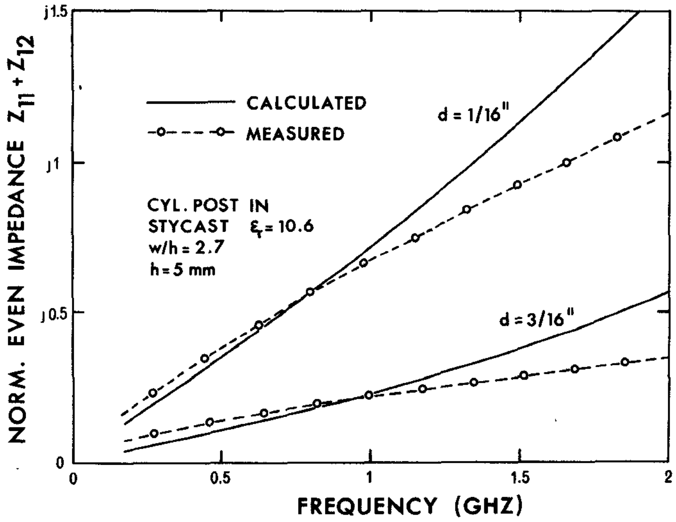

Fig. 8. Normalized even impedances of centered metallic posts of circular cross section in a 27-Ω line on Stycast (ϵ_r = 10.6; h = 5 mm; W/h = 2.7). Post diameter d = 1/16 in, post diameter d = 3/16 in.

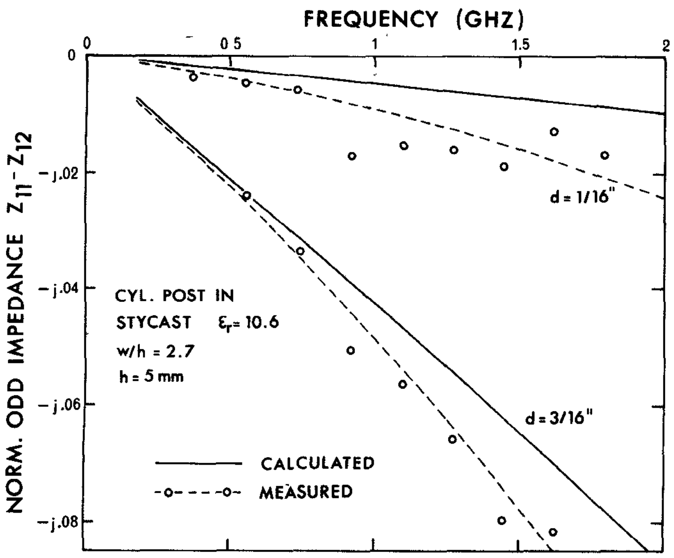

Fig. 9. Normalized odd impedances of centered metallic posts of circular cross section in a 27-Ω line on Stycast (ϵ_r = 10.6; h = 5 mm; W/h = 2.7). Post diameter d = 1/16 in, post diameter d = 3/16 in.

measured and calculated reactive impedances of posts of diameter $\frac{1}{16}$ and $\frac{3}{16}$ in for even and odd excitation.

It can be seen that the measured values for the even impedances form a smooth curve while the values measured for the odd impedance are more scattered. This confirms the observation made with respect to (15): the more the impedances differ from $\pm j1$, the more sensitive they are to errors in frequency measurements.

Taking into account the residual changes in temperature, small imperfections in the ring, and the uncertainties in the localization of resonance peaks, the total uncertainty in both f_p and f_e [in (15)] was typically ±50 kHz over the whole frequency range covered in the experiment. On this basis, the relative error in the discontinuity impedances is as shown in Table I.

The theoretical values which are presented in Figs. 8 and 9 for comparison are only approximate and differ from experimental results for the following two reasons.

1) They have been calculated using the variational principle and assuming only approximate simple current-distribution functions on the obstacle. If the correct current distribution were known, theoretical results would be closer to experimental values.

2) In these calculations, the microstrip was replaced by an idealized parallel plate model with magnetic sidewalls. This model does not incorporate radiation effects. Even though radiation losses appear to be very small, the reactive part of the radiation impedance affects the energy stored at the discontinuity. This phenomenon is currently being studied by the authors.

