathode starting or to lack of precision in setting and maintaining the discharge current) and the limited calibration services available. In practice, diode suppliers tend to use r-f noise sources for relative measurements, with sets of previously measured diodes as their reference.

1.3 History of Mixer Standards

In the 1940s and 50s, coordinated diode calibration services were maintained by the U.S. Army Signal Corps Laboratories at Fort Monmouth, N. J., and by the Material Laboratory of the N. Y. Naval Shipyard, Brooklyn, N. Y.* These laboratories also maintained the primary standard holders against which other diode holders could be tuned. Following the *de facto* termination of these services, fixed-tuned holders reproducible solely from their mechanical dimensions were developed under military contract. It was hoped that these fixed-tuned holders, together with the development of the gas-discharge r-f noise source, would compensate for the lack of the calibration services. As has been mentioned, the r-f noise source proved to have limited usefulness as an absolute standard. The fixed-tuned standard holders were found by diode suppliers to be insufficiently reproducible, although there is disagreement between these suppliers as to the relative merits of the several holder types (one in each band) developed to date. The result has been that the major diode suppliers rely almost entirely upon their historic internal "standard" diodes for calibration of their measurements, and strive to maintain a uniform product rather than conform to an absolute standard. One supplier expressed the belief that this practice has resulted in an absolute drift of as much as 1.5 dB in the overall noise figure limit for some diode types. The development of the Schottky-barrier diode for microwave mixer use has brought new suppliers into the market. This and tightened military specifications requiring control of measurement uncertainty have made the lack of absolute standards a significant problem, both to industry and to the government. The NBS mixer measurement program resulted from expressions of concern by industry, through the JEDEC High-Frequency Diode Committee JS-3, and from support by the Naval Electronics Systems Command. Unfortunately, the latter agency was forced to curtail its intended support in 1972 because of funding difficulties (which affected the standards area generally), and NBS was unable to obtain sufficient expressions of need for this work from the private sector to warrant the use of Department of Commerce funds. The mixer program

* These calibrations were started at the MIT Radiation Laboratory during World War II (Military Standard Test Methods for Semiconductor Devices, revised May 1970, p. 265).

was therefore terminated before the principal objectives were met. This report describes the work actually carried out, including the successful demonstration of an improved incremental modulation technique. Not included were any of the contemplated noise measurements, nor, as a consequence, the much-needed intercomparison between the modulation and noise-source methods. (Appendix E constitutes a distillation of unpublished work on mixer noise measurements conducted by the author prior to his joining the NBS.)

2. Design and Analysis of Measurement System

The immediate problem was not one of device characterization — although characterization of mixer diodes does leave much to be desired — but of refinement and evaluation of the quality of the measurements made of those parameters generally accepted for characterization. Accordingly, the decision was made to construct an experimental measurement system at NBS to study the practical problems faced in measuring the traditional mixer parameters. (Certain alternative parameters for the same basic operational, single-frequency characterization were also to be considered.) The measurement methods to be used were preferably to include all those that had been devised to date, for comparison purposes, but the principal goal was to be the minimization of measurement uncertainty, whether by improved engineering practices in implementing previously used methods, or through the development of new methods. It was recognized that the measurement methods in use are numerous and have large numbers of possible variations in their implementation, so that the possible number of intercomparisons is also large, but it was hoped that the main routes to the best techniques using current technology could be explored. In particular, it was hoped that the discrepancies that have been noted between measurements made using a 30-MHz intermediate frequency and those made using a very low (audio or d-c incremental) intermediate frequency could be resolved. It has been the general experience of those working in this area that these two types of measurements frequently yield substantially different results even though there is no theoretical reason for a measurable difference (no methodical investigation of this discrepancy has been reported, however).

A second intended purpose of the NBS system was to enable absolute measurements of high quality to be made for the development of calibration standards or for evaluating diodes for experiments. With respect to this goal, a study of Schottky-barrier diode radiation hardness was completed,* but it was not possible to begin work on calibration standards prior to project termination.

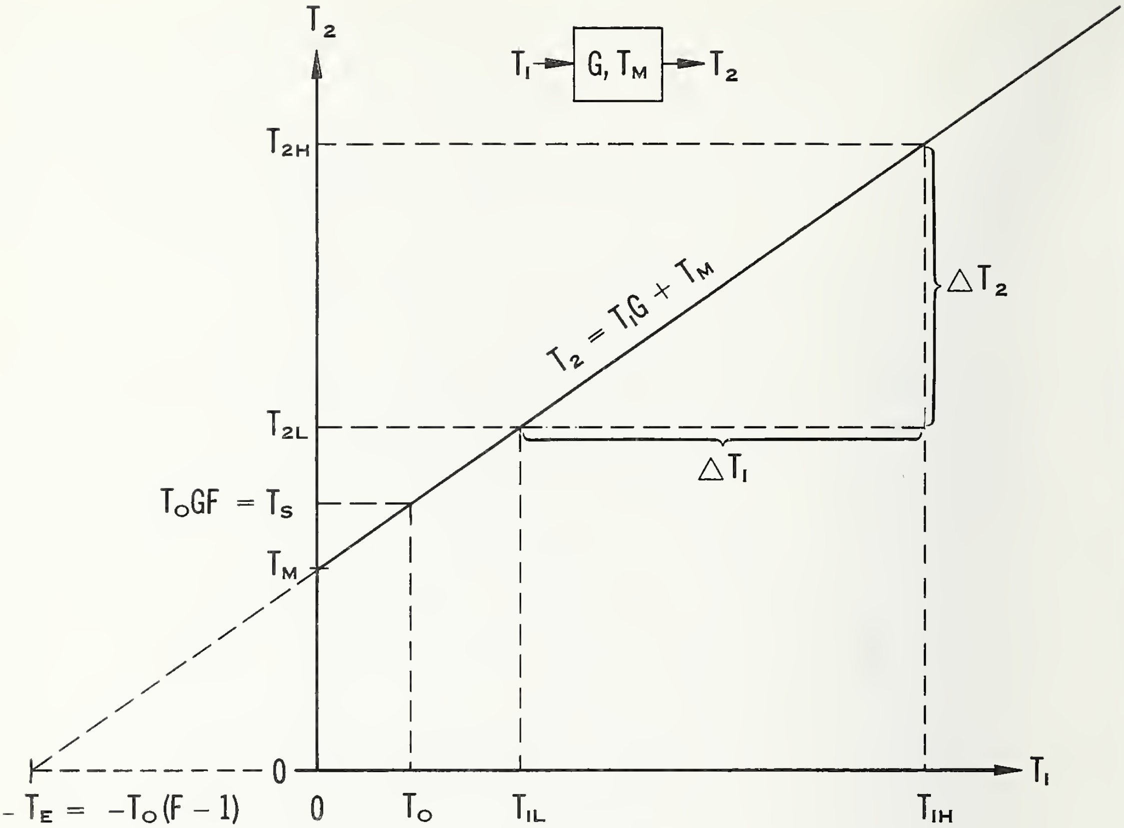
2.1 General Nature of Measurement System

It may be seen from the equation for overall average noise figure given in 1.2 that one mixer parameter may be calculated if certain others are known; there are only two independent parameters if the intermediate-frequency average noise figure is held constant. The mixer is measured

* This work is described in reference [22].

as a two-port network, with the image response assumed equal to the signal response, and combined with it in a manner dependent upon the nature of the signal (quadratically, or powerwise, for uncorrelated inputs and linearly, or voltagewise, for correlated inputs). The other spurious sources are combined with the mixer to form a two-port network as shown in figure 1. (The input and output signals shown are noise temperatures — see Appendix D.) This two-port is essentially linear for signal powers much smaller than the local-oscillator power. (The local oscillator has been assumed to be a part of the mixer.) By definition, such a linear two-port has a linear relationship between input and output, as shown in figure 2 (again, using noise temperatures) [2]. The straight line representing the signal transfer characteristics is completely determined (so long as it remains linear) by only two points, or by one point and the slope of the line, the slope being the power gain (or loss) of the network. (Where the gain is less than unity, as in the case of a mixer, it is usually expressed by its reciprocal, which, for mixers, is conversion loss. For the noise temperature representation of figure 2, the gain or loss is for available power, the ratio of the power available <u>from</u> the network to the corresponding power available <u>to</u> the network.) The calculation of one mixer parameter from two others requires that the latter be measured under identical conditions. If the calculated parameter is to be directly measured for comparison, then it too must be measured under those same conditions. For any of these measurements to be duplicated elsewhere using the same device, with different equipment, these conditions must be sufficiently defined and reproducible. These elementary requirements for objective measurements have unfortunately not always been met in measurements on mixer diodes. Frequently, the different parameters are measured in different holders at different stations of a production line. Occasionally, even the operating frequencies are different, and different criteria are used for setting the local-oscillator drive, as when millimeter-wavelength diodes are measured for output noise by inserting them into an X-band test set using an adaptor and adjusting the local-oscillator drive for a given rectified current (the conversion loss being measured at the nominal operating frequency with a given available local-oscillator power).

Even when every attempt is made to obtain identical measurement conditions, discrepancies have been noted by various investigators. These are frequently attributed to a difference in the r-f environment of the diode at frequencies above the operating band, at and near harmonics of the local-oscillator frequency. Adequate control of this "harmonic environment" is exceedingly difficult because the frequencies involved are above the normal operating range of the waveguide, thereby permitting multimodal transmission. Multimodal transmission is difficult to analyze precisely, because every waveguide discontinuity tends to transfer power between modes. The behavior of these high frequencies in the waveguide components designed for lower frequency, single-mode propagation has neither been established nor controlled by the component manufacturers. Because of the experimental difficulties involved, the degree of dependency of the mixer parameters on the harmonic (and near-harmonic) terminations has not been established, as noted in 1.2. Until proven other-



$$G = \frac{\Delta T_2}{\Delta T_1} = \frac{T_{2H} - T_{2L}}{T_{1H} - T_{1L}}$$

$$T_M = T_2 - T_1 G$$

$$T_E = \frac{T_M}{G} = \frac{T_2}{G} - T_1$$

$$T_S = T_O G + T_M = (T_O + T_E)\,G = (T_O - T_1)\,G + T_2$$

$$F = \frac{T_S}{T_O G} = 1 + \frac{T_M}{T_O G} = 1 + \frac{T_E}{T_O} = 1 + \frac{T_2}{T_O G} - \frac{T_1}{T_O}$$

Figure 1. Temperature relationships for a linear two-port network and its measurement.

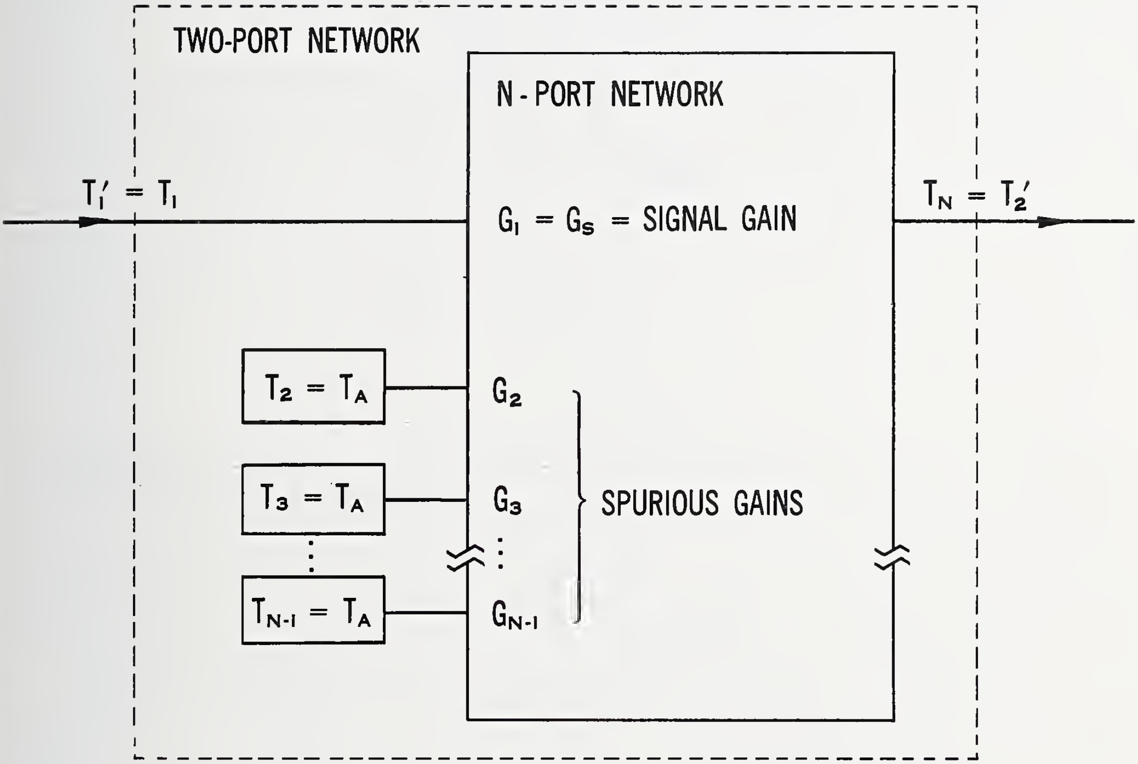


Figure 2. Two-port measurement of an n-port network, at ambient temperature T_A.

wise, it is well to consider these harmonic terminations as having significant effect on precision measurements.

Since the harmonic environment is difficult to reproduce, it is useful to keep this factor constant by performing all of the measurements using the same holder and the same measurement system and with all switches and attenuators that are operated in the course of the measurements isolated from the mixer under test by means of low-pass filters or pads. Other factors that can add uncertainties to the intercomparisons are also eliminated or reduced by this approach: lack of repeatability in the diode-to-holder contacts; differences in match at the signal, image, and local-oscillator frequencies; differences in local-oscillator power or frequency; difference in d-c and intermediate-frequency load, etc. The price to be paid for these advantages is system complexity and, as will be seen, certain compromises with the optimum location of the waveguide components for each type of measurement. The same arguments for a common measurement system for the various parameters apply equally well for measurements of a single parameter using different methods; all factors affecting the measurements should be kept constant except for those unavoidably changed by the difference in the methods.

2.2 Choice of Measurement Frequency

The choice of an operating frequency for the NBS system resulted from a consideration of diode usage, availability of fixed-tuned standard diode holders, and the availability, performance, physical size, and cost of the other microwave components. Two types of point-contact diodes are produced in the largest volume: the 1N21 S-band diode, measured with a 3060-MHz local oscillator, and the 1N23 X-band diode, measured with a 9375-MHz local oscillator. The distribution of use for Schottky-barrier diodes is not known, but may be expected to follow a similar pattern. (There is a tendency to measure them at the same frequencies as the point-contact types, even using the same holder when they have identical external dimensions or using adaptors when they do not.) X-band waveguide components are much smaller and lighter than S-band waveguide components and also less expensive. Coaxial components could be used for parts of an S-band system, but the best performance still requires waveguide, despite the recent improvements in coaxial connectors. Frequencies higher than X-band would yield an even smaller and lighter system, but at greater cost and with reduced component availability and performance.

2.3 Details of Measurement System

The final form of the microwave circuit of the mixer measurement system is shown in the schematic block diagram of figure 3 and in the photos of figures 4 through 17. Overall views of the system are shown in figures 4 and 5. Somewhat greater detail can be seen in figures 6 and 7, each of which shows half the system. These can aid in locating the detail views shown in the remaining photos (figs. 8 through 17), which are sequenced from left to right across the system, seen from the front as shown in figures 4 and 5. This progression also corresponds in a large degree

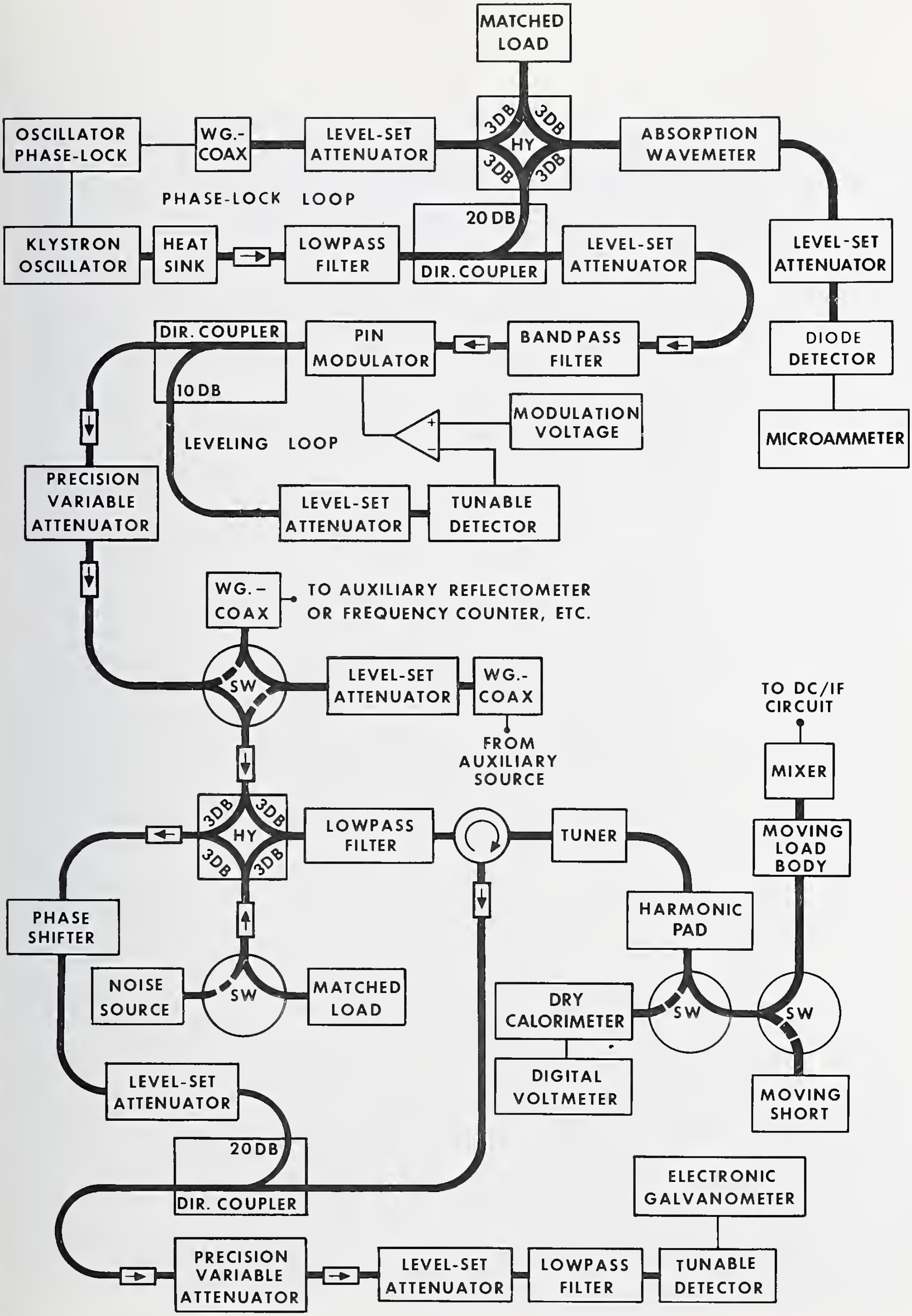


Figure 3. Block diagram of microwave circuit of mixer measurement system.

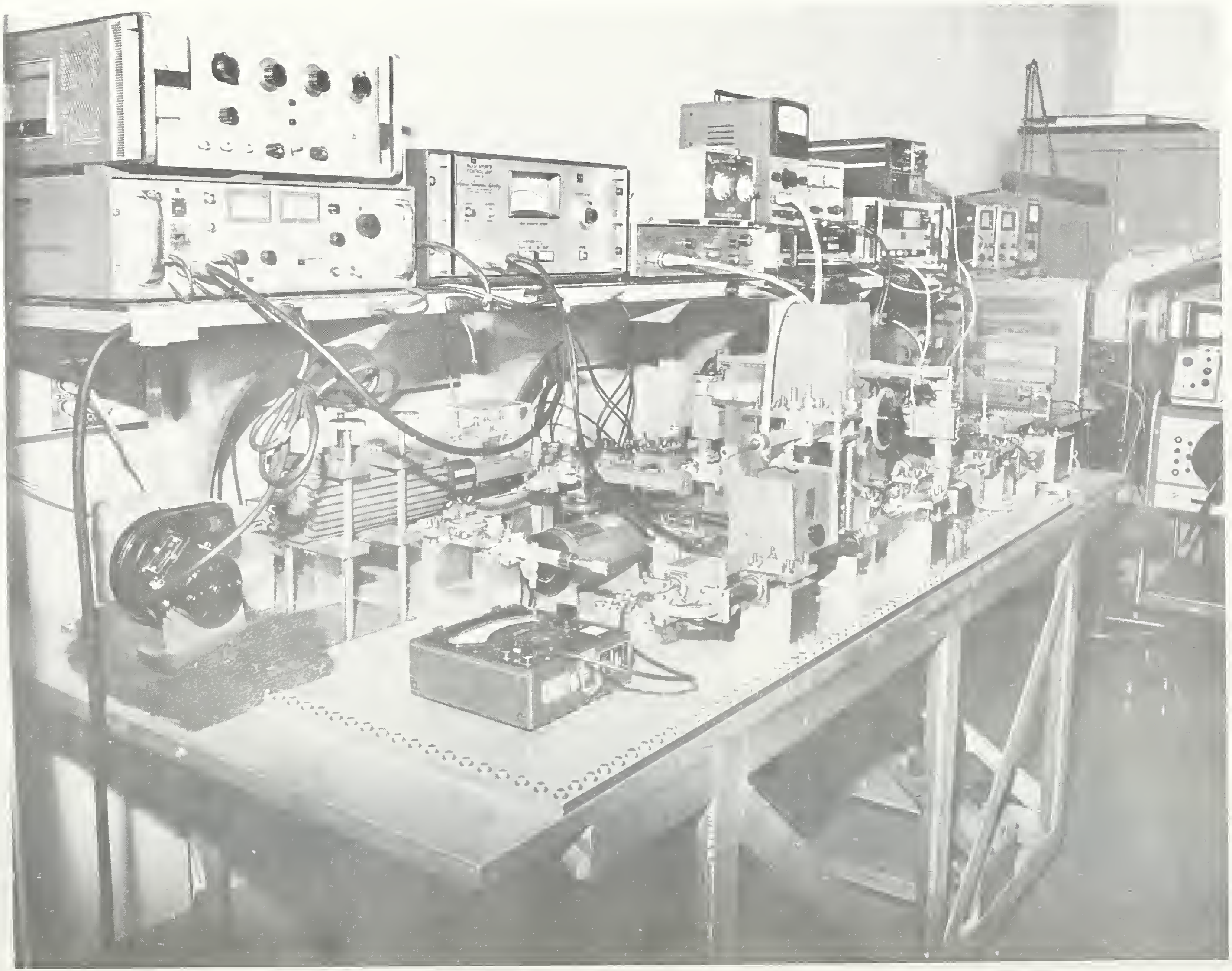


Figure 4. Overall view of mixer measurement system, from left front.

