nonreflecting waveguides, preventing mismatch errors.

Standards of phase shift and impedance may be made to high precision by careful machining techniques. With refined instrumentation, the system should prove capable of impedance measurements of the highest accuracy and therefore useful in calibration work.

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On the Noise Temperature of Coupling Networks*

When a passive coupling network, such as a waveguide, transmission line, matching filter, etc., is used to connect a source to a receiver, it is apparent that it will contribute noise to the output because of its lossiness. If the noise temperature of the source is T_s and the temperature of the coupling network is T_n, then the noise temperature, T_o, at the output (under matched conditions) is given by[1]

$$T_o = \frac{T_s}{L} + T_n\left(1 - \frac{1}{L}\right), \qquad (1)$$

where L is the coupling-network power loss ratio. This relationship was derived by constructing a transmission line analog to the coupling network and treating the source and loss noises as propagating signals. An alternative derivation based on a more physical representation is presented in this note.

Consider the coupling network as a generalized two-port with matched input and output. Its noise power output, P, can be written

$$P = \frac{kT_sB}{L} + kT_nBf, \qquad (2)$$

where the first term is simply the attenuated source noise power and the other is some fraction, f, of the noise power available from the coupling network. Since (2) is true for all values of the parameters, it is true, in particular, when the coupling network is at the same temperature as the source, yielding

$$P_{T_n=T_s} = kT_sB\left(\frac{1}{L} + f\right). \qquad (3)$$

However, the noise contributions to the output from the source and from the coupling network become indistinguishable when both are at the same temperature. That is, the output from the coupling network then looks exactly like that from the source itself, so

$$P_{T_n=T_s} = kT_sB, \qquad (4)$$

and combining (3) and (4) yields

$$f = 1 - \frac{1}{L}\cdot \qquad (5)$$

Writing the general noise power output, P, as kT_0B then gives

$$T_0 = \frac{T_s}{L} + T_n\left(1 - \frac{1}{L}\right), \qquad (6)$$

Q.E.D.

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[1] P. D. Strum, "A note on noise temperature," IRE TRANS. ON MIROWAVE THEORY AND TECHNIQUES, vol. MTT-4, pp. 145–151; July, 1956.

A Logarithmic Transmission Line Chart*

In his article above,[1] Hudson raises the question: "What length of line of what impedance will match a given impedance?" He states, "Conventional charts do not answer this question explicitly."

This problem can be solved on the "conventional" Smith Chart (Fig. 1) explicitly without trial-and-error, by the following method. If A and B are two quite general impedances, the matched condition requires that A be transformed into B^*, the complex conjugate of B.

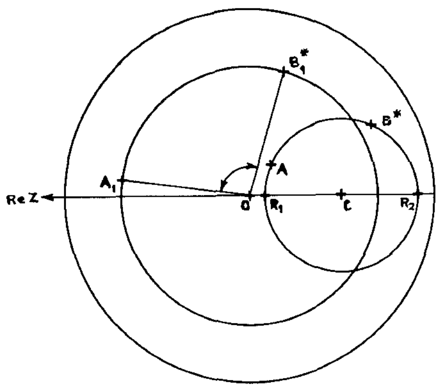

Fig. 1—Smith Chart.

1) Plot A and B^* on the Smith Chart and draw the circle through these two points which has its center (C) on the real axis. If this circle lies fully within the Smith Chart, the question has a solution, otherwise not.
2) Read off the values at the intersections of the real axis and the circle (R_1, R_2), and determine their geometric mean $\sqrt{R_1 \times R_1}$, which will be the characteristic impedance of the matching line.

To find the length of this line,

3) On the Smith Chart normalized to $\sqrt{R_1 \times R_2}$, represent A and B^* by A_1 and B_1^*.
4) The electrical length of the matching line will be given by half of the angle $A_1OB_1^*$.

This method is based upon the following two properties of loss-free transmission lines:

1) The locus of the impedance along a line is always a circle on the Smith Chart (having its centre on the real axis) irrespective of the value of the normalizing resistance.
2) The characteristic impedance of the line is given by $Z_0 = \sqrt{(\text{Re } Z)_{\max} \times (\text{Re } Z)_{\min}}$, where Z is the impedance along the line.