TABLE I
RELATIVE ERROR IN IMPEDANCE MEASUREMENTS AS A FUNCTION OF THE IMPEDANCE VALUE

$\lvert Z_i \rvert$	1	0.1	0.01	0.001
$\lvert \Delta Z_i/Z_i \rvert$	0.3%	1.8%	18%	180%

VI. CONCLUSION

The evaluation of the equivalent circuit of microstrip discontinuities in a resonant ring avoids the problems encountered when measuring through coaxial–to–microstrip transitions, the characteristics of which are poorly known. The Z parameters of a discontinuity are calculated from the resonant frequencies and Q factors of the ring before and after the introduction of the discontinuity. Normalized reactances between $\pm j100$ and $\pm j0.01$ can be measured with satisfactory accuracy. Evaluation of losses is more difficult since this involves measurement of changes in Q factors, a procedure which is inherently less accurate than the measurement of resonance frequencies.

The method is excellent for checking theoretical expressions for microstrip discontinuity parameters, and for characterizing microwave-integrated-circuit elements.

ACKNOWLEDGMENT

The authors wish to thank Dr. D. S. James for fruitful discussions and for supplying the microstrip ring.

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The Accurate Measurement of Range by the Use of Microwave Delay Line Techniques

GERALD F. ROSS, FELLOW, IEEE

***Abstract*—A scheme is presented for accurately measuring range to a radar target by the use of microwave delay line techniques and the use of solid-state subnanosecond digital threshold circuitry. The scheme obviates the need for expensive high-speed counters or analog thresholding and is cost effective to implement. A breadboard**

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The author is with the Sensor Systems Department, Sperry Research Center, Sudbury, Mass. 01776.

design of the technique was constructed and schematic diagrams are presented in this short paper. The results of the breadboard tests indicate that the range to a target can be measured and indicated up to 250 ft in 2-ft increments at a cost below $100.00.

A common problem in short-range radar applications is the accurate measurement of range to a target. At distances of several hundred feet, range accuracies of around 1 ft have been obtained with CW-FM radars [1]; in this short paper we describe a new ranging measurement technique with comparable accuracy for BAseband Radars (BAR). In short-pulse BAR applications [2], [3] a pulse several hundred picoseconds in duration is radiated into free space, reflects off a target that is perhaps several hundred feet away, and returns to a receiver where a threshold detector is activated, producing a reconstituted short pulse. The problem is to cost-effectively measure the time between the two epochs (i.e., the main bang and the output of the threshold detector) and convert this time duration to an accurate range measurement.

Both analog and digital approaches can be employed to solve this problem. In the analog approach, a ramp function is initiated at the start of the main bang and stopped when a target is received; the dc voltage obtained at this point is held until the next pulse. When the ramp duration is measured in nanoseconds and the pulse train period in hundreds of microseconds, it is difficult to accurately hold and read the resulting dc voltage (i.e., the range). A range measurement can also be obtained digitally by counting the number of RF cycles produced by an L-band oscillator, a start command to the counter being given at the time of the main bang, and a stop command being issued when a return is received To count at a gigahertz rate, however, requires expensive, elaborate hardware, especially if range accuracy in the order of 1 ft is required.

The purpose of this paper is to describe a scheme involving the use of a tapped microwave delay line which achieves the accuracy obtainable by high-speed digital techniques but at a significant cost reduction (e.g., a factor of 10 or more). To explain the scheme, consider the two identical pulse generators connected together through a length L of a TEM-mode transmission line and of characteristic impedance r_0 as shown in Fig. 1.

Assume that at $t = 0$ each generator produces a rectangular pulse $p(t)$ whose duration is $\tau \ll L/c$, where c is the speed of light in the line. Then an observer standing $L/2$ m from either source would experience a voltage $2p(t)$ for τ seconds, and at the other instants the voltage would be identically zero. An observer standing at any other point on the line would experience the pulse $p(t)$, only at two different times, provided the source impedance of each generator equaled r_0, the characteristic impedance of the line (i.e., there are no reflections). Thus a simple length of TEM-mode line acts as a most efficient summing network.