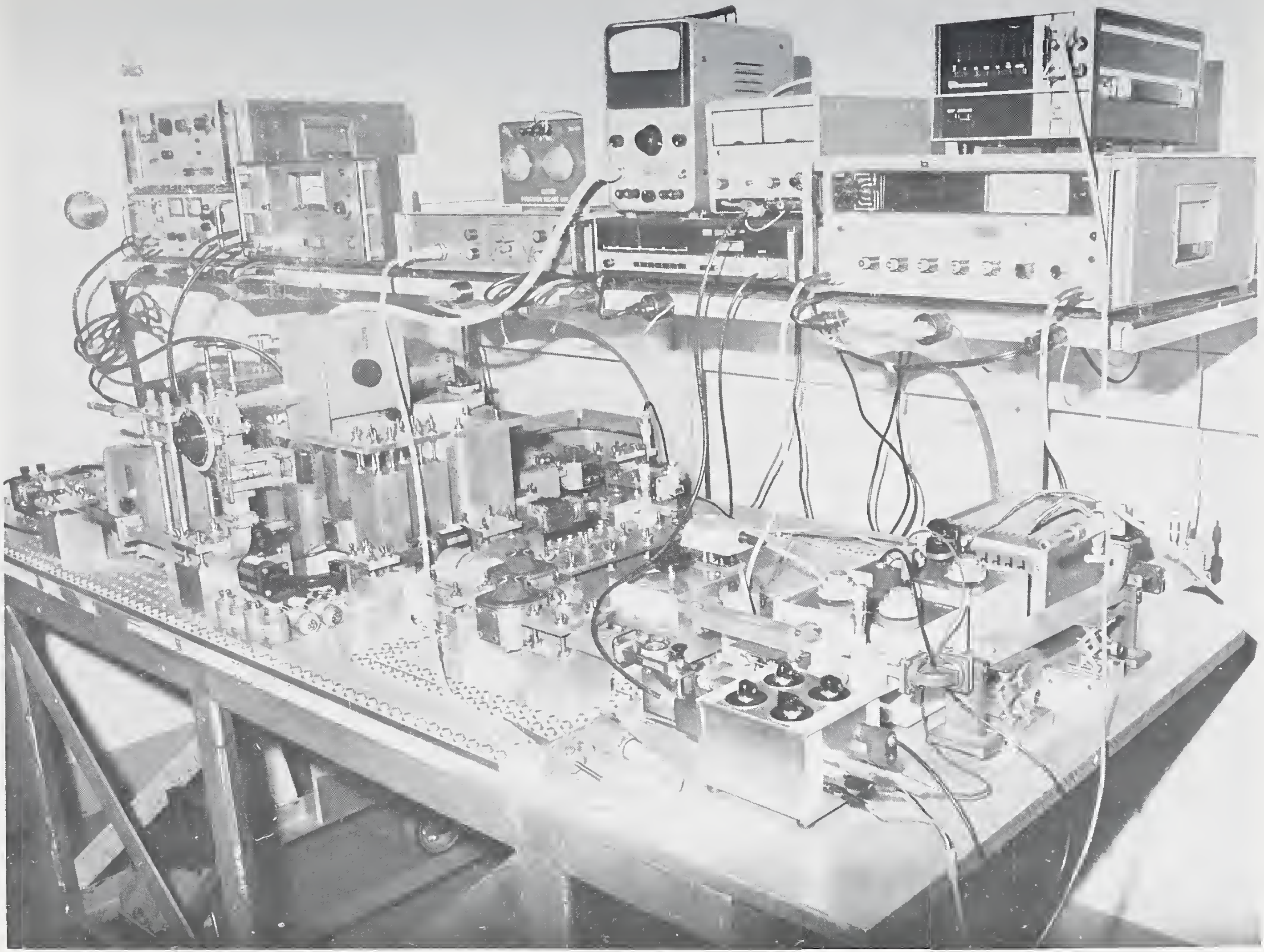


Figure 5. Overall view of mixer measurement system, from right front.

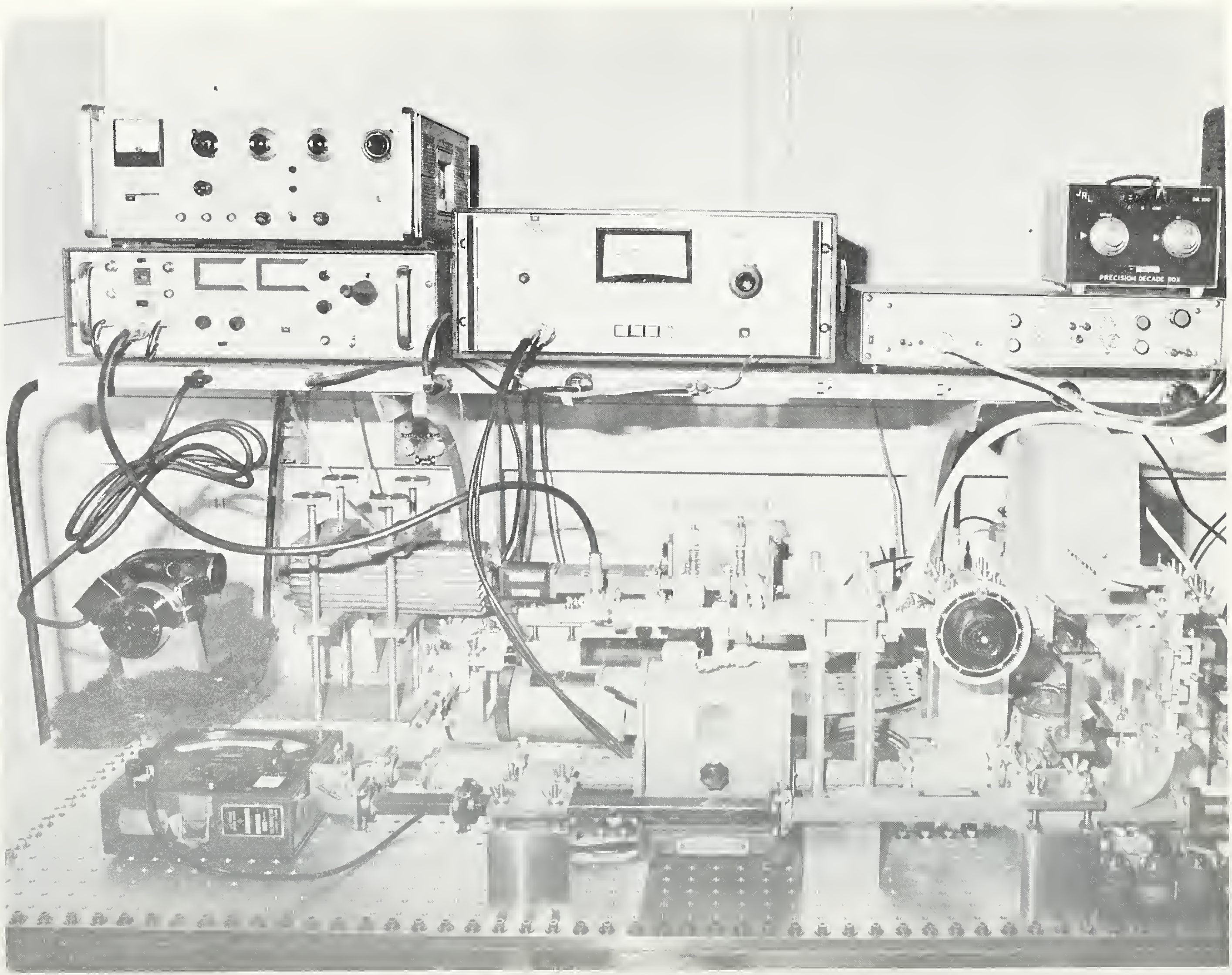


Figure 6. Left half of mixer measurement system.

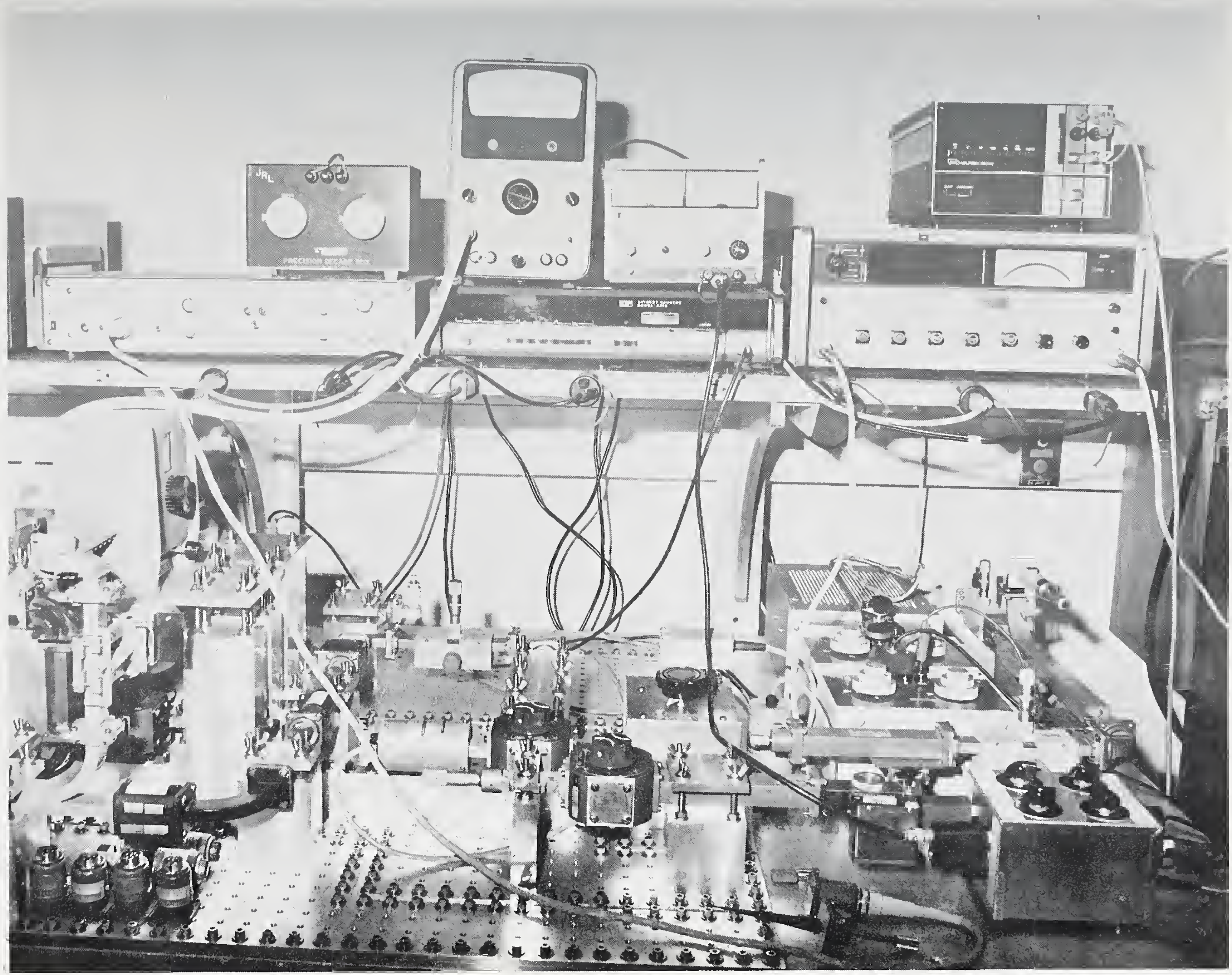


Figure 7. Right half of mixer measurement system.

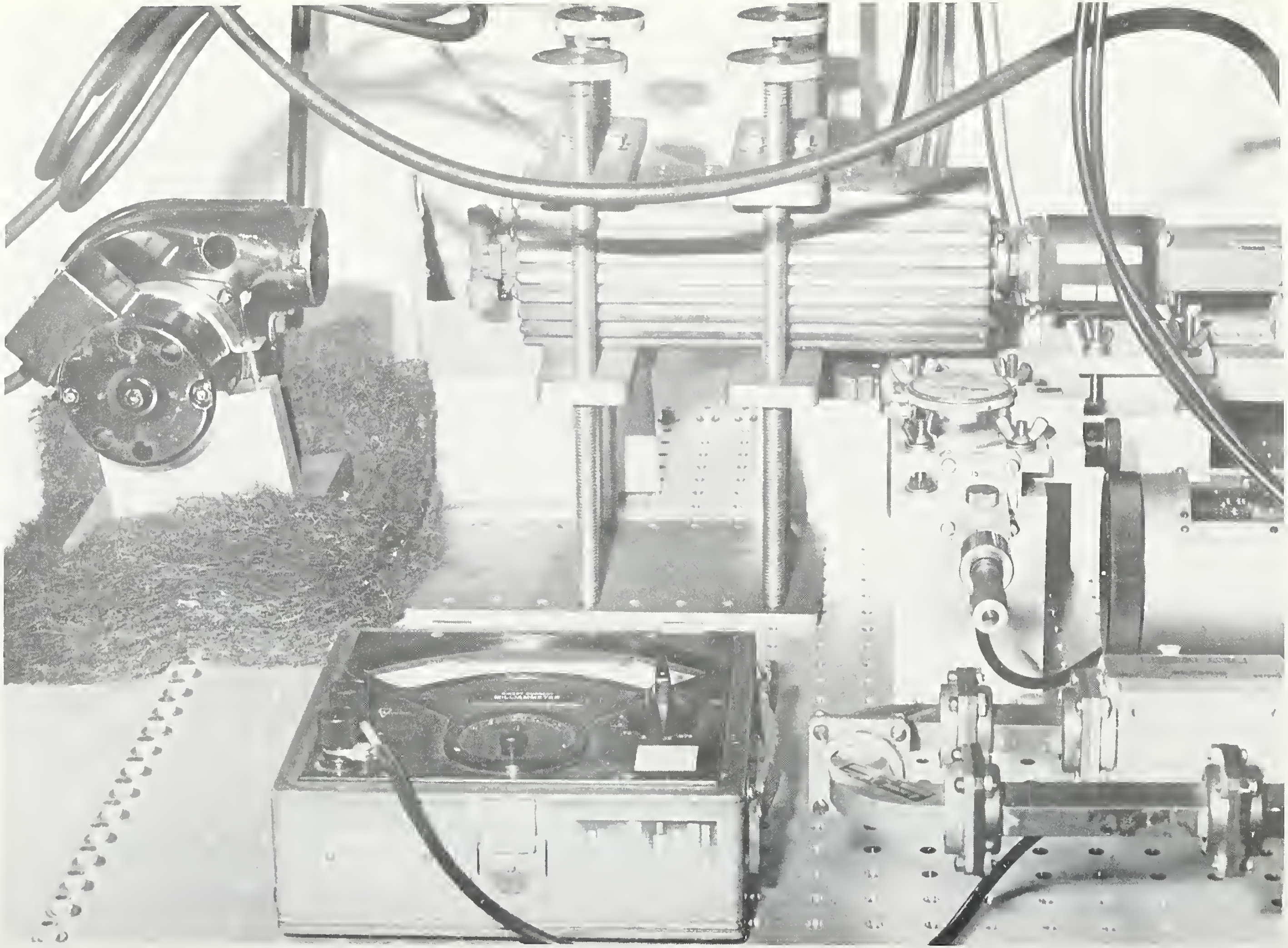


Figure 8. Leftmost side of waveguide system showing heat sink at rear center, with klystron mounted on its left end and an isolator on its right end, with cooling blower at far left and microammeter in the foreground.

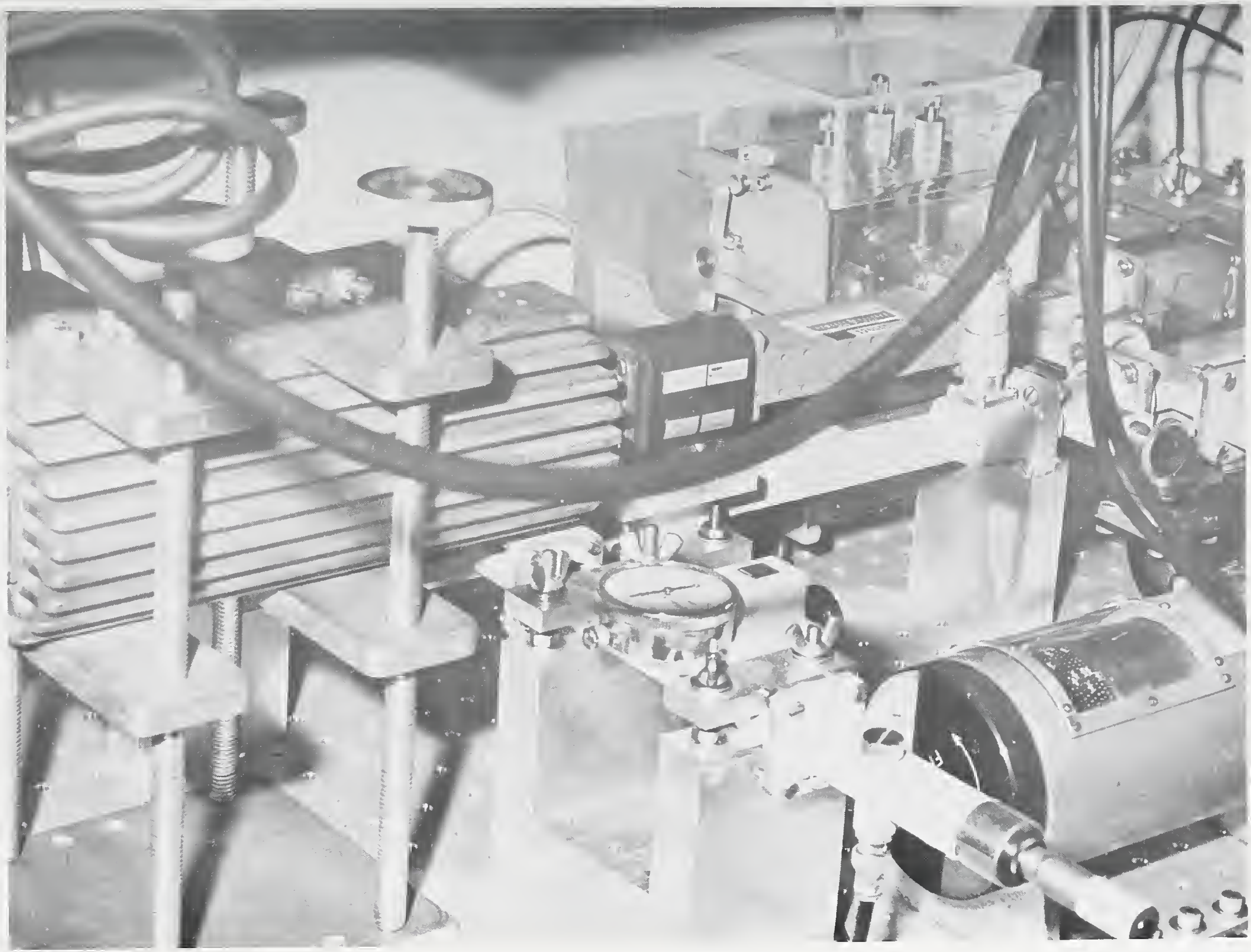


Figure 9. Close-up of heat sink at left and leveling-loop detector and attenuator in the foreground, with modulation attenuator at right rear and absorption wavemeter at right front.

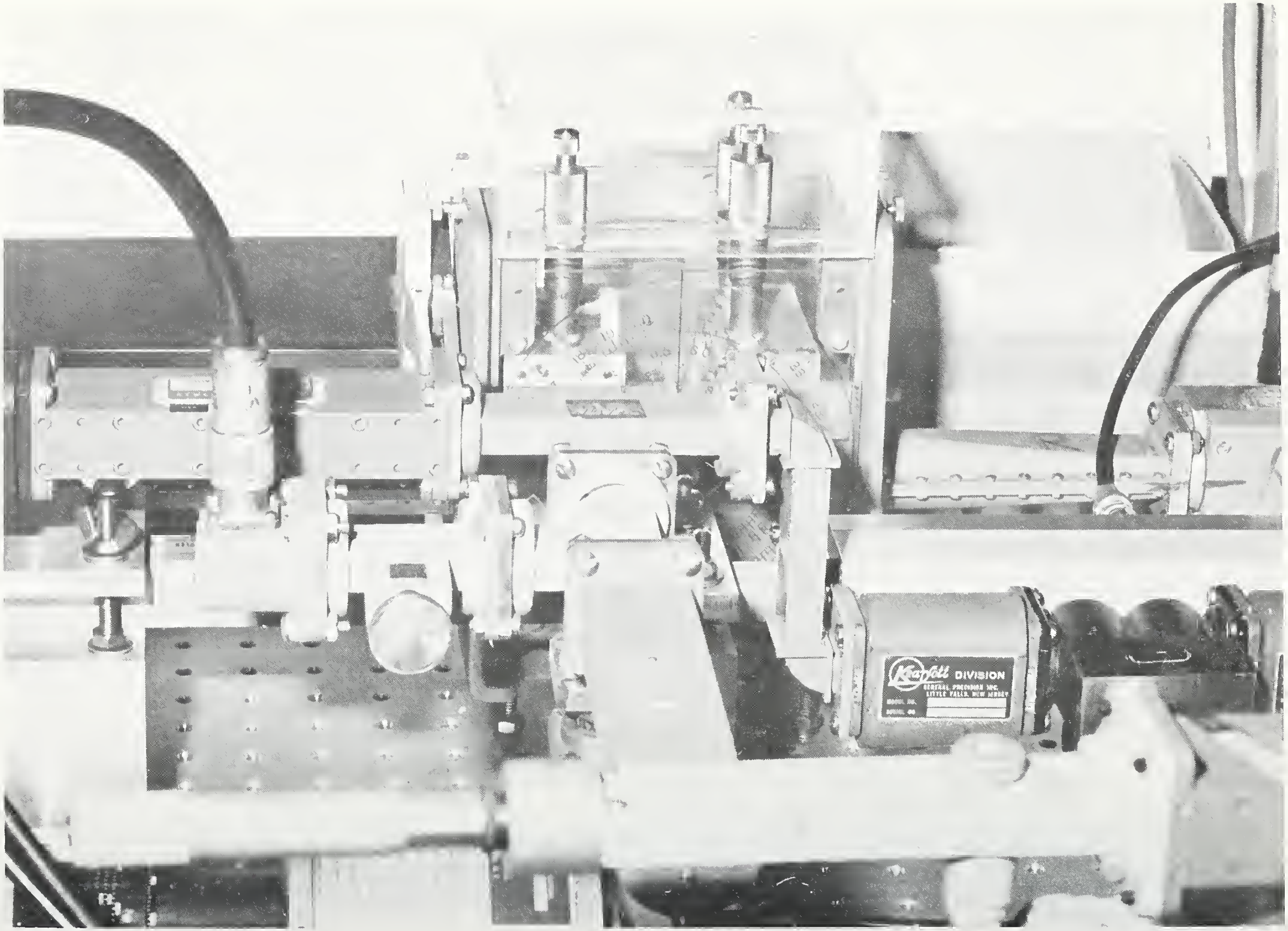


Figure 10. Modulation attenuator at rear center. The single micrometer head on the left is used in setting the nominal local-oscillator power, while the other two are used in setting peak and trough powers. In the foreground is the reflectometer detector and back of it, at the far right, is the two-cavity bandpass filter.

