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]

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

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

$$P = \frac{kT_sB}{L} + kT_nBf, \tag{2}$$

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

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

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

$$P_{T_n=T_s} = kT_sB, \tag{4}$$

and combining (3) and (4) yields

$$f = 1 - \frac{1}{L}\cdot \tag{5}$$

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

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

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.

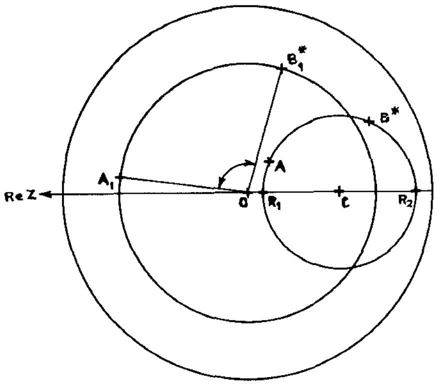

Fig. 1—Smith Chart.

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 $A_1 O B_1^*$.

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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[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.