If a time delay is placed in series with one generator then it is clear from superposition that a doubling of the voltage will occur at some new point on the line. By placing a series of taps along the line, it is possible to estimate this delay by, essentially, utilizing space–time sampling. For example, consider that the generator at the left in Fig. 2 produces a pulse $p(t)$, synchronized with and reconstituted from the main bang of a BAR system. The returned signal from a target in space initiates a reconstituted video pulse T seconds later forming a new pulse $p(t - T)$. If the maximum and minimum ranges to an expected target are $D_{\max}$ and $D_{\min}$, respectively, then for the pulses $p(t)$ and $p(t - T)$ to coalesce somewhere within the line length L, two equations must be satisfied; namely,

$$cT + L = 2D_{\max} \tag{1}$$

and

$$cT = 2D_{\min} + L. \tag{2}$$

Solving (1) and (2) for cT and L yields

$$L = D_{\max} - D_{\min} \tag{3}$$

$$cT = D_{\max} + D_{\min}. \tag{4}$$

For example, if $D_{\max} = 250$ ft and $D_{\min} = 50$ ft, then from (3), 200 ft of line length is required. By placing taps on the line 2 ft apart, a range-resolving capability of 2 ft can be obtained. Although the use of 100 taps is an approach, a more desirable alternative involves limiting the number of taps to, perhaps, 10 and limiting the distance $D_{\max} - D_{\min}$ to a 20-ft increment. From (4) the delay length cT would be set first to 120 ft, then to 160 ft, then 200 ft, $\cdots$, and then finally to 480 ft. It should be noted that current technology permits tap spacing of only inches if desired; this also corresponds to a range resolving capability of only inches.

The stepping of the time delay T can be accomplished by initiating an impulse comb train of pulses in synchronism with the main bang as shown in Fig. 3(a). In the example, the spacing between pulses is 40 ns. The spacing of pulses in the comb oscillator train can be made to depend upon a length of line in the feedback loop of a pulse oscillator and therefore can be accomplished very accurately. A scanning pulse [see Fig. 3(b) and (c)] selects the desired member

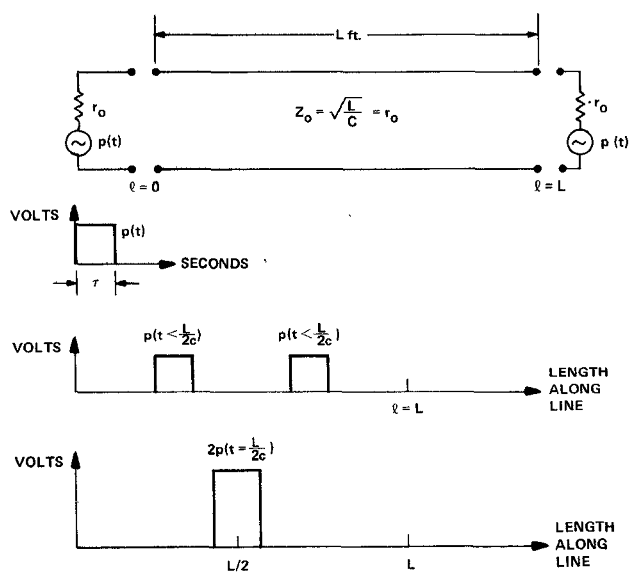

Fig. 1. Basic transmission-line concept.