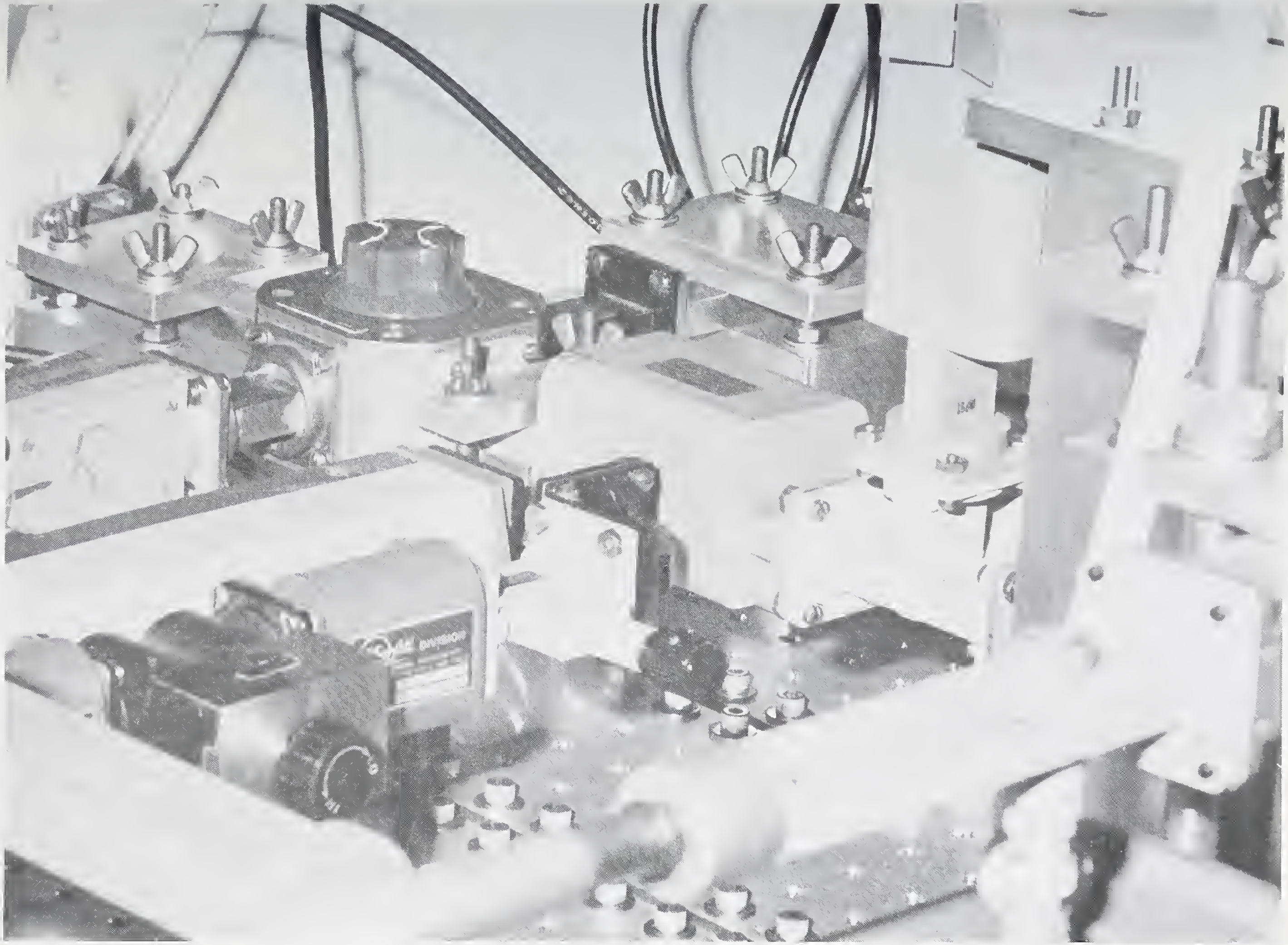


Figure 11. With the reflectometer detector in the foreground, as in figure 10, the bandpass filter is now seen at the far left. The waveguide switch at the left rear is intended primarily to divert power to the auxiliary reflectometer.

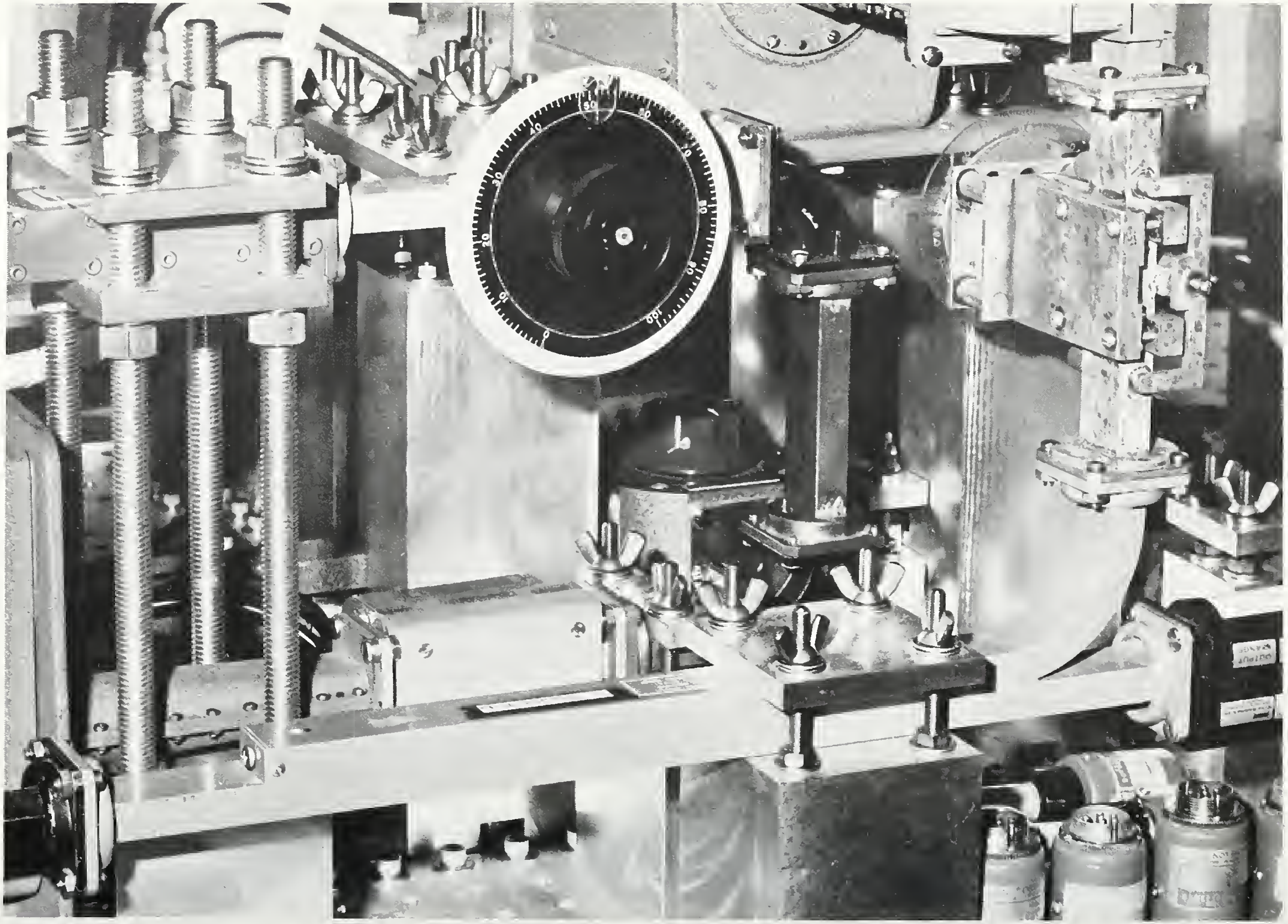


Figure 12. The elevated components comprise most of the reflectometer. At the top rear is the phase shifter, connected to the attenuator at the right foreground, which in turn is connected to a directional coupler to permit cancellation of all reflections other than from the mixer. The waveguide switch in the center is for the noise source.

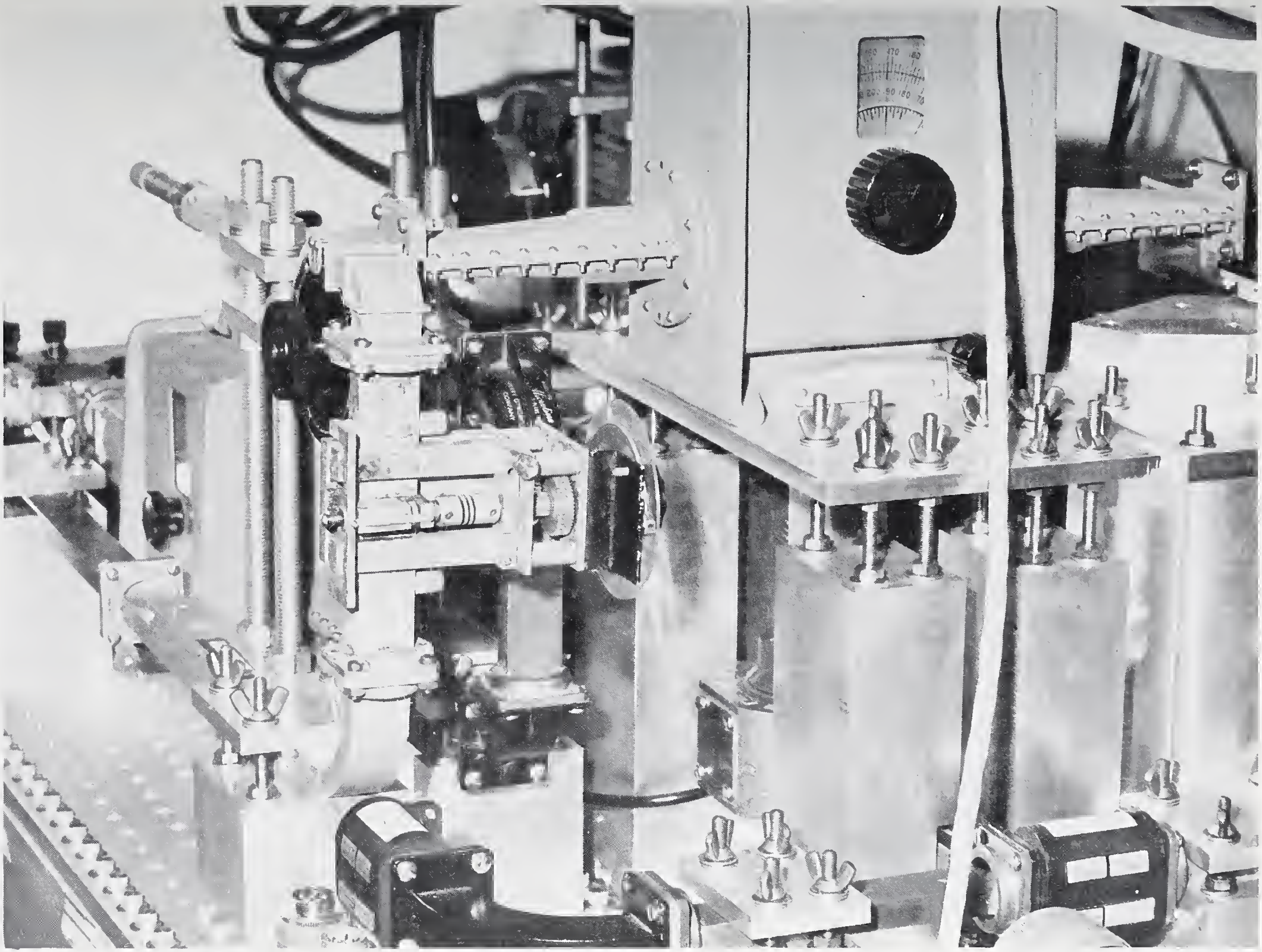


Figure 13. The reflectometer components appear from a different view. The phase-shifter is at the top right, with the associated attenuator at left foreground.

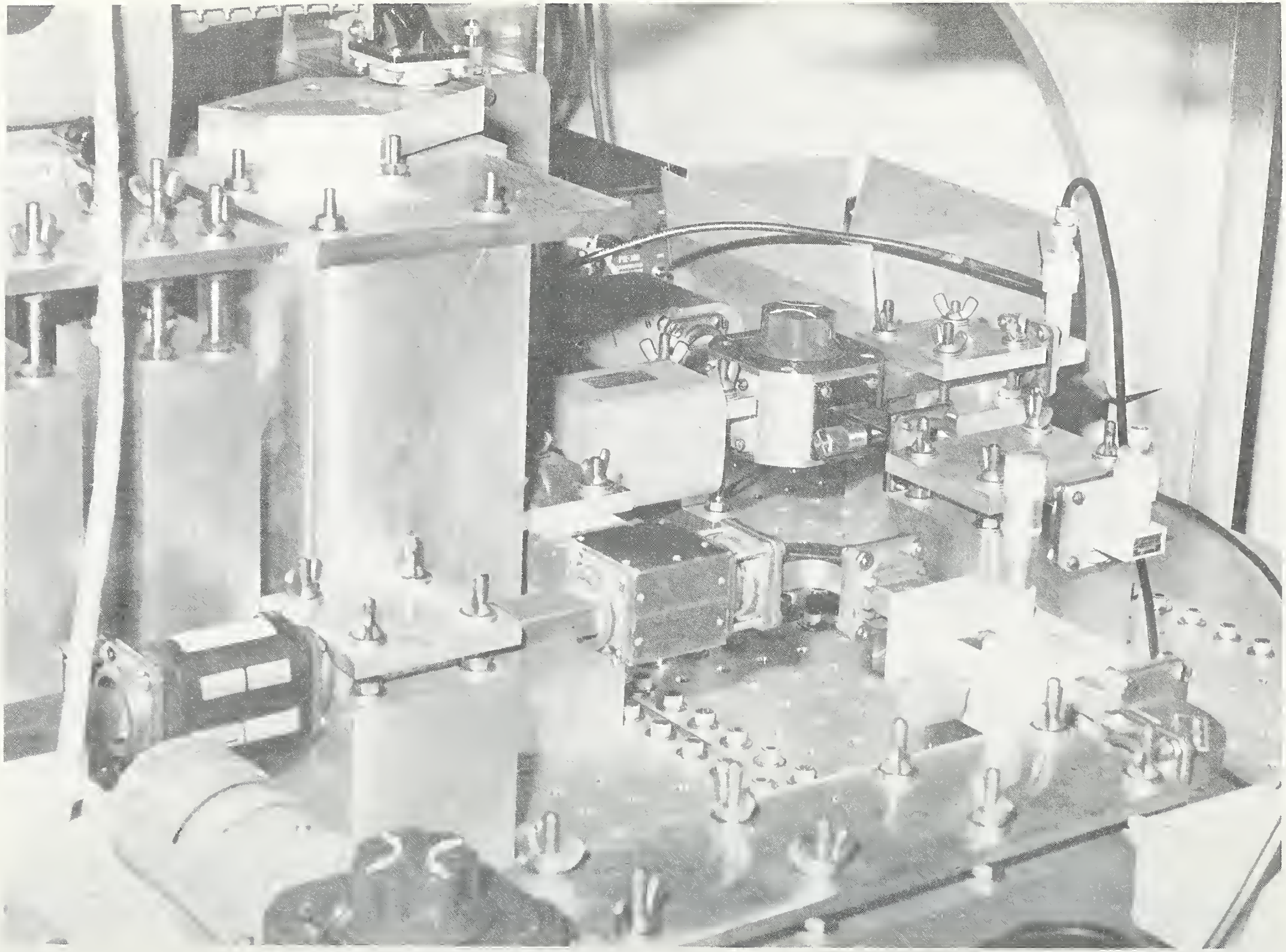


Figure 14. Toward the rear of the bench is seen the waveguide switch, used primarily to divert power to the auxiliary reflectometer, behind the elevated reflectometer components at left. In the foreground is the calorimeter switch next to the harmonic pad (clamped down with eight wing nuts).

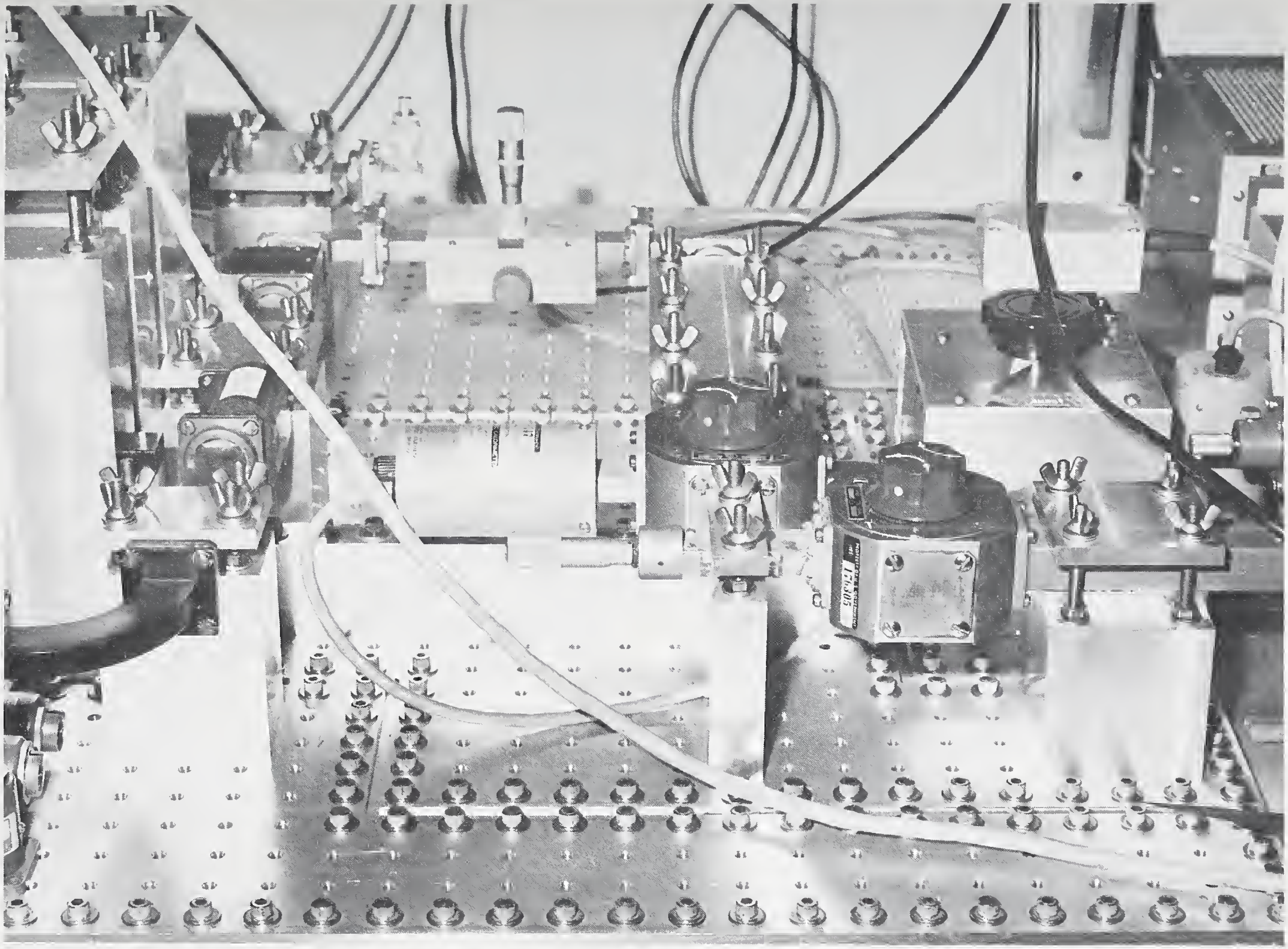


Figure 15. The waveguide switches near the center divert power to the calorimeter (at left of left switch) for short-term power adjustment and to the reference short (at left of right switch) for reflectometer readings.

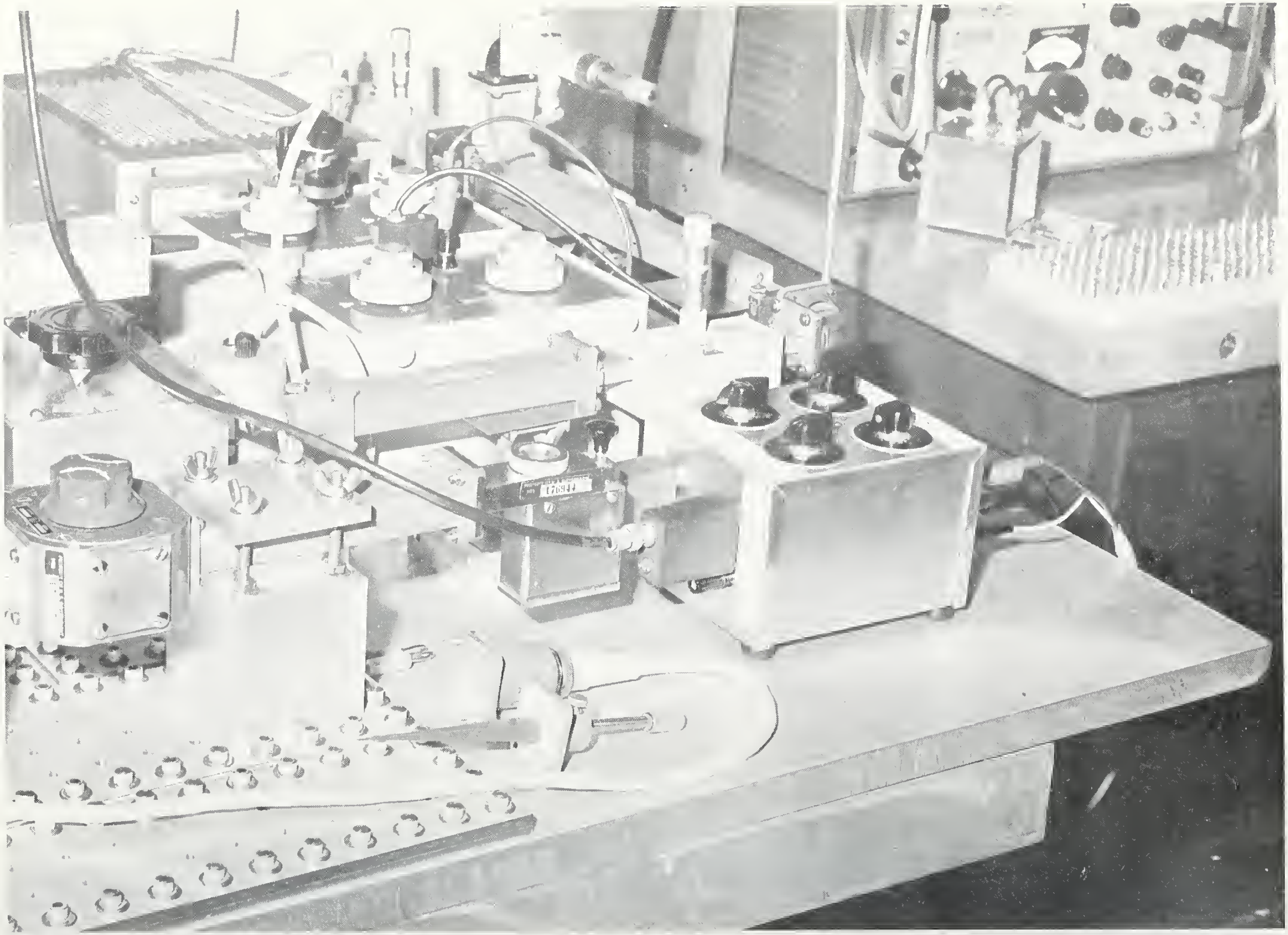


Figure 16. The auxiliary reflectometer is next to the right edge of the bench. The mixer (holder for the diodes under test) and the intermediate-frequency output circuit box (with four knobs) are in the center of this view. Near the front edge of the bench is a sharply pointed moving-load element and, just behind it, a compensated bolometer.