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* Received by the PGMTT, April 1, 1960.

[1] A. C. Hudson, IRE TRANS. ON MICROWAVE THEORY AND TECHNIQUES, vol. MTT-7, pp. 277–281; April, 1959.

Velocity Sorting Detection in Backward Wave Autodyne Reception*

An electronically tunable microwave receiver which uses an oscillating backward wave amplifier driving a crystal detector has been described previously.[1] This receiver has the advantages of large dynamic range and good rejection of unwanted signals, but has the disadvantage that its frequency response can be no better than that of its crystal detector. Since a variation in sensitivity of greater than 3 db over the tuning range of 8 to 12 kmc would seriously lower its usefulness as a spectrum display device, the restrictions on the crystal detector performance are quite severe.

In the paper describing the operation of the device, the author made the suggestion that it might be possible to detect the video output by means of a suitable collector. This letter describes the results of an experimental velocity sorting detector used with the backward wave autodyne receiver.

The first tube used was a Varian VAD-161-2. The collector in this tube was not designed for depressed operation, and, as a result, when the collector voltage was lowered to within a few volts of the cathode potential a virtual cathode was formed near the collector.

A three-dimensional plot of collector current vs collector voltage and beam current is presented in Fig. 1. The current is a multivalued function of collector potential which resulted in the production of oscillations when a load resistance was connected to the collector. Because the oscillations occurred at the setting of collector voltage

* Received by the PGMTT, December 28, 1959; revised manuscript received, April 4, 1960.

[1] J. K. Pulfer, "Application of a backward wave amplifier to microwave autodyne reception," IRE TRANS. ON MICROWAVE THEORY AND TECHNIQUES, vol. MTT-7, pp. 356–359; July, 1959.

which produced maximum sensitivity, detection in this tube was not satisfactory.

A new tube designated type VAD-161-4 was developed which had a collector redesigned for depressed operation. With this tube no instability was observed at any value of collector voltage, beam current or beam velocity.

Fig. 2 is a three-dimensional model of collector current for the VAD-161-4. The collector current below the oscillation starting current increases with both beam current and collector voltage, indicating that no virtual cathode exists.

At the start of oscillations (a beam current of about 3.5 ma), there is an abrupt change in collector current. At a low collector voltage the collector current rises at the oscillation starting point, but at a higher collector voltage the beam current drops steeply when oscillations begin. This steep drop may be used to detect small changes in beam velocity distribution.

Fig. 3 is a plot of detected output signal level for a constant input vs collector voltage. Beam current was optimized for maximum output.

An important characteristic of the autodyne detector is the rate at which the output signal is reduced as the beam current is increased beyond starting current. In Fig. 4, normalized output from the two types of detector is plotted against normalized beam current. One hundred on the abscissa represents starting current. The crystal detector is much less sensitive to variations in beam current. Note that in contrast to the crystal detector the collector detection efficiency falls off almost as rapidly above starting current as it does below.

The mechanism of detection by a negatively biased collector is such that it is the beat frequency between the oscillation and the signal which is actually detected. When either of these becomes large relative to the other, the change in collector current due to the beat becomes small and detection efficiency drops off.

When the beam current is increased above starting current, the amplitude of the oscillation increases extremely rapidly, so that detection efficiency decreases rapidly for beam currents above starting current.

Because of this effect, a velocity sorting detector has not proved superior to a crystal detector in the electronically tunable receiver application, although sensitivity equivalent to a good crystal detector can be easily obtained.

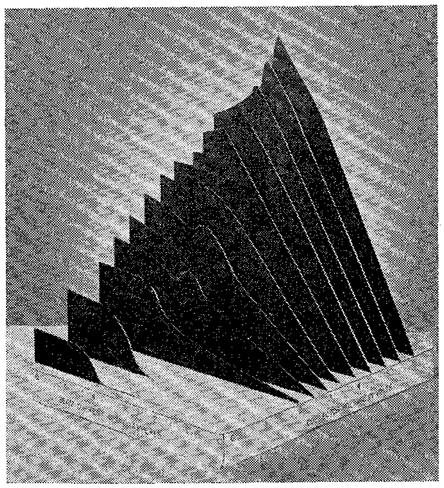

Fig. 2—Three-dimensional graph of measured collector current for type VAD-161-4 tube.