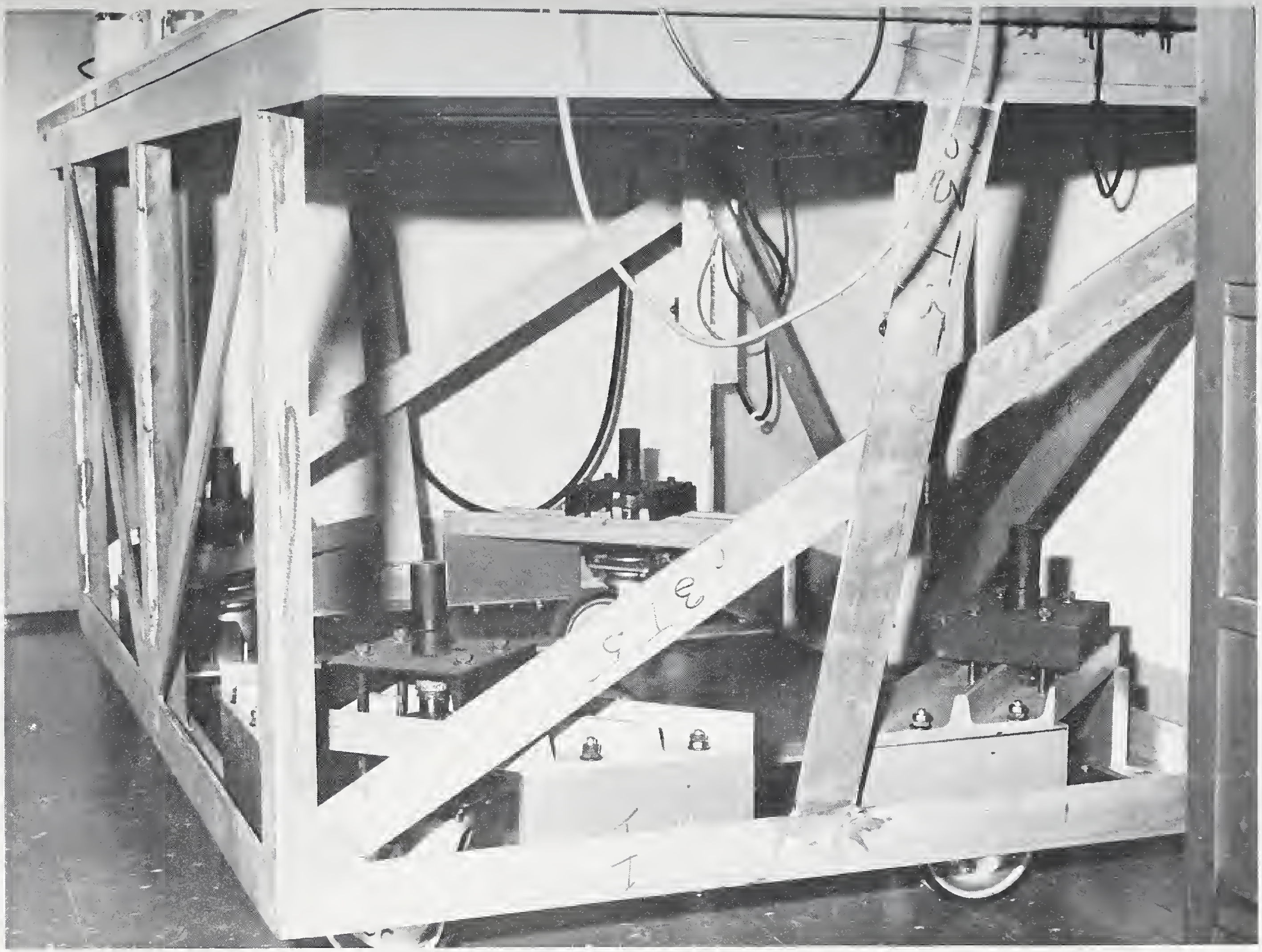


Figure 17. Structure of metal bench. Shock mounting of the casters reduces stresses on the framework as the table is rolled, permitting access to the rear with minimum disturbance of system tuning.

with power flow in the system, as the klystron generator is at the far left and the mixer under test, which constitutes the termination, is at the far right. A more detailed discussion of the various elements of the system and a rationale for their use is given below. The intermediate-frequency output circuit of the system will be discussed later.

2.3.1 Local-Oscillator Stabilization

Since the mixer parameters are functions of the local-oscillator power, it is essential that this power be stabilized to the greatest possible extent. Since the mixer and other components are frequency sensitive, it is necessary to stabilize the local-oscillator frequency as well. At the start of its construction, the X-band system was provided with an oscillator synchronizer to phase-lock the klystron to a single harmonic of a crystal-controlled oscillator, but no active amplitude stabilization was used.* It was found that the local-oscillator power tended to drift during the course of the measurements, even though the klystron and its power supply were highly stable types (and despite the later addition of a massive copper heat sink for the klystron, shown in fig. 8). It was subsequently determined that the PIN modulator, installed primarily to modulate the local oscillator for making conversion loss measurements, could be driven directly by a common type of operational amplifier when used in a leveling (amplitude-stabilizing) loop formed by connecting the amplifier input to a microwave detector monitoring the r-f power after the power had passed through the modulator. Three methods for setting the power level at which limiting takes place were successfully used at various times: (1) adjustment of the operational amplifier offset potentiometer, (2) adjustment of an r-f attenuator to reduce the power reaching the detector, and (3) adjustment of a d-c reference voltage at the operational amplifier input. Small drifts observed in the leveled power led to the solution of an operational amplifier having the lowest temperature coefficient of offset voltage (0.25 μV/C) available in a chopperless type. The amplifier selected had no offset adjustment, however, resulting in the use of method (3), a d-c reference voltage, for setting the power. The use of this more stable amplifier noticeably reduced the power drift but did not eliminate it. The remaining drift may have been caused by changes in the reference voltage, the detector diode, or the waveguide components following the leveling loop. To enable this drift to be corrected prior to each measurement, a dry calorimeter (thin-film thermocouple) and a waveguide switch were installed near the mixer to monitor the power after leveling. A dielectric-filled length of waveguide between the monitor and the mixer, used to provide padding to improve the match to local-oscillator harmonics generated by the mixer, proved to be temperature sensitive, leading to the temporary installation of a second calorimeter in place of a moving short used as the reflectometer incident-power reference. The system was then rebuilt so that both the calorimeter and short could be present together at the mixer end of the harmonic pad. Ideally, the calorimeter and short should each be as close to the mixer as possible; in this case, the reference short is given priority, with its switch separated from the mixer only by a short

* See figure 19 on p. 43 of reference [3].

length of waveguide which was designed to accept a moving load. The characteristic resistance of this precision waveguide constitutes the r-f immittance standard.

The waveguide was clamped against blocks bolted to a rigid base plate in order to eliminate the large and partly irreversible power changes that were observed when forces were applied to the waveguide or the supporting bench. Such forces occur during the measurements in the course of inserting and removing diodes, interchanging the mixer with bolometers and moving loads, and operating switches and attenuators in the system. A system susceptible to such forces may also be expected to move erratically as the temperature changes, due to sticking-and-slipping. In addition to changing the local oscillator and signal powers available to the mixer, this waveguide motion may also be expected to alter the r-f source immittance presented to the mixer, thus making precision tuning impractical.

The two 1/4-in. (6-mm) aluminum plates against which the system was clamped proved to be insufficiently rigid when supported by a wooden work bench. Fortunately there became available as surplus property an all-metal bench consisting of a 1-in. (25-mm) aluminum slab supported by a very rigid welded aluminum framework of triangular members formed by welding strips to L-beams. (See fig. 17.) Although the aluminum top was probably not as rigid as some other materials, such as granite, it appeared adequate for this application, even without the two 1/4-in. (6-mm) plates having been bolted to it. The bench was mounted on large casters to permit it to be easily moved for access to all sides; shock mounts were used with these casters to minimize distortion of the framework, as the bench was moved. A large, sturdy overhead shelf was added to support the power supplies and instruments. (See figs. 4 through 7.)

Details of the clamping system may be seen in figures 8 through 16. The waveguide rested on aluminum blocks drilled to clear 1/4-in. (6-mm) bolts held by the upper baseplate, which was drilled and tapped on 1-in. (25-mm) centers over its entire surface by a tape-controlled automatic vertical milling machine. This upper plate was divided into two sections, bolted around their peripheries to the lower baseplate. One of the sections held that portion of the waveguide (bounded by isolators) that could be removed for calibration of the noise temperature at the mixer input. This removable section was subsequently made more rigid by a bar across its longest dimension. A noise calibration was not made, nor was this section ever actually removed once the waveguide was mounted; it was contemplated that if its rigidity after removal was inadequate, additional reinforcement would be added.

2.3.2 R-F Immittance

To determine the magnitude of the reflection by the mixer of local-oscillator power, a reflectometer circuit was added to the system. In its final form, as shown in the lower left of figure 3, it consisted of (a) a circulator to separate the incident and reflected waves; (b) a reference short to reflect all of the incident power; (c) a calibrated attenuator to reduce this incident power to the level of the reflection from the mixer, thereby setting it equal to the return loss (from which the SWR may be calculated); (d) a diode detector and an electronic galvanometer to indicate this equality of power levels; (e) a level-set attenuator for zeroing the galvanometer when the galvanometer zero adjustment is not adequate; and (f) a level-set attenuator and a phase shifter which allow the cancellation of reflections from other than the mixer port. The power used for this cancellation was that half of the local-oscillator power lost in the process of combining, in a hybrid ("magic") tee, the local-oscillator power with noise power from the gas-discharge noise source.

2.3.3 System Tuning

For the low intermediate-frequency measurements considered in this phase of the program, the signal, local-oscillator, and image frequencies are essentially the same, and the tuning can thus be narrow band (neglecting, for the moment, the harmonic termination). To obtain a local-oscillator frequency match, an auxiliary reflectometer was used, consisting of a tuned 3-dB directional coupler with a diode detector (see fig. 16). No attenuator or reference short was provided with this auxiliary reflectometer since its function was to indicate when a perfect match had been obtained, not the degree of mismatch. It received power via a switch in the local oscillator line and a coaxial cable. Slide-screw tuners were used to tune both the auxiliary reflectometer and the mixer source; future work at a 30-MHz intermediate frequency would require more elaborate tuners (multistub, etc.) to obtain sufficient bandwidth, and a separate generator (possibly swept) for tuning. The output port of the auxiliary reflectometer and the port to which the mixer was attached both consisted of moving-load bodies to facilitate using such loads with a minimum of waveguide flange connection uncertainty.

The tuning procedure was as follows: (1) With power switched to the auxiliary reflectometer, the electronic galvanometer attached to it, and a moving-load element inserted into its output port, tune for a galvanometer null, then refine this match by moving the load element back and forth through its range and tuning carefully for a galvanometer deflection nearly independent of load position (this is done by pushing down the peaks with the tuner). This refinement allows for the imperfect match of the load element. (2) Remove the load element and attach the auxiliary reflectometer to the system in place of the mixer, then tune the mixer source to obtain a null. This completes the tuning for a reflectionless mixer source at the local-oscillator frequency. (3) To tune the system reflectometer, connect the electronic galvanometer to it, switch the power back to the main line, insert a load element in the mixer source port after disconnecting the auxiliary reflectometer and adjust the phase shifter and the adjacent level-set attenuator for a galvanometer null. As with the auxiliary reflectometer tuning, this match may then be refined by tuning for a nearly constant galvanometer deflection as the load element is moved. With the power switched to the moving short, adjust it for a maximum reflection. (There should be little power variation over its range.) The assumption that this reflection is nearly equal to a perfect reflection at the mixer port may be checked by inserting a (noncontacting) moving-short element into the mixer source port and comparing the reflections with the calibrated reflectometer attenuator. With perfect tuning, the galvanometer deflection is independent of the position of the short or any load element, since all reflections reaching the detector have been cancelled except that from the moving element; when the tuning is not perfect, the reflection from the moving element adds vectorially to the residual fixed reflections to form a resultant that varies in amplitude with the phase changes produced by moving the element. Note that by leaving the moving load body in the system, reflections due to imperfections in the mixer flange connection are included in the evaluation of mixer mismatch.

To improve the match at harmonics of the local-oscillator frequency, a section of waveguide 11 in. (280 mm) long containing a lossy dielectric was placed near the mixer. To insure attenuation of all possible waveguide modes, the dielectric was shaped to fill completely the waveguide at its center; to insure a good match for all frequencies it was tapered linearly for 5 in. (130 mm) at each end in both planes, with the points laid in diagonally opposite corners of the waveguide to reduce sensitivity to point imperfections. (The fields are zero at the corners for all modes.) The first dielectric material used was unfired steatite, giving a 7-dB loss at the local-oscillator frequency. When the steatite proved to be exceedingly temperature sensitive, it was replaced with nylon, giving a 2-dB loss. While not as temperature sensitive, the padding provided by the latter may not have been adequate, although a high local-oscillator power loss cannot be tolerated, and the r-f noise source is even more sensitive in this regard; the 7-dB loss of the steatite pad was probably excessive for optimum noise measurement sensitivity.

2.3.4 Local-Oscillator Modulation

2.3.4.1 Introduction

To measure conversion loss, it is necessary to introduce a stable signal of known amplitude which is very small compared with the local-oscillator power (which, for X-band diodes, is 1 mW available to the mixer). It is generally believed that a signal of at least 20 (or possibly 30) dB below the local-oscillator level is sufficient to insure adequate mixer linearity (although this remains to be verified). Such a signal could be produced by an independent signal source but, except for a uniformly distributed signal such as white noise, this would pose a severe frequency stability problem which would be worse at the lower intermediate frequencies. Since the intermediate frequency is the difference between the signal and local-oscillator frequencies, its absolute instability is the sum of the instabilities of those two sources; percentagewise, the instability of the intermediate frequency is much greater than that of either signal or local oscillator, being inversely proportional to the frequency. One solution to this problem is to synthesize the signal from the local oscillator by amplitude modulating the latter at the intermediate-frequency rate. One of the two symmetrical sidebands ordinarily produced by such modulation could be suppressed, either by modulator design or by filtering, but it is simpler to use both sidebands and make allowances for their linear (voltagewise) addition by the mixer. For mixer measurements, this modulation was once commonly done with a mechanical modulator which periodically attenuated the local oscillator by means of an eccentrically mounted motor-driven disk of lossy material [4]. In recent years, ferrite and PIN modulators have come into favor. The commonly used rectified-average electronic type of a-c voltmeter to indicate the mixer output voltage would require that the modulation be perfectly sinusoidal for absolute measurements, i.e., where the modulation factor is explicitly known. As a consequence of the difficulties in obtaining a known sinusoidal modulation, almost all mixer measurements using this method have relied upon calibrated "standard" diodes, thereby making the measurement a relative one and allowing the use of an arbitrary modulation waveform.

Methods of obtaining an absolute measurement of the modulation factor, and utilizing an arbitrary waveform in making absolute conversion loss measurements based upon this modulation factor, were devised by the author at the Material Laboratory of the N. Y. Naval Shipyard (later renamed the Naval Applied Science Laboratory) almost two decades ago. These methods were privately circulated but did not come into general use. Improvements in equipment since that time have made these methods appear even more promising for high-quality absolute measurements. As will be seen, however, refinement of these periodic modulation methods at NBS has led logically to a return to the incremental modulation method traditionally used for absolute conversion loss measurements [4]. The periodic and incremental modulation methods have traditionally been regarded as completely distinct, with apparently unrelated derivations.* The work begun at NASL and continued under this program has stressed their fundamental identity and reduced the distinction to a choice of the best instrumentation to achieve the smallest measurement uncertainty. While periodic modulation was not used at NBS, the calibration and mixer output measurement methods may be of value in other circumstances, for measurements other than those of mixers as well as for their original use.

2.3.4.2 Periodic Modulation

As shown in Appendix A, linear sinusoidal modulation of modulation factor m results in a maximum ("crest") r-f power which is 20 $\log_{10}$ (1 + m) decibels above the carrier power, and a minimum ("trough") r-f power which is 20 $\log_{10}$ [1/(1 - m)] decibels below the carrier power, for a total power excursion of 20 $\log_{10}$ [(1 + m)/(1 - m)] decibels. (The "instantaneous" — crest or trough — r-f power is the average power over a full r-f cycle, which can be regarded as sinusoidal when its period is very small compared with the modulation period, as it is assumed to be

* Pound [5] did comment (p. 103) that the incremental method "may be regarded as an extrapolation of the modulation method to zero frequency," but he did not relate the methods quantitatively or otherwise use this observation. The MIT Radiation Laboratory approach [4-6] is more general, but considerably more complex than the analysis to be given here.

here.) This crest-to-trough power ratio can be measured by displaying the detected modulation on a d-c coupled oscilloscope (so that the display is a unique function of "instantaneous" power), and using a calibrated r-f attenuator following the modulator (and isolated from it) to attenuate the crest until it is equal to the trough, as shown by coincidence with a horizontal reference trace on the oscilloscope. The attenuation required for this is the crest-to-trough power ratio. The demodulation must be accomplished with sufficiently low capacitance that the filter time constant is very small compared with the modulation period (again, so that the trace position is a unique function of "instantaneous" power). The most stable reference trace is generated by switching the oscilloscope periodically between the detector and a stable adjustable d-c reference voltage, thus eliminating the effect of oscilloscope drift. The optimum relation between the switching frequency and the modulation frequency must be determined by experiment; a 100-Hz switching rate was found to be suitable for 1000-Hz modulation, but lower switching rates (below flicker-fusion) were sometimes deemed preferable. The detector can be the mixer under test, since it will generally have a small output filter time constant, usually suitable for at least 30-MHz intermediate-frequency use.

With a calibrated reference voltage, the crest-to-trough mixer output voltage (peak-to-peak of the a-c component) was determined with the NBS X-band system. Periodic modulation was accomplished by applying the modulating voltage to an input terminal of the operational amplifier used in the leveling loop; the r-f power was thus constrained to follow the modulation voltage to the extent that the leveling detector was linear over the relatively small excursions of the shallow modulation used. For such shallow modulation, modulation linear with power is essentially identical to modulation linear with voltage since $[1 + m \cos(\omega_m t)]^2 \approx 1 + 2m \cos(\omega_m t)$, for $m << 1$, where m is the modulation factor and ω_m the modulation frequency in radians.

Precise comparison of the crest and trough voltages with the reference trace is facilitated by using square-wave modulation, so that the crest and trough are lines rather than points. All calculations can still be carried out in the same manner as for sinusoidal modulation having the same total excursion, provided that the measured instantaneous values are effectively independent of the history of the waveform, i.e., that there is no significant factor having a time constant that is appreciable by comparison with the modulation period, such as is associated with large electrical or thermal capacitance and resistance. A low modulation frequency obviously favors this condition. A small modulation factor has the same effect by minimizing the variation in power dissipated in the diode and therefore keeping its temperature changes small. Since shallow modulation is required for mixer linearity anyway, this requirement is not an added burden. A low modulation frequency is more difficult to separate from d-c bias in the mixer output, but the load circuit that will be described in conjunction with incremental conversion loss measurements has no low-frequency limitation.

It was found that the best measurement resolution could be obtained by using an oscilloscope having a very high gain vertical amplifier with adjustable offset. This enabled the demodulated signal to be amplified to many times the size of the oscilloscope screen, the offset being used to bring either the crest or the trough on screen for comparison with the d-c reference voltage.

Attenuator calibration and mixer output measurements using this oscilloscope comparison technique with periodic modulation have two major disadvantages. First, a precise bilateral increment cannot be easily obtained, i.e., the crest and trough can be located with respect to each other but not to the carrier (local-oscillator) level. This is because the modulation cannot be altered without affecting the carrier level, making it difficult to set the latter precisely. The use of independently adjustable alternating positive and negative modulation pulses would allow the preservation of a base line representing the unmodulated local oscillator, but with considerable modulation circuit complexity. For shallow modulation, the range of local-oscillator power uncertainty is small, and no change in the crest-to-trough output voltage was actually noted as the mean power was varied over roughly this range, but this observation was made at a time when the output voltage resolution was limited. The second major disadvantage is that the oscilloscope comparison for periodic modulation requires the use of a sizable bandwidth in order to have the fast rise and fall times and low tilt necessary for adequate flatness of the top and bottom of the square wave. This wide bandwidth resulted in broadening of the oscilloscope trace by noise, which limited the precision with which the comparison could be made. A lesser disadvantage is the requirement of high stability of both the modulating voltage source and the modulator.

2.3.4.3 Incremental Modulation

Implicit in the oscilloscope comparison method using squarewave periodic modulation is the assumption that the mixer output voltage is independent of output frequency, i.e., that the conversion loss is unrelated to the intermediate frequency. For a d-c coupled output, with low pass filtering in the many-megahertz-cutoff range, this is a reasonable assumption, since the reactances and susceptances associated with the diode must be very small to permit its performance at microwave frequencies. The apparent conversion loss would thus be expected to approach an asymptote as the intermediate frequency goes to zero, and be essentially constant below a good fraction of the local-oscillator frequency. The value of this asymptote would in fact be the least arbitrary definition of conversion loss as far as the intermediate frequency is concerned.

In the oscilloscope comparison method, only one level of the square-wave modulation can be measured at a time. In the absence of any low limit to the intermediate frequency, it is possible, then, to eliminate the automatic switching that produced the square-wave modulation, and perform it manually, thereby eliminating the need for a rapid rise time and permitting a much narrower bandwidth, reducing the noise limitation on resolution. With a "zero" intermediate frequency, the only limitation on bandwidth reduction is the corresponding increase in settling time.

With incremental modulation, the comparison circuit is unnecessary, thus eliminating the need for the oscilloscope, the adjustable reference voltage, and the automatic switch. A d-c digital voltmeter can now be used to measure the crest and trough output voltages directly.

With incremental modulation, the electrical modulator is no longer required; the attenuator used previously to calibrate the electrical modulation can now be used directly as the modulator. In addition to its crest and trough settings, this "mechanical modulator" can also be used at an intermediate attenuation position for setting the local-oscillator power. The use of a passive component, the attenuator, in place of an active component, the electrical modulator, could be expected to improve the modulation stability, but only if the attenuator action has good repeatability.

2.3.4.4 Modulation Attenuator

The attenuator used in the X-band system for mechanical modulation (and, earlier, for calibration of electrical modulation) is a modified commercial rotary-vane attenuator. The theory of operation of this type of attenuator [7-9] need not be discussed in detail here, but for those unfamiliar with its operation a brief description may help to clarify the modifications to be discussed.

The rotary-vane attenuator consists of three cascaded sections of circular waveguide, with tapered transitions to rectangular waveguide at the extreme ends of the outer sections. These transitions transform the $TE_{10}(H_{10})$ mode in the rectangular guide to $TE_{11}(H_{11})$ in the circular guide. Each of the circular waveguide sections is longitudinally divided by an attenuating vane of metallized mica clamped between the two semicylindrical halves forming each section. Low-loss propagation can take place in these sections only for a wave whose electric vector is normal to the vane. The center section can be rotated through a 90-deg angle by a worm-drive mechanism. For any angle other than zero (coincident alignment of all vanes), the wave propagated through one of the end sections enters the center section with an electric vector forming a nonnormal angle with the center-section vane, an angle complementary to that between the given end-section and center-section vanes. This vector may be resolved into two components, one normal to the center-section vane and the other parallel to it. The wave represented by the latter is rapidly attenuated as an exponential function of distance, and the length of the section is such that very little of this reaches the other end. For a relative vane angle θ, the original electric field has been reduced by a factor $\cos\theta$, which represents that component which was normal to the vane; the power is thus reduced by a factor of $\cos^2\theta$. After the remaining wave reaches the end of the center section and enters the other end section, it again experiences the same vane-angle difference and is again reduced in power by a factor of $\cos^2\theta$, for a total reduction factor of $\cos^4\theta$. In decibels, the attenuation is thus $-40 \log \cos \theta$ [10,11].

The scale of the available rotary-wave attenuator did not afford sufficient resolution, and there was appreciable backlash in the drive mechanism. It was therefore decided to use adjustable stops to limit the rotation of the center section precisely. Hemispherical head screws, turned from tool steel (drill rod) and hardened to Rockwell C61 or C62, were inserted as stops in the flange used to hold together the two halves of the center section. Lockable vernier micrometer heads to engage these stops were mounted on heavy platforms bridging the sides of the attenuator frame, which was cut down to receive them so that the spindles engaged the stops.

The provision of a stop for setting the local-oscillator power posed a formidable problem. It was not considered practical to reset a micrometer head for each use of this stop. It was therefore necessary to mount the micrometer head for this stop on a hinged platform so that it could be raised to allow the trough-power stop to engage. (The crest-power stop was on the opposite side of the center section, and was set for minimum attenuation, nominally 0 dB.) To obtain a repeatable action for this hinged platform, dowel pins were inserted in holes drilled through both this platform and a fixed locating block while they were clamped together, thus assuring accurate location of the mating members. The platform hinge was sufficiently loose that it did not interfere with the seating of the platform. A cam, operated by a handle on the left side of the modulation attenuator (see figs. 9 and 10), was used to press the platform against the locating block, with a very heavy leaf spring under the cam to maintain the platform position against the stop force, and against a weaker spring that raised the platform when the handle was raised.

A toggle mechanism was used to spring load the stops via a rack-and-pinion coupling between the toggle and the center section. This type of coupling minimized the radial forces that would increase friction in the bearings at the end of the center section. It was hoped that this spring loading would take up any slight bearing clearance and provide a more repeatable structural stress than that resulting from only friction as the operating knob is released.

This toggle mechanism was added after the attenuator appeared to become less repeatable in its operation and sensitive to changes in operating torque. This difficulty was believed to have resulted from a poor fit between the steel rings forming the center-section bearings and the aluminum center section over which they were press fitted. The contacting surfaces of the latter had been made undersized and were evidently dimpled with a punch to displace enough metal to retain the steel bearings. This fit was probably too poor to prevent relative motion between the parts as radial forces were applied due to pressure on the stops (forces for which the attenuator was not originally designed).

To correct this problem, all of the bearing-retaining surfaces were turned down and new bearings made for the center section. The end sections, which were similarly made, were also turned down, but the original bearings were salvaged by the use of bushings between them and the aluminum surfaces. To assure a good bearing fit, the alignment surfaces of the attenuator frame were remachined. The steel bearings were lapped in place and then lubricated with a molybdenum-disulphide compound and firmly tightened to eliminate all discernible end-play. This tight fit increased the driving force, requiring a heavier spring holding the worm against the center-section drive gear. These modifications eliminated the irregularities originally noted.

Isolators on either side of the modulation attenuator were left attached during calibration in order to maintain constant matching and thus preserve the calibration. (Only attenuation increments are of interest; the minimum "insertion" loss was of no concern if not excessive.)

2.3.5 Mixer Load Circuit

Mixer diode specifications generally require different values of d-c (self-bias) load and a-c (intermediate-frequency) load. For the 1N23 X-band type, the specified d-c load is 100 Ω, and the specified a-c load is 400 Ω (purely resistive).

At very low (audio) intermediate frequencies, it is difficult to construct a filter with a d-c resistance less than 100 Ω together with an accurately known a-c impedance. Stray magnetic field pickup is also a severe problem at these frequencies, for which only ferromagnetic shielding is effective. For incremental measurements, no inductive or capacitive filtering can be used at all, requiring a completely different approach.

In the simplified circuit of figure 18, the mixer output is presented with a load consisting of the series combination of a 100-Ω resistor and a 300-Ω resistor. An adjustable current source (a d-c source with essentially infinite internal resistance) is connected in parallel with the 300-Ω resistor. Also connected across the 300-Ω resistor is a d-c null indicator with effectively infinite impedance.

To provide the correct d-c and intermediate-frequency loads, the current source is adjusted to obtain a d-c null across the 300-Ω resistor. This must be done with no modulation or with an antisymmetrical (a-c only) modulation that does not change the d-c mixer output. The mixer terminal voltage is now equal to the voltage across the 100-Ω resistor, through which all of the d-c mixer output current flows. With no a-c component, therefore, the mixer output terminal conditions are the same as though the load consisted of only the 100-Ω resistor. Since the current source and the instrumentation are postulated to have virtually zero admittance, any variation in mixer output current (such as superimposed a-c or incremental change in the d-c) must therefore flow through both resistors. For the intermediate frequency (a-c or d-c increment), therefore, the

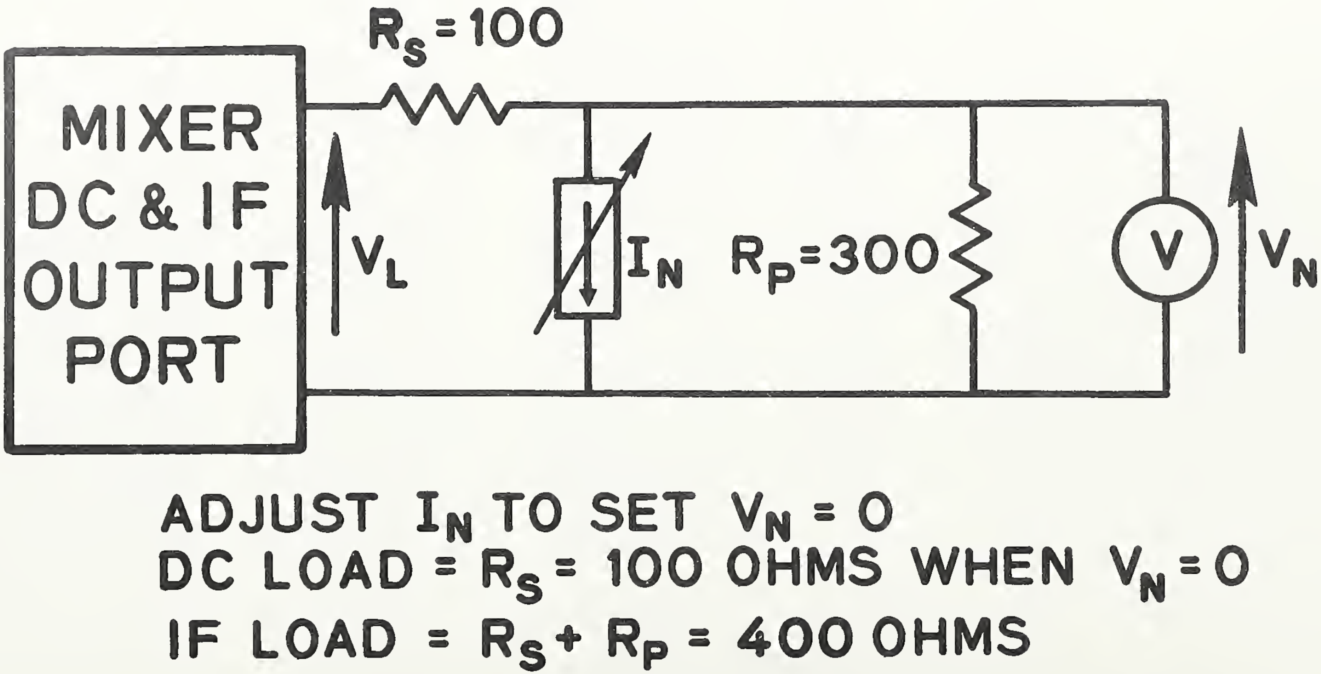


Figure 18. Simplified mixer output circuit to provide distinct d-c and a-c loads by nulling the d-c voltage across part of the a-c load with a constant-current source.

mixer terminal conditions are the same as though the load consisted of both resistors in series.

The complete load circuit shown in figure 19 includes the conductance measurement circuits described in 2.4.1.

The output circuit has four switches: Switch S_L is used to select the appropriate load for each measurement step. Switch S_V selects the appropriate point for voltage measurement; when set in position 6, the points normally required for each load position are selected automatically by one pole of switch S_L. Switch S_T selects the type of conductance measurement circuit, and switch S_M operates the circuit selected.

2.4 Measurement Procedure and Calculations

In the normal operation of the load circuit shown in figure 19, switch S_L is sequenced in order of the numbered positions, with switch S_V in position 6. With S_L in position 1, the digital voltmeter is short circuited, allowing a zero check and adjustment, and grounding one end of the 100-Ω d-c load resistor. In position 2, the DVM is transferred to the mixer output, where it can indicate the self-bias voltage across the 100-Ω d-c load, which still goes to ground. The DVM is read with the specified unmodulated local-oscillator power available to the mixer. In position 3, the 300-Ω resistor is increased by an additional 1200 Ω to improve the null sensitivity. (A higher value, such as an open circuit, would add little additional sensitivity, and is not used in order to keep the impedance level low to reduce pickup and to limit the voltage that could appear across the diode terminals.) The DVM is now across 1500 Ω. The current source, I_N, is then adjusted for a null. In position 4, the intermediate-frequency output conductance of the mixer may be obtained by applying modulation (incremental crest or trough, or periodic) and noting the DVM indication with switch S_M in position 1, then switching to position 2 and restoring this reading by adjusting calibrated resistor R_x (a decade resistance box). In the presence of output drift, switch S_M may then be thrown back and forth periodically and the resistance readjusted until no change in output appears to occur due to this switching. The intermediate-frequency output conductance is the reciprocal of the sum of R_x and the 100-Ω d-c load, but it is always expressed in ohms, i.e., as R_x + 100. This procedure is the same for either the series (Mance's) or shunt circuits, as selected by switch S_T (positions 1 and 2, respectively). For incremental modulation, only small differences have been found in the intermediate-frequency output conductance found by using the modulation crest as opposed to the modulation trough.

With switch S_L in position 5, the a-c load is the sum of R_x and the 100-Ω d-c load resistance, and is therefore the reciprocal of the intermediate-frequency output conductance. Since the output admittance of the mixer is purely real for a zero intermediate frequency and very nearly so for an audio intermediate frequency, the mixer is thus presented with a matched load. The mixer terminal voltage with this match is half the open circuit voltage, allowing the calculation of available

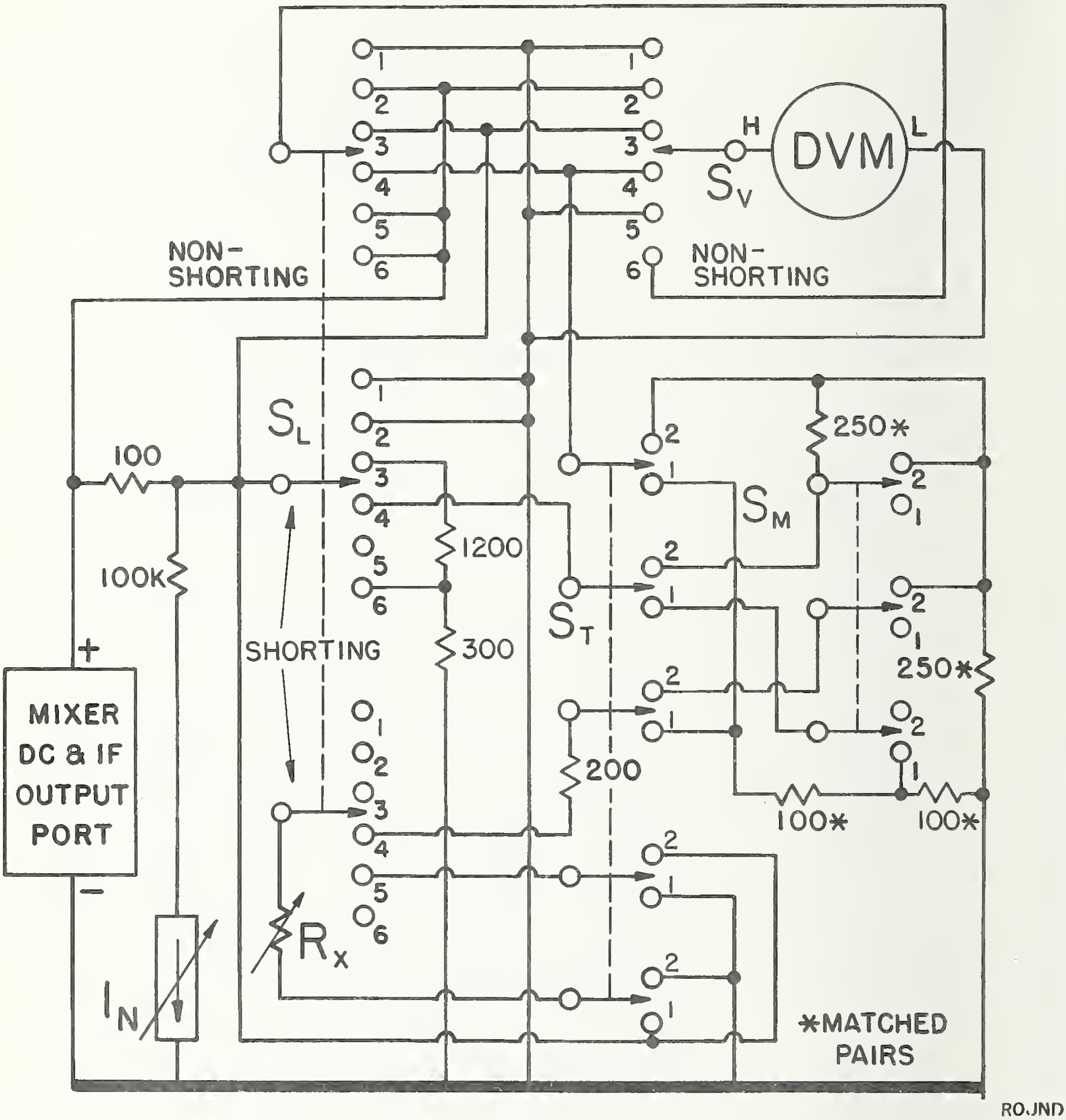


Figure 19. Mixer load circuit. Resistor values are given in ohms.

intermediate-frequency power and, hence, the conversion loss. For incremental modulation, the DVM, which is now across the mixer output, is read first with the modulation at crest and then with it at trough, and the difference used in the calculation. Unfortunately this calculation also requires the value of intermediate-frequency output conductance previously determined, making it difficult to tabulate in advance since a two-dimensional entry would be required. This is no problem, however, if a programmable calculator is available. The equation (derived in Appendix A, with mismatch factor M = 1) is

$$L_m = \frac{8m^2 P_o R}{(\Delta V)^2} \tag{1}$$

where L_m = conversion loss (power ratio)
m = modulation factor (voltage ratio)
P_o = unmodulated available local-oscillator power (watts)
R = intermediate-frequency load resistance (ohms)
(= reciprocal of intermediate-frequency output conductance)
ΔV = incremental or peak-to-peak intermediate-frequency load voltage (volts) corresponding to m.

The modulation factor, m, is obtained from the crest-to-trough power ratio, which is equal to the attenuation, L_a, used to produce (or calibrate) it:

$$m = \frac{\sqrt{L_a} - 1}{\sqrt{L_a} + 1} = \frac{10^{L'_a/20} - 1}{10^{L'_a/20} + 1} \tag{2}$$

where $L'_a = \log_{10} L_a$ (decibels).

With switch S_2 in position 6, the a-c load is 400 Ω, as in the simplified circuit of figure 18. The intermediate-frequency load voltage is obtained as with the matched load (and is, of course, a different value corresponding to a different load). This voltage determines the intermediate-frequency power delivered to the 400-Ω intermediate-frequency load. The calculation of conversion loss differs from the matched-load case in that an intermediate-frequency mismatch factor, M, is introduced into the equation to account for the difference between available and delivered intermediate-frequency powers:

$$L_m = \frac{8m^2 P_o RM}{(\Delta V)^2} . \tag{3}$$

The intermediate-frequency mismatch factor is obtained from the intermediate-frequency conductance, G, as

$$M = \frac{4R(1/G)}{[R + (1/G)]^2} . \tag{4}$$

The above expression for conversion loss is more general than the one given for a matched intermediate-frequency load since, for the latter case, $R = 1/G$, and $M = 1$. The advantage in using a fixed load resistance ($R = 400\ \Omega$) and a separate mismatch factor is that each of these two factors in the calculation can be tabulated in one-dimensional form, i.e., requiring only a single entry. The division of the original single expression (for a matched load) into these particular factors (of the many pairs that could have been formed) has two advantages.

First, the mismatch factor, M, is unity for the matched case and very close to unity for a reasonable range of intermediate-frequency conductance. For a two-to-one ratio in either sense between load resistance and the reciprocal of intermediate-frequency conductance, $10 \log_{10} M$ ($= M'$) is about -0.5 dB. For the intermediate-frequency conductance reciprocal range limits specified for the 1N23WE and 1N23WG diodes of 335 to 465 Ω, M' is about -0.035 dB. For many measurements, therefore, M may be considered unity ($M' = 0$ dB) with little error.

The second advantage of this formulation is that the fixed-load loss equation with M omitted has an important physical interpretation: It is analogous to the microwave concept of "insertion loss" of the mixer, which is defined (for example, in the IEEE Standard Dictionary of Electrical and Electronics Terms) as the factor by which the power reaching a line-matched load is reduced when the transmission line connecting this load to a line-matched generator is broken and the device under test inserted. Insertion loss thus accounts for power lost due to mismatch as well as dissipation within the device. This original concept of insertion loss can obviously not be literally applied to mixer conversion loss measurements because of the different input and output frequencies involved; no loss-free frequency converter exists as a physical reference analogous to the straight-through feed in the single-frequency case. The basic concept of holding the device under test accountable for mismatch to both source and load, each having a specified fixed immittance, however, is perfectly applicable. (The IEEE has also defined the term "transducer loss" to have this same basic definition, but this is a debatable use of the term "transducer," which has more value in another sense.)

There are good reasons, in fact, for preferring to define conversion loss in this way rather than by the conventional available-gain concept in which the mixer is held accountable for only the source mismatch. In field use, when mixer diodes are replaced, there is generally no retuning on either the r-f or intermediate-frequency sides; the resulting performance degradation caused by this mismatch is thus a part of normal operation. Using the mismatch factor to correct for the increased loss resulting from an intermediate-frequency conductance that is not nominal is therefore unrealistic in terms of the best characterization for field-performance prediction. A good device characterization should, as its most fundamental criterion, be capable of ranking devices in the order of their field performance. Consider two diodes of differing conversion loss as conventionally defined: If the one exhibiting a lower loss has an intermediate-frequency conductance nearer to a specification limit, it may prove inferior to the other in field performance by having a larger intermediate-frequency mismatch. Fortunately, the best-grade modern diodes have sufficiently tight intermediate-frequency conductance limits that the difference would be very small, as has been shown above, and there are more important limitations on the accuracy of field performance prediction, such as frequency and holder differences. For conversion loss measurements of the modulation type, it is obviously easier to use only the insertion loss factor, although for the noise substitution measurement method (not considered here) the reverse is true.

Although the insertion loss definition must be considered further, particularly with respect to noise measurements, before it can be recommended for general use, it does seem sufficiently important to warrant giving a distinct title to conversion loss defined on this basis. The term "conversion insertion loss" is recommended (see Appendix D). Conversion loss (based upon available powers) can be said to consist of the product of an "insertion loss factor" and a "mismatch factor", where all terms are power ratios, or the sum of these factors, where all terms are expressed in decibels, with appropriate signs.

2.4.1 Intermediate Frequency Output Conductance

It has been traditional to specify limits on the "intermediate-frequency impedance" of a mixer diode. The parameter generally measured, however, has been a resistance, measured at 60 or 1000 Hz, and the specified limits are always purely real. This measured value has commonly been used in the calculations required for 30-MHz intermediate-frequency noise measurements made with a temperature-limited thermionic diode, where the approximate expression "20 IR" is used. The "I" in this expression is the d-c anode current of the noise diode; the "R" required is actually the reciprocal of the real part of the mixer output admittance. (The exact expression is $eI/2kT_oG$, where e is the elementary (electronic charge), k is Boltzmann's constant, T_o is a standard reference temperature, and G is the output conductance. A complete derivation is beyond the scope of this report, but the conductance term is due to the use of Norton's equivalent circuit for the mixer as necessitated by the noise-current source and effectively zero shunt conductance representing the noise diode. The two parallel current sources can thus be combined; an available power computation effectively cancels the total susceptance, leaving only the mixer conductance.)

As the intermediate frequency approaches zero, the mixer output impedance and admittance become purely real, and their real parts thus become reciprocal. For a low intermediate-frequency measurement, therefore, it is immaterial whether the measurement is one of conductance, resistance, admittance, or impedance, since these (or their reciprocals) are all nearly identical in value when expressed in the traditional units of ohms. As the intermediate frequency is raised, it is the bypass and stray capacitance shunting the conductance that will first disturb this identity; at 30 MHz, some series inductance will be present, but the intermediate-frequency line lengths are a very small fraction of the (approximately 10-m) wavelength, so that the parasitics are still largely susceptive.

It is expected, therefore, that the conductance measured at a very low intermediate frequency is a close approximation to the 30-MHz conductance, provided that the corresponding r-f signal source immittance is unchanged (e.g., a broadband match). Taub [6] has stated that "... within the accuracy of the measurements mixer output conductances are essentially constant from 60 cps (60 Hz) up to at least 30 mc (30 MHz)." The low frequency resistance or impedance, however, would not be expected to approximate the 30-MHz resistance or impedance, which is an additional reason why conductance is the preferred parameter. In some military specifications [12,13,14], the term "intermediate-frequency conductance" has now replaced the older term "intermediate-frequency impedance," (though not in all [15]) but the unit is still the ohm (for 1/G).

The use of a bridge to measure intermediate-frequency output conductance at low frequencies was originally contemplated, as was the use of a 60-Hz comparison method (equipment for which was transferred from NASL), in which the voltage drop across the mixer resulting from a small 60-Hz current is matched by the drop across a decade resistance box switched in place of the mixer, with the same value of current. The advantage of the comparison method is that the current source and detector can both have a grounded terminal, whereas a bridge requires that one or the other be floated. The bridge and comparison methods share the disadvantage of requiring an intermediate-frequency generator of arbitrary output voltage (or current) and frequency. Unless battery power were used, such a generator would introduce an additional power-line connection with its attendant grounding problems. The NASL equipment was known to suffer from a problem of 60-Hz pickup by a tuned choke used to obtain the required 100-Ω d-c load without unduly reducing the measurement sensitivity. The less-than-zero admittance of this resonant circuit, however, required that it be left across the detector, with the decade resistance box switched in place of only the mixer; the resultant opening of the mixer output is potentially dangerous as it can result in large transients across the diode.

While the problems associated with the bridge and comparison methods could eventually be overcome, it was felt that a more elegant method could be found by using the normal intermediate-frequency output voltage as a signal source, i.e., by measuring the mixer as an active circuit, rather than a passive one, via the application of the same r-f signal (local-oscillator modulation) used for the conversion loss measurements and, hopefully, obtaining both the intermediate-frequency output conductance and the conversion loss together from the same measurement.

The essence of an active circuit (generator) measurement is to vary the load and observe the resulting change in terminal voltage, in load current, or both. The conceptually simplest method is to obtain the open-circuit voltage and short-circuit current, their ratio being the source immittance. Another conceptually simple method is to vary the load resistance until the terminal voltage is half the open-circuit voltage or until the load current is half the short-circuit current, after first varying the load reactance for maximum output, if an a-c measurement is being carried out. The load impedance obtained in this way is the complex conjugate of the source impedance.

These methods, while conceptually simple, present practical difficulties for mixer intermediate-frequency output conductance measurements. For the range of conductances to be measured, a sensitive current measurement is impractical, since the meter impedance would be a large fraction of the normal load, and a short-circuit current measurement impossible (except by use of a balancing circuit requiring an external source of identical frequency with adjustable magnitude and phase, thereby destroying a major advantage sought for this method of not requiring an external generator).

The half-open-circuit voltage method was tempting, but it was not known if the mixer output voltage-current characteristics were sufficiently linear to permit going to an open circuit. In addition, for the 1000-Hz intermediate frequency originally used, it was feared that the mixer bypass and stray capacitive susceptance, while negligibly small compared with the normal 400-Ω intermediate-frequency load, might appreciably compromise an "open-circuit" load.

Open- and short-circuit conditions are not required for a source-immittance determination. The load voltage, V, can be written in terms of a generator's open-circuit emf, E, source impedance, z, and load impedance, Z: $V = EZ/(z + Z)$. For a different load impedance, Z', the load voltage can be written: $V' = EZ'/(z + Z')$. These equations may be combined and solved for E and z:

$$E = \frac{Z' - 1}{\frac{Z'}{V'} - \frac{Z}{V}} = \frac{Z}{\Delta\left(\frac{1}{I}\right)} \tag{5}$$

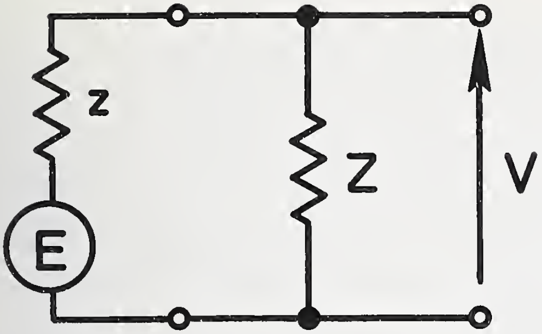
$$z = \frac{V' - V}{\frac{V}{Z} - \frac{V'}{Z'}} = - \frac{V}{\Delta I} \tag{6}$$

where Δ indicates the difference between primed and unprimed values of the variable in question and I is the load current. The equations for the conceptually simple methods mentioned earlier can all be obtained as special cases of these. This method was successfully tried with a 1000-Hz intermediate frequency for a number of load resistance pairs over a wide resistance range, with one member of each pair being the nominal 400 Ω, and also with 400 Ω as the geometric mean of each pair. The observed variation in the calculated values of E and z seemed to be small and uncorrelated with the load variation, but the voltage resolution at the time was rather poor.