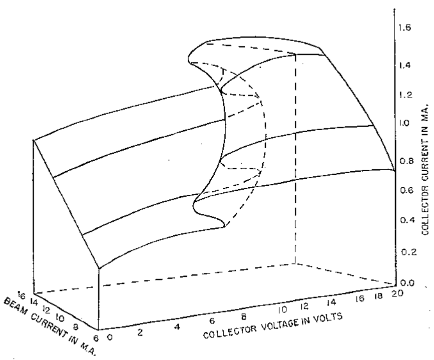

Fig. 1—Three-dimensional graph of measured collector current for type VAD-161-2 tube.

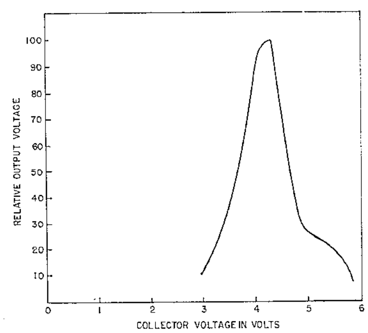

Fig. 3—Detected output level vs collector voltage.

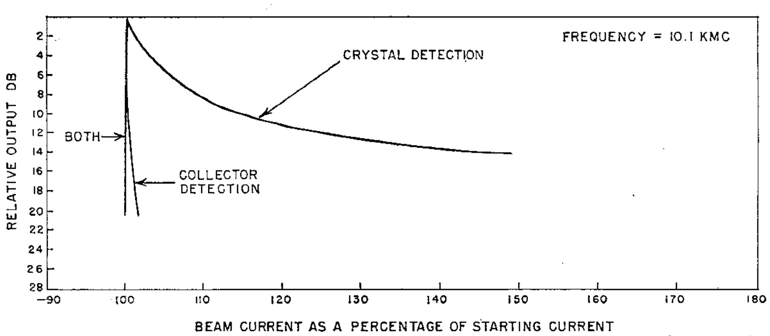

Fig. 4—Normalized output for two types of detector vs beam current.

The backward wave oscillator tubes used in these experiments were developed by Varian of Canada, Ltd., under the auspices of the Defence Research Board, Canada, (Electronic Components Research and Development Committee).

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A Broad-Band Crystal Mount 10.5 kmc to 20 kmc*

Laboratory measurements at microwave frequencies very often require the use of a sensitive detector with broad-band characteristics.

A crystal mount has been developed with a nearly uniform response from 10.5 kmc to 20.0 kmc using coaxial crystals type 1N26 or 1N78A. The best over-all sensitivity is obtained with 1N78A crystals, although the 1N26 crystals are more sensitive from 17 to 20 kmc.

Basically, the mount consists of a section of type RG-91/U waveguide containing a tapered ridge guide-to-coaxial-line junction [1]. The first design matched the waveguide to a 65-ohm coaxial line, but as the crystal impedance was not 65 ohms the dimensions had to be modified considerably to obtain the maximum sensitivity across the frequency band. Dimensional details of the crystal mount are shown in Fig. 1. A low impedance between the crystal body and the mount is obtained by means of an insulated sleeve which forms a capacity of 50 $\mu\mu f$. With a bias current of 75 μamp the video impedance is about 700 ohms, which results in a rise time of less than 0.1 μsec.

The narrow dimension of the waveguide was reduced from the standard 0.311 inch to 0.281 inch to eliminate a sharp dip in sensitivity at 19.2 kmc. In the final version of this mount, the tapered ridge and the insert which reduces the narrow dimension of the guide were machined from one piece of metal, and then placed in the standard waveguide. This method of construction places the junction of the insert and the waveguide at the walls of the guide instead of at the base of the ridge. The results of electrical tests indicated that the performance of the mount does not depend on a good electrical contact at this junction. With this method of fabrication, the assembly of the mount will be greatly simplified.

During all measurements, crystals were biased with 75 μamp of forward current to improve both the detection efficiency and the RF impedance of the complete mount. The video amplifier following the crystal detector had an input impedance of 50 ohms and a bandwidth of 0.5 mc. A pulse-modulated signal was used, with a pulse width of 1 to 2 μsec. The measurements were limited to a few crystals of each type, as the effort

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