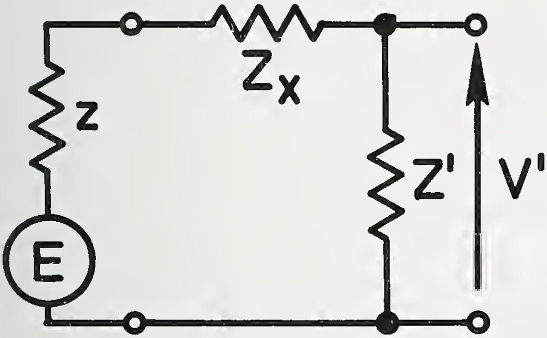
It was thought that accuracy of the load perturbation method could be improved if the voltmeter were used at only a single voltage by providing an adjustable tap on the load. As the load impedance was increased, increasing the terminal voltage, the voltmeter would be moved from its initial position at the terminal to a lower tap where the voltage was the same as the initial terminal voltage. The circuits for the two states are shown in figure 20, along with the calculation of source impedance. As shown in the figure, the denominator in this expression, Z' - Z, can be physically realized by an impedance Z_y in series with the initial load, Z, and shorted out, along with Z_x, for the initial reading. It was seen that this circuit could be <u>direct reading</u> (no calculations required) if, as shown in the figure, Z_x is made variable, and Z and Z_y proportioned so that their ratio is an integral power of ten, thereby forming a decimal multiplier for the value of Z_x indicating z. Since use of a very-high impedance voltmeter was contemplated, it was thought that a single-pole switch would suffice for effectively removing the added impedances Z_x and Z_y. This was a fortunate choice, because subsequent analysis showed that at balance (V' = V) the impedance presented to the voltmeter by this circuit is independent of switch position. A voltmeter with a finite impedance could therefore be used, provided that a trial-and-error procedure (switching back and forth repeatedly while adjusting Z_x for a constant voltmeter reading) was acceptable.

The properties of this circuit seemed so remarkable that a search was made for it in the literature. It was found [16-20] under titles such as "Mance's method for measuring battery resistance," but no reference to its origin has been found other than the name of its inventor, Sir Henry Mance. It is similar to Kelvin's method for determining galvanometer resistance [21]. These texts treat the circuit as a balanced bridge, where the switch and detector are in "opposite" arms which are isolated when the other four arms are equal or appropriately proportioned. Perhaps a simpler view, when an infinite impedance voltmeter is used, is that the voltage division between the source impedance and the load is not disturbed if the impedances are added in series to each of these in the same proportion, i.e., if $Z_x/Z_y = z/Z$ (fig. 20). Strangely, Mance's method seems to have been forgotten in recent times, even for battery resistance measurements, the last known reference to it appearing in 1912 [18].

In Mance's method, the load perturbation can be made as small as desired by the appropriate ratio of Z_y to Z, although the balance sensitivity is correspondingly reduced. It was originally intended to have Z_y = 0.1 Z or 0.01 Z, so that z = 10 Z_x or 100 Z_x. The intermediate-frequency voltage was later found to be insufficiently stable for this, requiring the higher sensitivity of a unity arm ratio. With a high arm ratio, the decade resistance box used for R_x would have to be read to 0.1 Ω or 0.01 Ω in order to obtain z to the nearest ohm. It would thus be highly susceptible to errors caused by contact and residual resistances, and be limited by the relatively lower accuracies generally obtainable with low-resistance decade boxes. An attempt was therefore made to find a direct reading circuit in which small load perturbations could be obtained by use of a decade resistance box larger than the source impedance by a power of 10, rather than smaller. It was thought that this could be done by adding impedances in shunt with the original load rather than in series. The dual of Mance's circuit yielded such a shunt circuit, but was unusable because it required a detector of zero impedance to avoid a trial-and-error procedure.



$$\frac{V}{E} = \frac{Z}{z+Z}$$



$$\frac{V'}{E} = \frac{Z'}{z+Z_x+Z'}$$

If $V' = V$:

$$\frac{Z'}{z+Z_x+Z'} = \frac{Z}{z+Z}$$

$$z = \frac{ZZ_x}{Z'-Z} = Z_x\frac{Z}{Z_y} = Z_x(10^n)$$

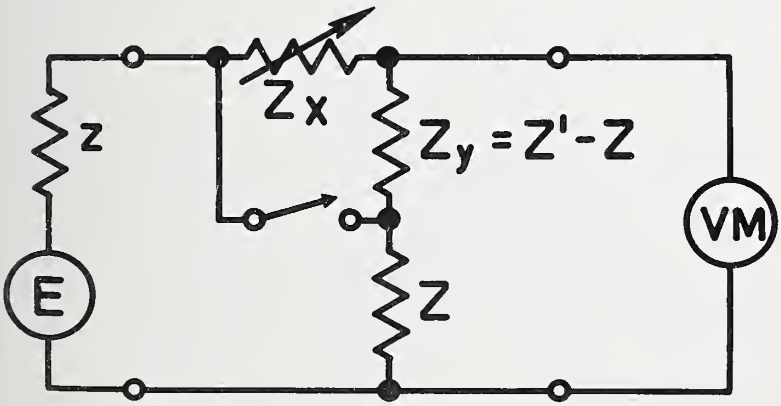


Figure 20. Derivation of direct-reading, series type, load-perturbation method of determining a source impedance (Mance's method).

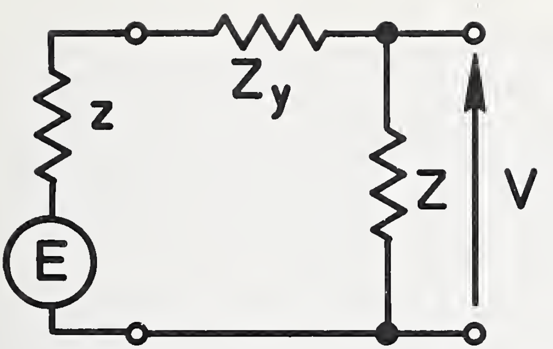
As shown in figure 21, two different ways of reducing the voltage across a given load impedance by adding series or shunt impedance may be alternated and one of the impedances varied to equate the outputs. If the shunt impedance is varied, it can directly indicate the source impedance provided that the ratio of Z to Z_y is an integral power of 10. This circuit requires a double-pole switch, as shown in figure 21. It can be seen that for small load perturbations Z_x is larger than z, rather than smaller as in Mance's method. Unfortunately, unlike Mance's, this circuit does not present a constant impedance to the voltmeter at balance, thus requiring an infinite voltmeter impedance.

As has been mentioned, practical considerations of voltage stability dictated a unity arm ratio ($Z_y = Z$, $z = Z_x$). The arm resistances (Z_y and Z) were calculated for both circuits to obtain load resistances which varied with geometric symmetry about the specified 400-Ω intermediate-frequency load. For Mance's circuit, the resulting perturbation factor is 2, i.e., the load varies between 400/2 = 200 Ω to 400(2) = 800 Ω. For the shunt circuit, the perturbation factor is 1.5, i.e., the load varies between 400/1.5 = 266 2/3 Ω to 400(1.5) = 600 Ω. Sensitivity calculations indicated that, for a given error in setting Z_x, the resulting changes in output voltage for the two methods as the switches are thrown are roughly proportional to their perturbation factors. Mance's circuit is thus somewhat more sensitive, reducing random measurement error, but at the expense of a larger load perturbation (equivalent to a larger signal) which could possibly result in a greater systematic error (bias) due to mixer nonlinearity.

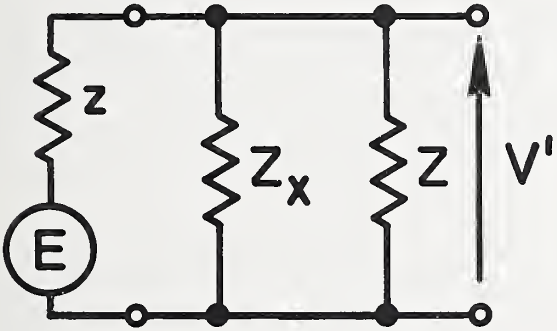
To compare the load-perturbation circuits, they were both incorporated in the load circuit shown in figure 19. The unity arm ratio permitted the decade resistance box to be used directly as the load (allowing twice for the added 100-Ω d-c load resistance) when measuring conversion loss in a single step (requiring a matched load). Originally, it was expected that a separate decade resistance box would be required, with the readings transferred with an appropriate shift in decimal point position — a procedure that would have introduced additional measurement uncertainty and appreciably lengthened the measurement procedure.

3. RESULTS

The results of the work to date fall into two categories: (a) hardware, and (b) software. The hardware consists of the completed portions of the measurement system, which can be evaluated in terms of the associated measurement uncertainties. The software consists of the measurement techniques and engineering refinements introduced or developed in the course of this work that may be considered advances in the mixer measurement art.



$$\frac{V}{E} = \frac{Z}{z + Z_y + Z}$$



$$\frac{V'}{E} = \frac{\dfrac{Z_x Z}{Z_x + Z}}{z + \dfrac{Z_x Z}{Z_x + Z}}$$

If $V' = V$:

$$\frac{Z_x Z}{z(Z_x + Z) + Z_x Z} = \frac{Z}{z + Z_y + Z}$$

$$z = Z_x \frac{Z_y}{Z} = Z_x (10^n)$$

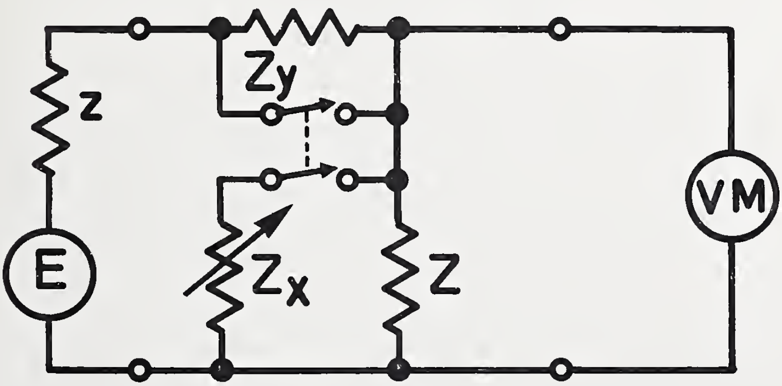


Figure 21. Derivation of direct-reading, shunt type, load-perturbation method of determining a source impedance.

3.1 Conversion Loss Measurement Uncertainties

3.1.1 Systematic Measurement Uncertainty

The overall systematic measurement uncertainty is determined in the manner of a total differential:

$$dy = dx_1 \frac{\partial y}{\partial x_1} + dx_2 \frac{\partial y}{\partial x_2} + \ldots , \qquad (7)$$

where the partials represent the sensitivity of the measured parameter, $y = f(x_1, x_2 \ldots)$, to each of the error sources, x_i, whose uncertainties are represented by the independent differentials (dx_i).

This is reasonably valid for finite uncertainties as long as they are small compared with the measured parameter. It has been argued that independent uncertainty terms should be combined quadratically rather than linearly, but where the number of significant terms is small and the probability distribution of each is unknown, it would seem advisable to use the more conservative linear addition.

In the determination of conversion loss by the modulation techniques, two major sources of systematic uncertainty have been identified to date. The first of these is the attenuation of the modulator. It was found that the partial derivative of conversion loss with respect to total modulator attenuation (crest to trough) is a function only of attenuation, and not of conversion loss or any other mixer parameter, provided that the conversion loss is expressed in decibels:

$$\frac{\partial L'_m}{\partial L'_a} = \frac{2}{10^{L'_a/20} - 10^{-L'_a/20}} \qquad (8a)$$

$$\simeq \frac{8.69}{L'_a} \text{ dB/dB} \qquad (8b)$$

where $L'_m = 10 \log_{10} L_m$ = conversion loss (decibels)
$L'_a = 10 \log_{10} L_a$ = attenuation (decibels)
and L_m = conversion loss (power ratio)
L_a = attenuation (power ratio).

A probably quite conservative systematic uncertainty of ± 0.005 dB was tentatively assigned by the former Electromagnetics Division* at the NBS Boulder Laboratories to their calibration of the 1-dB attenuation increment. Using the sensitivity of 8.67 dB/dB obtained as the exact partial derivative, this is equivalent to a conversion loss uncertainty of ± 0.043 dB.

The second major source of systematic uncertainty that was identified is the local-oscillator power. The modulation method of measuring conver-

* This work is now the responsibility of the Electromagnetic Technology Division of the Center for Electronics and Electrical Engineering.

sion loss is doubly sensitive to local-oscillator power: First, the actual conversion loss is in itself a function of local-oscillator power, as are all of the other mixer parameters. Second, the signal power generated by modulation of the local oscillator is directly proportional to the local-oscillator power, since the modulation factor is fixed. Unfortunately, the measured conversion loss is changed in the same direction by both these factors as local-oscillator power is changed.

The variation of actual (but nonstandard) conversion loss with local-oscillator power is a complex function, and no theoretical analysis has been attempted, but the apparent change in conversion loss due to the signal change resulting from a given local-oscillator power change can be obtained from the measurement equation, using the same approach as with attenuation, as shown in Appendix B. The sensitivity of measured conversion loss to local-oscillator power, P_o, considering the true loss to be constant, is

$$\frac{\partial L'_m}{\partial P_o} = \frac{4.34}{P_o} \text{ dB/W} . \qquad (9)$$

An empirical study of the total sensitivity of conversion loss to local-oscillator power, including both the actual and apparent changes, was made by deliberately introducing power errors for every fourth measurement in a series of diode measurements. The absolute magnitude of this sensitivity was determined to be about 0.0047 dB/μW for a local-oscillator power of 1 mW, as compared with 0.0043 dB/μW calculated from eq (9), and showed no significant difference from one diode to another. (Only point-contact diodes were used.) The close agreement with the calculated measurement error and the consistency between diodes both seem to indicate that the variation of actual conversion loss with local-oscillator power is negligible by comparison with the measurement error due to signal power variation over the same small local-oscillator power range. Whatever small true loss sensitivity that does exist will be the same for all measurement methods, regardless of how the signal is produced, so that it may be largely ignored in intercomparing the different methods. Only the 4.34 dB/W factor is unique with the modulation method.

The local-oscillator power was set by a trial-and-error measurement using a dual semi-automatic power bridge and a group of compensated bolometers (mounted dual thermistors) each of whose calibration factor (available r-f power from a line-matched source/substituted d-c power) was determined by the Electromagnetics Division at the NBS Boulder Laboratories.

The uncertainty in d-c substituted power was stated by the bridge manufacturer to be ± 0.1 percent. The bolometer calibration factor 3σ uncertainty was stated to be ± 0.7 percent. The individual bolometers were repeatable to about ± 0.05 percent, including flange contact repeatability; this became a systematic uncertainty factor when the bolometers were used to determine an output voltage from the monitor calorimeter used to hold the power constant. A random error in this determination thus became a constant (systematic) error in the local-oscillator power until the next determination was made. These uncertainties totaled ± 0.85 percent; using the 0.0434-dB/percent sensitivity, this was equivalent to a conversion loss uncertainty of ± 0.037 dB. An additional possible power uncertainty whose significance was not determined was the spread in the power (totaling several microwatts) determined by the several calibrated bolometers (currently five). It is not known if this spread can be included in the quoted ± 0.7-percent uncertainty in calibration factor. If not, then the total power uncertainty may have been as large as ± 1.2 percent, equivalent to a loss uncertainty of ± 0.052 dB.

The conversion loss uncertainty from the two factors considered so far total at least ± 0.080 dB (± 0.095 dB if the bolometer spread is included). If the estimated limit of variation of true conversion loss with power is included as well, the total would be about ± 0.10 dB.

The other systematic uncertainty sources that have been considered make negligible contributions to the total. As shown in Appendix B, the sensitivity of conversion loss to intermediate-frequency load resistance is (4.34/R) dB/Ω. For a resistance uncertainty of ± 0.1 percent, this causes only a ± 0.004-dB loss uncertainty. The sensitivity to incremental output voltage determination is (-8.69/ V) dB/V. For a conservative ± 0.02-percent voltage increment uncertainty (using a ± 0.01-percent voltmeter), the systematic contribution is only ± 0.002 dB.

From charts prepared to evaluate the intermediate-frequency mismatch factor, M, it can be shown that a systematic ± 0.1-percent uncertainty in load resistance and hence in measured intermediate-frequency conductance reciprocal results in a negligible ($<\pm 0.001$-dB) conversion loss uncertainty.

In the series of diode loss measurements during which the power sensitivity was ascertained, every fourth measurement was performed with a deliberate error in the nulling procedure used to establish the correct d-c load of 100 Ω. (The 400-Ω intermediate-frequency load was not affected by the null error.) This nulling error was expected to affect the conversion loss by altering the self bias on the mixer diode and thereby changing the r-f match (among other things). The results of this experiment showed no detectable effect for any reasonable null error. For most diodes, the conversion loss appeared to be slightly higher for one or both extremes in null error (± 25 mV); for some, a slight improvement was noted, but in all cases where any significant change could be seen (beyond the random fluctuations in repeatability), the null errors were several orders of magnitude larger than those normally occurring. Since the uncertainty due to nulling error is diode-dependent, it is particularly fortunate that it is small enough to be considered negligible.

The effect on conversion loss of errors in r-f tuning were not methodically investigated, but the mixer source standing-wave ratio (SWR) appears to hold within about 0.01 over a prolonged period. The power reflected due to a 1.01 SWR mismatch is only 0.0025 percent, and is therefore not expected to have a significant effect on the mixer parameters as compared with the effect of the much larger local-oscillator (and signal) power uncertainty. The much larger reflections expected at the local-oscillator harmonic frequencies and at frequencies near these harmonics constitute a potential source of significant systematic error. This source is the most difficult to deal with of any of the error sources identified, since the sensitivity of conversion loss to these reflections is completely unknown, as are the reflection magnitudes. As in the case of actual conversion-loss variation with local-oscillator power, however, the harmonic match is expected to affect all measurement methods equally, assuming that the line match near the harmonics is not exceedingly frequency sensitive so that the harmonic-mixer conversions are independent of the intermediate frequency used.

3.1.2 Random Measurement Uncertainty

The random measurement uncertainty was determined by a statistical study of the results of many repetitions of the measurement. "Repeatability" studies of mixer measurements are complicated by the fact that the diodes tend to be unstable. Some diodes are exceedingly sensitive to mechanical shock and cannot be ejected from the holder and reinserted without a change in their characteristics. It was noted that the mechanical shock transmitted to the mixer by operating a waveguide switch near it (see figs. 3 and 7) was sufficient to affect some diodes even after the waveguide was tightly clamped. Static electrical discharges through the diode, resulting from improper handling, can also affect a diode, even to the point of destroying it.

Even with the most careful handling, some diodes are far less repeatable in conversion-loss measurement than others. Unfortunately, there is no abrupt difference in repeatability to allow "stable" diodes to be easily distinguished from "unstable" ones. There is thus no way to completely separate the random variations of the measurement system from those of the diode. The total random uncertainty, system plus diode, is certainly of practical interest, but if the quality of the measurement system is to be evaluated, system performance must be isolated. An empirical approach to this isolation problem is to select the most apparently stable diodes from measurements of many diodes, to remeasure the selected diodes repeatedly, and then to attribute the statistical variability of these measurements entirely to the measurement system. This scheme was followed using point-contact diodes, but the number of diodes available at the time was limited, and defects in the modulation attenuator (described earlier) were subsequently found, interrupting the final, long-duration measurement run.

Every available point-contact diode (13 in all) having a conversion loss less than 6 dB was measured over a 24-h interval. Of the 35 conversion-loss measurements generally made on each diode during each run, usually started in an afternoon and completed the following morning, 17 (every other one) were made under standard conditions. The remaining 16 were used for empirical sensitivity determinations, eight for power sensitivity and eight for null sensitivity, the results of which have already been described.

The data taken under standard conditions were analyzed in various ways: In five of these daily measurement runs, more than 90 percent of the conversion loss values were within ±0.02 percent of the mean for the run. (In one run, all of the loss values were.) The sample standard deviations for all runs ranged from 0.006 dB to 0.198 dB (0.13 to 4.55 percent of the respective means), with an average of 0.023 dB (0.48 percent). These figures are based upon all data, with no outliers (grossly atypical data) removed. Removing one outlying datum reduces the largest standard deviation from 0.198 dB (4.55 percent) to 0.011 dB (0.24 percent), about equal to one of the smallest. The appearance of one or two such outliers in an otherwise compact group seemed to be typical.

One of the most apparently stable diodes was reinserted in the holder and remeasured for two additional 24-h runs without being removed from the holder (a weekend intervened between these two runs). After ejection and reinsertion (at a later date), it was remeasured repeatedly over a one-week period without being removed from the holder (with fewer measurements per day than were obtained previously). Considering these last measurements as constituting a single run, the four runs exhibited the following: the mean values of conversion loss, in chronological order, were 4.394, 4.373, 4.372, and 4.370 dB, showing a monotonic decrease totaling 0.024 dB; sample standard deviations corresponding to these means were 0.021, 0.019, 0.012, and 0.018 dB, respectively. Expressed as a percentage of their respective means, these standard deviations were 0.40, 0.43, 0.27, and 0.41 percent. The independently rounded corresponding 3σ values were 0.063, 0.057, 0.036, and 0.054 dB; or 1.44, 1.29, 0.81, and 1.23 percent, respectively. The measurements of this diode were interrupted when it became apparent that the modulation was sensitive to operating torque, leading to the repairs and modifications previously described. Disassembly of the attenuator led also to the discovery that the micrometer head used for the 1-dB stop had not been clamped tightly enough, allowing it to be unseated by about 0.003 in., causing a systematic measurement error (reducing the measured loss) estimated to be about 0.25 dB. In checking the data obtained for diodes for which more than one measurement run was made, none exhibited a change of this order of magnitude, but smaller movements may have caused the 0.034-dB decrease in the means for the selected diode mentioned earlier.

In an attempt to eliminate or at least reduce the measurement variations caused by diode instability, a Schottky-barrier diode was substituted for the point-contact type 1N23. Since it was desired to use the same holder, a Schottky diode was selected which had the same external package dimensions, and which was intended by the manufacturer to be a direct replacement for the 1N23 in many applications. Unfortunately, the Schottky diodes in this type of package are unbonded; i.e., they use a whisker, similar to that used in point-contact diodes, for one contact. The whisker apparently can move, altering the electrical characteristics. Particularly unfortunate was the choice of brands; the brand chosen for a long-term repeatability study was subsequently found to be unusually unstable, and appeared to be comparable to the point-contact diodes in this respect. The repeatability study made with this diode had to be terminated after the mechanical (shock and vibration) sensitivity of the diode led to its destruction. The substitution of a different brand diode was found to make a great improvement, although some mechanical instability is expected to be present in all unbonded diodes.

Before a long-term study could be made with this second brand of diode, a new project requiring repeated changing of diodes was begun using the NBS X-band measurement system. This project was the determination of the radiation sensitivity ("hardness") of Schottky-barrier diodes. A secondary objective of this project was the determination of the random measurement uncertainties for practical measurement conditions, for which diode ejection and reinsertion were necessary, and for different brands of diodes. A report on this work [22] has been issued and should be considered as an essential adjunct to this report, as it provides the best evidence for the success of the mixer measurement system as a practical tool. The following conclusions based upon the radiation-hardness work may be of interest:

The diode contacts for this type of package do not appear to present a significant problem, provided that angular orientation is preserved between measurements.

It is desirable to cycle the modulation attenuator from crest to trough three times and to use the median value of the three resulting mixer output voltage increments. This procedure tends to reject badly erroneous values (outliers) resulting from momentary electrical transients, faulty attenuator operation, voltmeter misreadings or drift, diode changes, etc. A comparison was made between the medians of different-sized data groups obtained in this way, for up to eleven attenuator cycles. It was found that the medians of sets of three cycles had a significantly lower standard deviation than had the same number of single measurements for which only one attenuator cycle was used for each measurement. (The power and output circuit balance were reset between separate entire measurements.) Using more than three attenuator cycles, however, did not make additional detectable improvements. A theoretical justification for this procedure is given in Appendix C. There are still unanswered questions concerning the effect of this median selection on the statistical results. If the triple attenuator cycle is made a permanent part of the measurement procedure, however, then there can be no objection to using the statistics of the data (medians) obtained in this way to characterize the system performance.

3.1.3 Measurement Uncertainties for Other Parameters:

The intermediate-frequency output conductance measurements were found to be repeatable to within a range of 2 to 3 Ω without extensive trial-and-error. As expected, the series (Mance's) method seemed slightly more repeatable. Over the 1N23WE and 1N23WG intermediate-frequency output conductance reciprocal (R) limits of 335 Ω to 465 Ω, each ohm of error generally results in a mismatch error of less than 0.001 dB. The roughly $\pm$1-Ω or $\pm$2-Ω random uncertainty would thus make a generally negligible addition to the total conversion loss uncertainty ($\pm$0.001 dB or $\pm$0.002 dB at the conductance limits; much less for most diodes).

All output conductance reciprocal data for the radiation-hardness study were taken as the averages of the pairs of data obtained by using first the crest power and then the trough power to produce the output voltage needed for this measurement. Consistent differences of several ohms were frequently noted with some brands of diodes, although for other brands the differences were slight or absent.

On the several occasions when the system was retuned, very little residual reflections were noted, probably corresponding to a source match of about a 1.01 SWR, or better. For the expected mixer mismatches (1.3 is the SWR limit for the 1N23WE and 1N23WG diodes), this tuning can be considered perfect. For small values of SWR, the measured SWR can be in error by as much of a factor as the source SWR; i.e., the true SWR of the load (mixer) will be measured within a range obtained by multiplying and dividing the load SWR by the source SWR. A value of 1.01 for the latter will therefore result in a load SWR of 1.30 being measured (with no other errors) within a range of 1.29 to 1.31, depending upon the relative phase relationship between the two reflections. A requirement for a closer mixer measurement than this seems unlikely, as the power uncertainty corresponding to a 1.01-SWR source is less than ± 0.0025 percent, for any load, and this power is less than 0.1 percent of that reflected from a 1.3-SWR load. Neglecting the small tuning uncertainty, the uncertainty in return loss measurement is equal to the uncertainty associated with the calibrated variable attenuator of the reflectometer plus the uncertainty in equating power levels. The latter is dependent upon power stability, and also upon zero and gain stability of the detector-galvanometer system. The measurements seemed to be repeatable almost to within the dial resolution of the (rotary vane) attenuator.

3.2 Contributions To The State-Of-The-Art

Most of the work in this program was in system design, construction, and evaluation. The experimental phase that was expected to resolve differences between modulation and noise measurements, etc., was not reached. However, some contributions to the mixer measurement art resulting from this work may be identified, some of which may be useful in other fields as well:

(1) It has been demonstrated that an attenuator with adjustable stops may be used to produce or calibrate a bilateral incremental modulation which simulates to a high degree of accuracy the crest, trough, and carrier powers of sinusoidal modulation.

(2) Equations for the incremental modulation measurement method have been derived from those for the periodic modulation method — a simpler and more intuitively obvious approach than has been previously used.

(3) A method has been developed for measuring the peak-to-peak value of an alternating voltage, in the presence of superimposed dc, by a precise comparison with a known adjustable d-c source. This method may have far wider applicability than for mixer measurements.

(4) Using essentially the same technique as for the a-c measurement of (3), a method has been developed for measuring the modulation factor of a modulated r-f source using a calibrated variable r-f attenuator.

(5) Equations have been developed for calculating conversion loss uncertainty. Of particular importance is the sensitivity of this parameter to modulator attenuation uncertainty.

(6) A novel reflectometer circuit has been introduced which uses a circulator to separate the incident and reflected waves. This reflectometer has a lower power loss than conventional directional coupler types and is thus more sensitive (an important consideration for low power c-w measurements).

(7) Two direct-reading load perturbation methods for measuring intermediate-frequency output conductance have been introduced. One of these is a long-forgotten method originally used for measuring battery resistance (Mance's method). These methods may be generally useful for measuring the internal immittance of any quasi-linear or high-power source.

(8) A data-smoothing technique has been developed in which the measurement, or part of the measurement, is repeated several times and only the median of the obtained data is then used. This method should have wide general applicability. An analysis of the technique is presented in Appendix C.

4. Conclusions

(1) With available equipment, incremental modulation is distinctly superior to periodic modulation for standards-quality measurements of conversion loss.

(2) Fundamental limitations on attenuation and power calibrations (roughly equal in importance) currently establish a systematic uncertainty in modulation-type conversion loss measurements of very close to ± 0.1 dB.

(3) A three-sigma random uncertainty (repeatability) of about ± 0.05 dB appears to be achievable for stable diodes. The reasonableness of this figure, originally estimated from point-contact diode data, was confirmed by the Schottky diode measurements made for a radiation-hardness study [22]. The total conversion loss uncertainty thus appears to be about ± 0.15 dB.

(4) If an overall average noise figure is desired, the uncertainty in the output noise ratio measurement must be added to the figure for total conversion loss uncertainty given in (3). Even if this noise measurement uncertainty were negligible, the resulting overall uncertainty is a good fraction of the 0.5-dB separation in overall average noise figure limits between diode grades (suffix-types). This uncertainty is probably not significant, *per se*, as far as most device users are

concerned, but may be critical to the device producer for the following reason: For the better grade diodes, the yield falls off precipitously with reduced limits on overall average noise figure, so that most diodes of a given grade (suffix) are close to the acceptance limit. Therefore, small change in the limit can change the yield of that grade diode far out of proportion to the ratio of the change to the 0.5-dB width of the acceptance "slot." A tightening of the limit by 0.1 or 0.2 dB may very well reduce to zero the yield of the best grade subtype. The economic impact of this yield reduction could be considerable, since sales price varies considerably with diode grade. (As an extreme example, a 35-GHz mixer diode type was advertised at $150, $200, and $340 apiece for noise figure limits of, respectively, 5.5 dB, 5 dB, and 4.5 dB; the 0.5-dB improvement offered by the best grade thus commanded a premium of $140.)

The measurement uncertainties that have been determined for the NBS apparatus and techniques as described are sufficiently large that even if all parties used the same design of apparatus and the same techniques, the interlaboratory variance is likely to give rise to the following difficulty: With diodes ranked in grades with incremental improvements in noise figure limit between grades of 0.5 dB, consider the outputs from two suppliers. Each supplier designates the diode grade as a result of measurement. Because the difference between supplier measurements can be a substantial fraction of the 0.5-dB increment (in fact, at the limit of twice the uncertainty, equal to the increment), two diodes of identical performance are classed in two different grades according to their suppliers. Put another way, if the incremental difference between diode grades is to remain at 0.5 dB and is to be meaningful, independent measurements will not suffice unless or until the basic measurement state of the art is improved to reduce the uncertainty.

(5) The uncertainties established for the attenuation and power calibrations can probably be reduced in the near future. Further improvements in power stability may also be possible. It seems unlikely, however, that the total measurement uncertainty can be reduced much below ± 0.1 dB, which may still be too large for independent producer measurements.

Note: The r-f noise measurements which were to have been investigated later in this program would perhaps have resulted in a somewhat lower measurement uncertainty. This uncertainty, however, would have been at least as large as the uncertainty in calibration of the r-f noise source, which is presently in the range of ± 0.06 to ± 0.08 dB over X-band and Ku-band (and greater in other bands). It is unlikely that a total uncertainty much below ± 0.1 dB can be achieved, even if special care had been used to reduce the calibration uncertainty somewhat below this range.

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555 September 1970, pp. 47-50.
560 November 1970, pp. 42-45.
571 April 1971, pp. 43-45.

592	August 1971, pp. 57-59.
598	October 1971, pp. 36-39.
702	December 1971, pp. 29-30.
717	April 1972, pp. 34-36.
727	June 1972, pp. 51-53.
733	September 1972, pp. 38-40.
743	December 1972, pp. 36-39.
754	March 1973, pp. 29-31.
773	June 1973, pp. 31-33.
788	August 1973, pp. 51-53.
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Quarterly progress reports on this work also appeared in the following NBS Special Publications under the title, Semiconductor Measurement Technology: Microwave Diodes:

400-1	pp. 44-46	March 1974.
400-4	pp. 75-77	November 1974.

APPENDIX A

DERIVATION OF EQUATIONS FOR THE MEASUREMENT OF CONVERSION LOSS USING PERIODIC OR INCREMENTAL MODULATION OF THE LOCAL OSCILLATOR

The wave travelling toward the load (incident wave) in a uniform transmission line of characteristic resistance R_o (however arbitrarily defined), and carrying a constant (unmodulated) power, P_o at a single frequency, ω_o, can be represented by an equivalent instantaneous emf

$$e = E_r\sqrt{2}\,\cos(\omega_o t)\ , \tag{A1}$$

where E_r is the Thevenin's equivalent rms open-circuit emf obtained from the available power expression

$$P_o = \frac{E_r^2}{4R_o}\ . \tag{A2}$$

so that

$$e = 2\sqrt{2P_oR_o}\,\cos(\omega_o t)\ . \tag{A3}$$

Linear amplitude modulation at a single modulating frequency, ω_m, impressed upon this wave results in an instantaneous emf

$$e = E_r\sqrt{2}\,[1 + m\cos(\omega_m t)]\,\cos(\omega_o t) \tag{A4}$$

where m is the modulation factor.

Representing the product of cosines by a sum-and-difference identity:

$$\begin{aligned} e = E_r\sqrt{2}\,\cos(\omega_o t) &+ \frac{mE_r}{\sqrt{2}}\cos\,[(\omega_o + \omega_m)t] \\ &+ \frac{mE_r}{\sqrt{2}}\cos\,[(\omega_o - \omega_m)t]\ . \end{aligned} \tag{A5}$$

This familiar representation shows the two sidebands resulting from the modulation, each of which carries a power $m^2E_r^2/16R_o$, by analogy with eq (A2). The carrier (local-oscillator) power is unchanged by the modulation (for $m \leq 1$).

In the mixer, the intermediate-frequency output voltages resulting from these sidebands add linearly when $m << 1$, so that they are equivalent to a single-frequency r-f signal of, not twice, but four times the power in each sideband:

$$P_s = \frac{1}{4R_o}\left[\frac{mE_r}{2} + \frac{mE_r}{2}\right]^2 = \frac{m^2E_r^2}{4R_o}\ . \tag{A6}$$

Relative to the carrier power, P_o, this equivalent single-frequency r-f signal power, P_s, is, from eqs (A2) and (A6):

$$P_s = m^2 P_o \, . \tag{A7}$$

The equivalent signal is thus 20 $\log_{10}(1/m)$ decibels below the carrier (the local oscillator).

The conversion loss, L_m, of the mixer is defined as the ratio of this equivalent single-frequency available r-f signal power, P_s, to the available intermediate-frequency output power, P_{ia}.

$$L_m \triangleq \frac{P_s}{P_{ia}} \, . \tag{A8}$$

P_{ia} can be expressed as

$$P_{ia} = \frac{E_i^{\,2}}{4(1/G)} \, , \tag{A9}$$

where E_i is the rms open-circuit intermediate-frequency output emf, and G is the intermediate-frequency output conductance. Equation (A8) can thus be written

$$L_m = \frac{4m^2 \, P_o \, (1/G)}{E_i^{\,2}} \, . \tag{A10}$$

In terms of E_i, G, and intermediate-frequency load resistance, R, the intermediate-frequency load voltage, V_i, is given by

$$V_i = E_i \frac{R}{R + \frac{1}{G}} \, . \tag{A11}$$

The intermediate-frequency power delivered to the load, P_{id}, is

$$P_{id} = \frac{V_i^{\,2}}{R} = E_i^{\,2} \frac{R}{\left(R + \frac{1}{G}\right)^2} \, . \tag{A12}$$

Mismatch factor, M, is defined as the ratio of this delivered intermediate-frequency power to the available intermediate-frequency power:

$$M \triangleq \frac{P_{id}}{P_{ia}} \, . \tag{A13}$$

From eqs (A9) and (A12):

$$M = \frac{4R(1/G)}{\left(R + \frac{1}{G}\right)^2} . \tag{A14}$$

From eqs (A8), (A7), (A13), and (A12):

$$L_m = \frac{m^2 P_o RM}{V_i^2} . \tag{A15}$$

In terms of the peak-to-peak intermediate-frequency load voltage, $\Delta V \triangleq 2\sqrt{2}\, V_i$,

$$L_m = \frac{8m^2 P_o RM}{(\Delta V)^2} . \tag{A16}$$

From eqs (A2) and (A4), it can be seen that the ("instantaneous") power over an r-f cycle is approximately

$$P_o [1 + m \cos (\omega_m t)]^2 , \tag{A17}$$

since $\cos(\omega_m t)$ is nearly constant over the r-f cycle when $\omega_o >> \omega_m$. The crest (maximum) value occurs when $\cos (\omega_m t) = 1$, at which time the power is

$$P_c = P_o [1 + m]^2 . \tag{A18}$$

The trough (minimum) value occurs when $\cos (\omega_m t) = -1$, at which time the power is

$$P_t = P_o [1 - m]^2 . \tag{A19}$$

The crest power is thus $20 \log_{10}(1 + m)$ decibels above the carrier, and the trough power is $20 \log_{10}[1/(1 - m)]$ decibels below it.

The total power excursion during modulation may be expressed as the attenuation needed to reduce the power at the crest to that at the trough:

$$L_a \triangleq \frac{P_c}{P_t} = \left(\frac{1 + m}{1 - m}\right)^2 . \tag{A20}$$

In decibels,

$$\begin{aligned} L'_a &\triangleq 10 \log_{10} L_a \\ &= 20 \log_{10} \left(\frac{1 + m}{1 - m}\right) . \end{aligned} \tag{A21}$$

Solving for m,

$$m = \frac{\sqrt{L_a} - 1}{\sqrt{L_a} + 1}$$

$$= \frac{10^{L'_a/20} - 1}{10^{L'/20} + 1} \,. \tag{A22}$$

The total power excursion may also be expressed in absolute terms, using eqs (A18) and (A19):

$$\Delta P \stackrel{\Delta}{=} P_c - P_t$$

$$= P_o\,[1 + m]^2 - P_o\,[1 - m]^2 \tag{A23}$$

$$= 4mP_o \,.$$

or

$$m = \frac{\Delta P}{4P_o}$$

Substituting in eq (A16),

$$L_m = \frac{(\Delta P)^2\ RM}{2\ P_o(\Delta V)^2} \,. \tag{A24}$$

In terms of load current excursion $\Delta I \stackrel{\Delta}{=} \Delta V/R$,

$$L_m = \frac{(\Delta P)^2\ M}{2\ P_o\ R(\Delta I)^2} \,, \tag{A25}$$

which is eq (37) on p. 214 of reference [4], given for the "incremental method," which was not derived on the basis of modulation.

APPENDIX B

DERIVATION OF SENSITIVITY EQUATIONS TO RELATE CONVERSION LOSS UNCERTAINTY TO UNCERTAINTIES IN SYSTEM PARAMETERS

From Appendix A, the most convenient expression for conversion loss [eq (A16)] is

$$L_m = \frac{8m^2 P_o RM}{(\Delta V)^2} . \tag{B1}$$

The sensitivity of L_m to each of the various quantities could be obtained by taking the partial derivative of L_m with respect to each of these. This would result, however, in "conditional" sensitivities, i.e., each partial would contain the other quantities as constants, whose values would require selection. Since conversion loss is generally expressed in decibels,

$$L'_m \stackrel{\Delta}{=} 10 \log_{10} L_m , \tag{B2}$$

eq (B1) can be expressed as a sum of terms (the logarithms of the quantities), by writing it in decibel form.

In the case of modulation factor m, since it is not known directly, it is desirable to express the sensitivity of conversion loss to the attenuation used to measure or produce m. Substituting from eq (A22) into eq (B1),

$$L_m = \frac{8P_o RM}{(\Delta V)^2} \left[\frac{10^{L'_a/20} - 1}{10^{L'/20} + 1}\right]^2 . \tag{B3}$$

Expressed in decibels [eq (B2)],

$$L'_m = 20 \log_{10} \left[\frac{10^{L'_a/20} - 1}{10^{L'_a/20} + 1}\right] + 10 \log_{10} \left[\frac{8P_o RM}{(\Delta V)^2}\right] . \tag{B4}$$

Taking the partial derivative of L'_m with respect to L'_a, the second term, containing all of the "other" quantities, which are held constant for this derivative, disappears:

$$\frac{\partial L'_m}{\partial L'_a} = \frac{2}{10^{L'_a/20} - 10^{-L'_a/20}} . \tag{B5}$$

The exponents in the denominator of this derivative may be expanded in series form:

$$10^x = 1 + x \ln 10 + \frac{(x \ln 10)^2}{2!} + \frac{(x \ln 10)^3}{3!} + \ldots \tag{B6a}$$

$$10^{-x} = 1 - x \ln 10 + \frac{(x \ln 10)^2}{2!} - \frac{(x \ln 10)^3}{3!} + \ldots \tag{B6b}$$

Therefore

$$10^{x} - 10^{-x} = 2x \ln 10 + 2\,\frac{(x \ln 10)^3}{3!} + \ldots \tag{B7}$$

For $x \ll 1$ ($L'_a \ll 20$ dB), this series is approximated by its first term, so that

$$\frac{\partial L'_m}{\partial L'_a} \simeq \frac{20}{L'_a \ln 10} = \frac{8.69}{L'_a}\ \text{dB/dB}\ , \tag{B8}$$

which may also be written as

$$\frac{\partial L'_m}{\partial L'_a} \simeq \frac{20 \log_{10} e}{L'_a} = \frac{8.69}{L'_a}\ \text{dB/dB}\ . \tag{B9}$$

The partials with respect to the other quantities in eqs (B1 or (B3) are obtained directly:

$$L'_m = 10 \log_{10} P_o + 10 \log_{10} \left[\frac{8m^2\ RM}{(\Delta V)^2}\right] \tag{B10}$$

$$\frac{\partial L'_m}{\partial P} = \frac{10 \log_{10} e}{P_o} = \frac{10}{P_o \ln 10} = \frac{4.34}{P_o}\ \text{dB/W}\ .$$

The derivatives with respect to R and to M are of identical form (the latter is not used).

For ΔV:

$$L'_m = - 20 \log_{10} (\Delta V) + 10 \log_{10} (8\ m^2\ P_o\ RM) \tag{B11}$$

$$\frac{\partial L'_m}{\partial(\Delta V)} = - \frac{20 \log_{10} e}{\Delta V} = \frac{-20}{(\Delta V) \ln 10} = - \frac{8.69}{\Delta V}\ \text{dB/V}\ . \tag{B12}$$

APPENDIX C

REDUCTION OF DATA CONTAMINATION BY THE USE OF DATA GROUP MEDIANS

Data may sometimes be "contaminated" by the random occurrence during measurements of error mechanisms not affecting most of the data, such as the misreading of a scale or the disturbance by a large power-line transient. The resulting "contaminants" frequently affect the smoothness of data plots by appearing as grossly atypical data commonly known as "outliers." Occasional outliers may be discarded by somewhat laborious methods [C1,C2] which consider the deviation of each relative to the standard deviation of the other data.

A simpler method under certain circumstances is to repeat the measurement several times for the same nominal conditions (same independent variables) and then discard all data from this group except the median. An odd number of repetitions simplifies the process.

Since a contaminant falling amid valid data would not be unrepresentative of the latter, it can be accepted as "valid" by definition. The word "contaminant" will therefore be reserved for scalar data taken during the operation of an unusual error mechanism (or one whose effects we wish to disregard) which lie above or below the data taken when no such mechanism is operative.

If the contaminants must lie outside the valid data, then the median itself will not be a contaminant unless, in a group of n data, there are at least $\frac{n+1}{2}$ contaminants of the same sense, i.e., either all higher than the valid data or all lower. The probability that this will happen has been calculated for groups of three, five, and seven data, and is significantly less than the probability of a (raw) datum being a contaminant for even moderately small values of the latter probability.

The calculations assume that the occurrence of each contaminant is a random event of probability P independent of the occurrence of other contaminants, and that the contaminant is equally likely to lie above or below the valid data. (For contaminants that lie on only one side, or which favor one side, somewhat different calculations are required.) A datum thus has a probability P/2 of being a contaminant of a given sense, and a probability (1 - P) of not being a contaminant. In a group of n data, the probability that a given particular sequence containing m contaminants of specified senses will occur is thus $(P/2)^m(1 - P)^{n-m}$ The number of such sequences having contaminant medians is determined for each value of m, as shown in Table C1. As a check, the sequences whose medians are not contaminants are also enumerated. Since contaminants of different senses are distinguished, there are three possible labels for a given datum, and therefore 3^n distinguishable sequences for a group of n data. As a further check, the probability of the median being a con-

Table C1. Sequence Count for Probability Calculations

Number of contaminants in data group	Number of ways in which the median is (is not) a contaminant obtainable in a group of 1 Datum		3 Data		5 Data		7 Data	
0	0	(1)	0	(1)	0	(1)	0	(1)
1	2	(0)	0	(6)	0	(10)	0	(14)
2	-		6	(6)	0	(40)	0	(84)
3	-		8	(0)	20	(60)	0	(280)
4	-		-		50	(30)	70	(490)
5	-		-		32	(0)	252	(420)
6	-		-		-		308	(140)
7	-		-		-		128	(0)
Total Ways:	3		27		243		2187	
(check)	$=3^1$		$=3^3$		$=3^5$		$=3^7$	

taminant and the probability of its not being a contaminant have been separately calculated and shown to add to unity.*

As shown in Table C2, the probability of the median being a contaminant may be expressed as the product of (1) a power of P and (2) a polynomial of P. For small values of P (infrequent contaminants), the value of the polynomial is determined largely by its lowest order (constant) term. General forms for the polynomial and for its first term have been sought but without success.

Of greater interest than the absolute probabilities are the ratios of these probabilities to the degenerate single-datum-group contaminant probability P, the case where no data are discarded. Normalizing the probabilities by P results in what is termed here the "Contaminant Reduction Factor" (CRF). The probability of a data group median being a contaminant is thus equal to the product of the CRF and the probability that any (raw) datum is a contaminant. Expression for the CRF for groups of seven data or less are given in Table C3 and plotted in figure C1. It should be particularly noted that the exponent $\frac{n-1}{2}$ which largely determines the value of the CRF is equal to the number of data pairs flanking the median. Every additional data pair reduces the contaminant probability by a factor of about P, since there is little change in the polynomial as n is varied, for small values of P.

Figure C1 suggests an analogy between this method of contaminant reduction and a low-pass electrical filter. Adding data pairs is analogous to adding filter sections, the effect being to increase the steepness of the curve of CRF *vs.* relative contaminant frequency (P), i.e., to "sharpen the response." For the degenerate single-datum-group case, the "response" is flat, analogous to having no filter.

Just as several low-pass electrical filters may be used in cascade, data group medians may be used in conjunction with other methods of data smoothing. For example, several such medians obtained under nominally identical conditions may be averaged, or several data closest to the median in a large group could be retained and averaged, discarding only a few pairs near the high and low extremes.

The price to be paid for the use of data group medians to reduce data contamination is, of course, the necessity to greatly increase the amount of (raw) data. To again use the electrical filter analogy, there is an appreciable "insertion loss." In some situations, it may be possible to repeat only a part of the measurement to generate a data group, where (e.g.) there are applied conditions that drift slowly enough not to require readjustment between every measurement (and which do not effect the contaminant incidence). It may be argued that the same amount of raw data could be put to better use, using other smoothing techniques. For

---

* These probabilities are independent of contaminant magnitude, which would not be the case if the definition of contaminants allowed them to fall amid the valid data.

Table C2. Probability Expressions.

Number of data in group	Probability that group median is a contaminant
1	$P = (\frac{P}{2})\ (2)$
3	$P^2(1.5 - 0.5\ P) = (\frac{P}{2})^2\ (6 - 2\ P)$
5	$P^3(2.5 - 1.875\ P + 0.375\ P^2) = (\frac{P}{2})^3\ (20 - 15\ P + 3\ P^2)$
7	$P^4(4.375 - 5.25\ P + 2.1875\ P^3 - 0.3125\ P^3)$
	$= (\frac{P}{2})^4\ (70 - 84\ P + 35\ P^2 - 5\ P^3)$
n	$P^{\frac{n+1}{2}}$ (polynomial of order $\frac{n-1}{2}$)

where P is the probability that a datum will be a contaminant. No general form for the polynomial has yet been found.

Table C3. Contaminant Reduction Factors.

Number of data in group	Contaminant Reduction Factor (CRF)
1	1
3	$P\ (1.5 - 0.5\ P) = (\frac{P}{2})\ (3 - P)$
5	$P^2(2.5 - 1.875\ P + 0.375\ P^2) = (\frac{P}{2})^2\ (10 - 7.5\ P + 1.5\ P^2)$
7	$P^3(4.375 - 5.25\ P + 2.1875\ P^2 - 0.3125\ P^3)$
	$= (\frac{P}{2})^3\ (35 - 42\ P + 17.5\ P^2 - 2.5\ P^3)$
n	$P^{\frac{n-1}{2}}$ (polynomial of order $\frac{n-1}{2}$)

where P is the probability that a datum will be a contaminant.

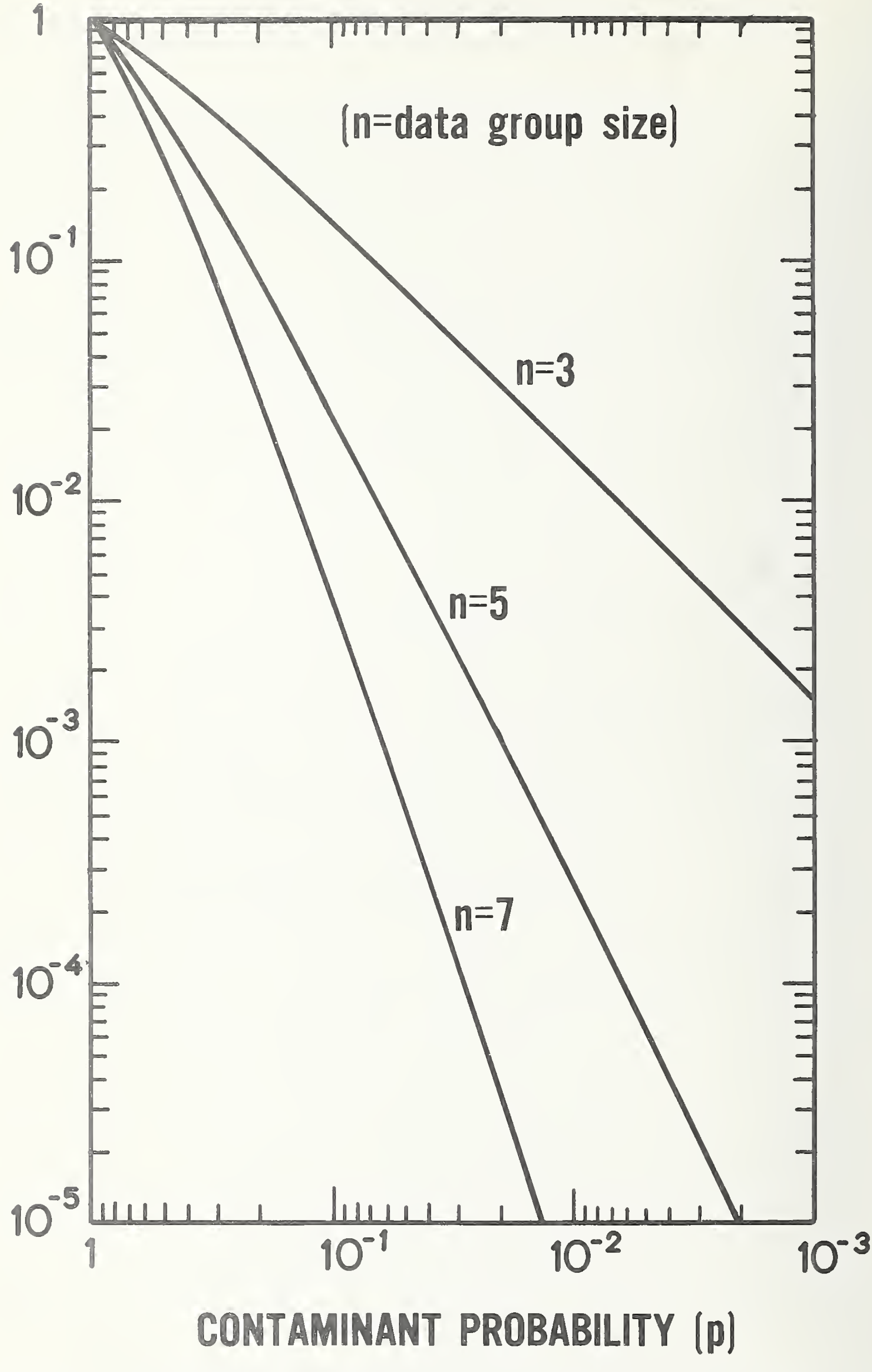

Figure C1. Contaminant reduction factors for groups of seven data or less.

small data groups, however, it is easy to select the medians rapidly, greatly reducing the data to be processed further. For these reasons, the efficiency of the data group median method in terms of relative speed of processed data production may be much better than the ratio of medians to total data (1/n).

The CRF calculations are also applicable to the improvement of system reliability by the use of active circuit redundancy. The outputs of several identical circuits (whose inputs are tied together) can be ranked electronically and only the median output used. If P is the failure rate for one of the circuits, then the overall failure rate would be reduced to CRF times P if the failure rate of the median selection circuit was negligible by comparison.

References:

C1. Natrella, M. G., *Experimental Statistics, NBS Handbook* 91, Washington: GPO, 1963, Chapter 17.

C2. Ku, H. H., (ed.), *Precision Measurement and Calibration,* NBS Special Publication 400, Washington: GPO, 1969, pp. 346-354.

APPENDIX D

DEFINITIONS OF MIXER TERMS

Conditions, standard mixer measurement: specified values for system variables outside the mixer (or diode) of which the measured values are functions. When the measurement conditions are standard, the measured values may be attributed to the mixer (or diode). When one or more conditions are not standard during the measurement, a correction may be made for the difference or the resulting change in the measured value covered by an additional uncertainty allowance. The measurement conditions to be specified include some or all of the following: Local-oscillator and intermediate frequencies; terminating immittances at the local-oscillator frequency and at all input frequencies for which there is a significant output response (e.g., nonreflective at all frequencies); local-oscillator power available to the mixer; intermediate-frequency and d-c loads and external bias, if any; physical temperature of diode or holder; and, for diode characterization, the physical construction of the diode holder and, if tunable, the tuning procedure to be followed. Signal levels are to be sufficiently small that the measured parameters are independent of signal strength. Other conditions to be standardized are noted under individual parameter definitions.

Conductance, intermediate-frequency output, G: the real part of the mixer intermediate-frequency output admittance. Standard mixer measurement conditions apply. Intermediate-frequency output conductance is commonly expressed in units of ohms (of its reciprocal). The traditional but erroneous use of the term "intermediate-frequency impedance" for this quantity is deprecated.

Conversion Loss, L: the ratio of available r-f power at a single designated signal frequency to that part of the available (or delivered) intermediate-frequency power which is a function of the r-f power (excluding that part which is independent of input). Standard mixer measurement conditions apply. When delivered intermediate-frequency power (to a specified load) is used, the loss is referred to as "conversion insertion loss."

Noise figure (noise factor), average, $\overline{F}$: for a multiport with specified terminating immittances, the ratio of (1) the total noise power available from its designated output port (or delivered to a specified load) within a designated output frequency band, when the noise temperature of all terminations is a standard reference temperature, to (2) that portion of (1) which is a function of the noise temperature of the designated signal input termination within a designated signal input frequency band. Average noise figure is a dimensionless power ratio, commonly expressed in decibels. In measuring mixers and mixer diodes, there are two average noise figures of common interest: (1) intermediate-frequency average noise figure, and (2) overall average noise figure:

Intermediate-frequency average noise figure, $\overline{F}_i$: the average noise figure of an intermediate-frequency amplifier. Its general dependency upon intermediate-frequency output conductance is to be particularly noted.

Overall average noise figure, $\overline{F}_o$: the average noise figure of the cascaded combination of a mixer and high gain intermediate-frequency amplifier.

When used to characterize a mixer, standard mixer measurement conditions apply, with the following additional qualifications: The intermediate-frequency average noise figure is standard (usually $\geq$1.5 dB), and the passband of the intermediate-frequency amplifier is sufficiently narrower than that of the mixer so that the mixer conversion loss and output noise temperature are constant over the intermediate-frequency passband. When these conditions are met, the prefix "standard" may be used to distinguish both the intermediate-frequency and overall average noise figures, and the subscript "s" added to the symbols, i.e.,

$\overline{F}_{is}$ = standard intermediate-frequency average noise figure,

and

$\overline{F}_{os}$ = standard overall average noise figure.

When the overall average noise figure is measured with a nonstandard (but known) intermediate-frequency average noise figure, the standard overall noise figure may be calculated from:

$$\overline{F}_{os} = \overline{F}_o + L\,(\overline{F}_{is} - \overline{F}_i)$$

or

$$\overline{F}_{os} = F_o \frac{N + \overline{F}_{is} - 1}{N + \overline{F}_i - 1}$$

where L = conversion loss
N = output noise ratio.

When the intermediate-frequency average noise figure is not known, but has a known lower bound, $\overline{F}_{ib}$, the following inequalities may be useful for qualification testing:

$$\overline{F}_{os} < \overline{F}_o + L\,(\overline{F}_{is} - \overline{F}_{ib})$$

or

$$\bar{F}_{os} < F_o \frac{N + \bar{F}_{is} - 1}{N + \bar{F}_{ib} - 1} .$$

To avoid confusion with standard values, measured (general) values of intermediate-frequency and overall average noise figure may be distinguished by the subscript "m," i.e.,

$\bar{F}_{im} = \bar{F}_i$ = (measured) intermediate-frequency average noise figure,

or

$\bar{F}_{om} = \bar{F}_o$ = (measured) overall average noise figure.

Noise ratio, output, N: the ratio of the noise temperature of the mixer intermediate-frequency port to a standard reference temperature when the noise temperature of all terminations is a standard reference temperature. Standard mixer measurement conditions apply. Output noise ratio is a dimensionless power ratio.

Temperature, noise, T: the uniform physical absolute temperature (kelvins) at which a network (and all its sources, if a multiport) would have to be maintained if it were passive in order to make available the same random power spectral density (watts/hertz) at a given frequency as it actually makes available, its output immittance being unchanged. For microwave, and lower, frequencies, noise temperature may be considered as equal to the power spectral density divided by Boltzmann's constant (1.38062×10^{-23} J/K).

Temperature, standard reference, T_o: a specified absolute temperature (kelvins) to be assumed as the noise temperature at the input ports of a network for the purpose of calculating certain noise parameters, and for normalizing purposes. The IEEE has established the value of T_o as 290 K, but a higher temperature may be preferred for mixer measurements since the presence of spurious responses of unknown magnitude makes impossible an exact correction for an ambient temperature different from T_o. The value of 300 K is tentatively recommended for mixer measurements, but other values may be used provided that consistency is observed for a given type of device and that the value used is clearly stated. Note: this is not a toleranced quantity.

APPENDIX E

OUTLINE OF MIXER NOISE MEASUREMENTS

STANDARD OVERALL AVERAGE NOISE FIGURE (FACTOR)

The standard overall average noise figure (factor) of a mixer is the average noise figure of the cascaded combination of the mixer and an intermediate-frequency amplifier when the average noise figure of the latter is a specified standard value (usually 1.5 dB) (see Appendix D).

The measurement is performed by increasing the noise temperature of the mixer r-f termination from T_l to T_h and noting the resultant relative (decibel) increase in intermediate-frequency output power. The r-f noise temperature may be changed by turning on and off the noise source itself (as is possible with a gas-discharge type), by switching between a source of relatively high noise temperature and one of relatively lower temperature, either one of which may be at ambient temperature (as is required for physically hot or cold terminations), or by inserting attenuation between a noise source of any type and the mixer, in all cases being careful not to alter the mixer r-f source immittance. The relative change in intermediate-frequency output power is commonly measured by using a calibrated intermediate-frequency attenuator following the preamplifier to maintain the output indicator reading constant. It is necessary to operate the intermediate-frequency preamplifier at high gain, but sufficiently below saturation (overload) to avoid nonlinearity due to significant clipping of noise peaks. Using the assumptions given below, the average noise figure may be obtained as:

$$\overline{F}_o = \frac{\left(\frac{T_h}{T_o} - 1\right) + L_o\left(1 - \frac{T_l}{T_o}\right)}{L_o - 1}\left(1 + \frac{L_{ms}}{L_{mi}}\right) - \left(\frac{T_a}{T_o} - 1\right)\sum_j \frac{L_{ms}}{L_{mj}}$$

$$= \left[\frac{T_h - T_l}{T_o(L_o - 1)} + \left(1 - \frac{T_l}{T_o}\right)\right]\left(1 + \frac{L_{ms}}{L_{mi}}\right) - \left(\frac{T_a}{T_o} - 1\right)\sum_j \frac{L_{ms}}{L_{mj}}, \qquad \text{(E1)}$$

where:

$\overline{F}_o$ = overall average noise figure (factor) (power ratio),

T_h = the higher mixer r-f source temperature (K),

T_l = the lower mixer r-f source temperature (K),

T_o = reference noise temperature (K),

T_a = ambient temperature (K),

L_o = increase in noise power output from intermediate-frequency preamplifier (= intermediate-frequency attenuation required to maintain constant postamplifier [main amplifier] output) as mixer r-f source temperature is changed from T_l to T_h (power ratio),

L_{ms} = mixer conversion loss from signal frequency (power ratio),

L_{mi} = mixer conversion loss from image frequency (power ratio), and

L_{mj} = mixer conversion loss from the jth r-f frequency other than signal or image frequency (power ratio) (summation is over all significant frequencies).

This equation assumes that:

1) the intermediate-frequency amplifier has a sufficiently narrow passband such that the mixer loss and output noise are effectively constant over this band;

2) the r-f source temperature is always the same at signal and image frequencies, and is equal to the ambient temperature at all other frequencies making significant contributions to the mixer intermediate-frequency output noise. (Note: A lowpass filter following the r-f noise source may be required if the mixer harmonic rejection is inadequate or unknown.);

3) the mixer and intermediate-frequency preamplifier are linear; and

4) the gains and noise contributions of the mixer and intermediate-frequency preamplifier are unchanged by the change in input noise.

Three slightly different operating procedures may be cited:

1. The r-f noise temperature is switched between two fixed values, and a variable intermediate-frequency attenuator or wide-range intermediate-frequency power meter is used to measure the relative (decibel) change in output power.

2. The r-f noise temperature is varied by means of an r-f attenuator between ambient (using a very large attenuation) and the value (using attenuation L_r) required to change the intermediate-frequency output power by a predetermined ratio, L_o. Using a "hot" (above ambient) noise source, the general equation is modified by the substitutions $T_h = T_a + (T_h' - T_a)/L_r$ and $T_l = T_a$ where T_h' is the mixer source noise temperature with the attenuator set to zero, i.e., the noise source temperature corrected for the insertion loss of the attenuator (set to zero) and other r-f components. Using a "cold" (below ambient) noise

source, the substitutions are $T_l = T_a - (T_a - T_l')/L_r$ and $T_h = T_a$ where T_1' is the mixer source noise temperature with the attenuator set to zero. L_o is most conveniently and accurately set by means of a switchable intermediate-frequency attenuator, the intermediate-frequency output indication being kept constant.

3. The r-f noise temperature is varied by means of an r-f attenuator between a maximum calibrated value (using zero attenuation) and the value (using attenuation L_r) required to change the intermediate-frequency output power by a predetermined ratio, L_o. The general equation is modified by the substitution $T_l = T_a + (T_h - T_a)/L_r$ if a "hot" noise source is used, or $T_h = T_a - (T_a - T_l)/L_r$ if a "cold" noise source is used. The noise source temperature must be calibrated at the mixer r-f port or corrected for the insertion loss of the attenuator (set to zero) and other r-f components.

If the average noise figure of the intermediate-frequency amplifier does not differ significantly from the specified standard value (e.g., 1.5 dB) when the overall average noise figure is measured, then the measured value is the *standard* overall average noise figure and may be ascribed to the mixer diode. If the intermediate-frequency average noise figure is not standard, then the following equations may be used to obtain the standard overall average noise figure, $\overline{F}_{os}$, from the measured overall average noise figure, $\overline{F}_{om}$; either the conversion loss, L_m, or the output noise ratio, N_r; and the standard and actual (measured) values of intermediate-frequency average noise figure, $\overline{F}_{is}$ and $\overline{F}_{im}$, respectively:

$$\overline{F}_{os} = \overline{F}_{om} + L_m (\overline{F}_{is} - \overline{F}_{im}) \tag{E2}$$

$$\overline{F}_{os} = \overline{F}_{om} \frac{N_r + \overline{F}_{is} - 1}{N_r + \overline{F}_{im} - 1} . \tag{E3}$$

If the actual intermediate-frequency noise figure is not known, but has a known lower bound, $\overline{F}_{ib}$, then the following inequalities may be useful for acceptance testing:

$$\overline{F}_{os} < \overline{F}_{om} + L_m (\overline{F}_{is} - \overline{F}_{ib}) \tag{E4}$$

$$\overline{F}_{os} < \overline{F}_{om} \frac{N_r + \overline{F}_{is} - 1}{N_r + \overline{F}_{ib} - 1} . \tag{E5}$$

The actual average noise figure of the intermediate-frequency amplifier may be determined from a plot of intermediate-frequency average noise figure *vs.* mixer intermediate-frequency output conductance, for the particular conductance of the diode under test. This plot is based on measurements made by substituting resistors for the diode or mixer, with the conductance varied but with the susceptance maintained at very nearly a constant value equal to that obtained with normal diode operation. The intermediate-frequency average noise figure for each value of conductance is calculated from a measurement of the average anode current of a temperature-limited thermionic diode (noise diode) connected across the mixer intermediate-frequency output port with the diode filament temperature adjusted to add sufficient shot noise to increase the intermediate-frequency preamplifier output by an arbitrary factor, L_o (e.g., L_o = 2), as determined by an intermediate-frequency attenuator, the output of which is kept constant. The intermediate-frequency average noise figure, $\overline{F}_i$, is then given by:

$$\overline{F}_i = 1 + \frac{eI}{2kT_o G\ (L_o - 1)} - \frac{T_a}{T_o}, \tag{E6}$$

where:

- e = elementary (electronic) charge (1.602×10^{-19} C);
- k = Boltzmann constant (1.381×10^{-23} J/K);
- I = average anode current (d-c component) of noise diode (A);
- T_o = reference noise temperature (K);
- T_a = ambient temperature = resistor temperature (K); and
- G = mixer intermediate-frequency output conductance (usually expressed in ohms of its reciprocal).

A coupling circuit equivalent to a 1/8-wavelength transmission line may be used at the input to the intermediate-frequency amplifier to reduce the dependence of the amplifier average noise figure on mixer intermediate-frequency output conductance over the anticipated conductance range. The conductance measurements for the diodes and resistors should either be made at the intermediate frequency used for the noise measurements, or resistor types selected that will provide very nearly identical conductance values at the frequency of measurement.

OUTPUT NOISE RATIO

To determine the output noise ratio of a mixer, the measurement must distinguish between the mixer output noise originating from within the mixer (and also converted from the T_o component of r.f. source noise) and the noise originating from other sources: (1) the excess (T_a - T_o) component of r-f source noise; (2) local-oscillator noise; and (3) the intermediate-frequency amplifier noise (which is a function of the mixer output admittance), represented by the amplifier average noise figure, $\overline{F}_i$. The excess r-f source noise can be accounted for in the noise equations from the r-f source (ambient) temperature, T_a, and standard noise temperature, T_o. Note, however, that the effect of ambient temperature on an active source such as a mixer cannot generally be accounted for theoretically. Local-oscillator noise can be removed by r-f filtering. The most troublesome problem in output noise ratio measurements has been to separate the mixer noise from the intermediate-frequency amplifier noise. Previous approaches to this problem have required either (1) a resistor to replace the diode (or a resistor-capacitor combination to replace the entire mixer), where the resistor conductance is the same as the mixer output conductance for each diode to be tested, or (2) a 1/8-wavelength coupling circuit to sufficiently reduce the dependency of amplifier noise upon mixer output conductance. Both approaches require good amplifier gain stability. To avoid these requirements, and to provide greatly increased measurement accuracy, the following procedure was originated, and successfully used, by the author at the Naval Applied Science Laboratory.

The measurement requires a temperature-limited thermionic noise diode (such as JAN-5722) connected directly across (and thereby defining) the junction between the mixer and the amplifier, as previously described for the measurement of $\overline{F}_i$ (circuit layout is critical). For accuracy and convenience, a calibrated (and padded) fixed attenuator of about 3.01 dB (L_o = 2) that may be easily switched in or out between the preamplifier and postamplifier (main amplifier) should be used. A further refinement is the addition to the attenuator of a switch to control the anode voltage of the noise diode, isolated electrically from the attenuator but mechanically connected so as to apply the anode voltage simultaneously as the attenuation is inserted. For this arrangement, the operating procedure is as follows:

1. Replace the mixer diode with a resistor in a diode package or replace the entire mixer with a shielded resistor-capacitor combination, keeping the susceptance the same as with the normal configuration.

2. With the attenuator out of the circuit (and the noise diode off), adjust the amplifier gain to provide a convenient and precise second-detector (d-c) output reading (reference value).

3. Insert the attenuator (applying the noise-diode anode voltage) and adjust the noise-diode filament temperature to restore the second-detector reference reading.

4. Note the average anode current, I, of the noise diode.

5. Check on system drift as follows: Switch the attenuator out of the circuit (turning off the noise diode). Check to see that the second detector reference reading has not drifted. If required, repeat steps 2 through 5.

6. Repeat steps 1 through 5 for a number of different simulated mixer conductance values covering the conductance range obtained for normal mixer operation using the diodes under test.

7. Plot either I or $\overline{F}_i$ [calculated from eq (E6)] as a function of G (or 1/G).

8. Insert a diode to be tested in the holder (normal mixer operation) and repeat the measurement procedure (steps 2 through 5) used with the resistors.

9. From a measurement of the mixer output conductance with this diode, use the plot obtained in step 7 to obtain a corresponding value of I or $\overline{F}_i$. Distinguish the average anode current of the noise diode obtained in step 8 from the current obtained from the plot for a corresponding conductance value by using the symbol I_d for the former and I_g for the latter.

6. Calculate output noise ratio, N_r, using the I_g *vs.* G plot as

$$N_r = 1 + \frac{e(I_d - I_g)}{2kT_o G\ (L_o - 1)} + \frac{T_a}{T_o} \tag{E7}$$

or from the $\overline{F}_i$ *vs.* G plot as

$$N_r = 1 + \frac{eI_d}{2kT_o G(L_o - 1)} - \overline{F}_i\ . \tag{E8}$$

Note that the first equation (E7) holds when L_o is identical for all relevant insertions of the attenuator, as should be the case for a switchable fixed attenuator with good switch contacts.

Fine control and stability of noise diode filament temperature is greatly facilitated by a filament supply which is servo-controlled by the anode current.

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☐ Document describes a computer program; SF-185, FIPS Software Summary, is attached.

16. ABSTRACT *(A 200-word or less factual summary of most significant information. If document includes a significant bibliography or literature survey, mention it here.)*

The measurement of mixer conversion loss using periodic or incremental modulation of the local oscillator, and the evaluation and minimization of the associated systematic and random uncertainties, are discussed in terms of an X-band mixer measurement system constructed at NBS. It is shown that the systematic uncertainty in the incremental modulation method of measuring conversion loss results largely from the uncertainties in the calibration of microwave attenuation and power.

It is also shown that the "modulation" (periodic modulation) and "incremental" (incremental modulation) methods of measuring conversion loss are essentially identical, the only practical distinction being in the somewhat different instrumentation required by the different modulation rates.

Several improvements in the periodic and incremental modulation techniques are introduced. Novel circuits for measuring intermediate-frequency output conductance and local-oscillator return loss are described which may also be useful for other immittance measurements.

17. KEY WORDS *(six to twelve entries; alphabetical order; capitalize only the first letter of the first key word unless a proper name; separated by semicolons)* Conversion loss; diode measurement; intermediate-frequency (i-f) output conductance; measurement uncertainties; microwave mixer diodes; modulation; point-contact diodes; reflectometer; return loss; Schottky-barrier diodes; semiconductor diodes; standing-wave ratio (SWR).

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