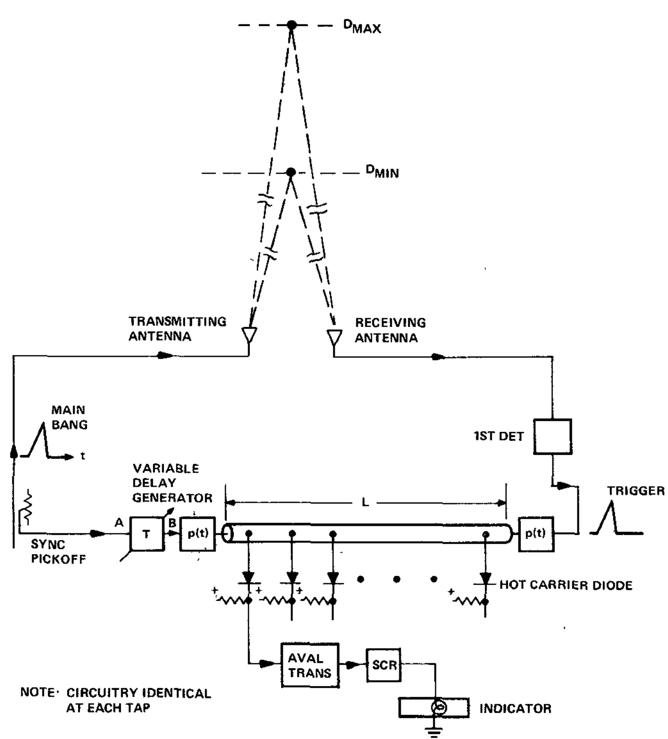

Fig. 2. Block diagram of ranging scheme in the BAR system.

of the pulse train (i.e., the delay T). When coincidence occurs at a given tap, the back voltage of a hot-carrier diode current is exceeded and an avalanche transistor is excited. This, in turn, illuminates and latches an indicator lamp.

Thus by using a number of taps on a transmission line in conjunction with a 25-MHz oscillator (i.e., the reciprocal of the 40-ns delay between pulses in the comb train illustrated in Fig. 3), the resolution previously obtainable only by counting at a gigahertz rate can be achieved. A model of a ranging unit employing this technique has been built and successfully demonstrated.

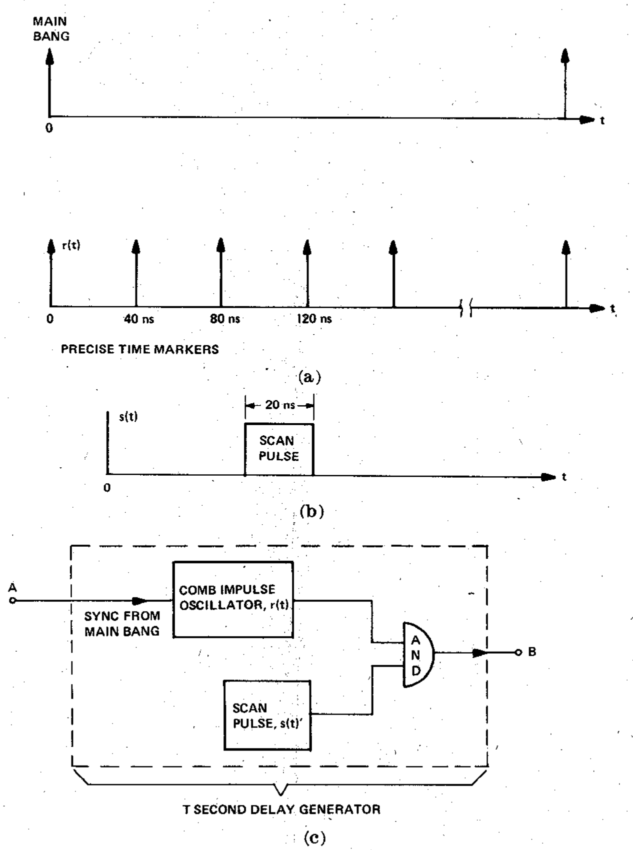

Fig. 3. Description of delay generator.

EXPERIMENTAL MODEL

An experimental model of the ranging scheme was constructed employing the ten-tap TEM-mode delay line described previously: the total length of the line L was 20 ft. The electronics at each tap location were mounted on a 2-in-sq circuit board and stacked in a chassis about 10 in in length. Each circuit board was interconnected by a 2-ft (electrical) length of RG174 cable. The duration of the pulse $p(t)$ was designed to be somewhat in excess of 2 ns to ensure that at least one vernier indicator light in the stack would operate when a target was present. A photograph showing the mounting of the circuit boards on the breadboard chassis and one section of the interconnecting cable is shown in Fig. 4.

To minimize the probability that any of the vernier range indicators would activate on false noise spikes during the time the system dwells in any range sector, the tap circuitry was designed to indicate only after four consecutive hits had exceeded the local threshold device. This is accomplished using the circuit shown in Fig. 5. The output from the hot-carrier diode at the tap feeds an avalanche transistor threshold device, a four-pulse decoder, and an N-P-N transistor and indicator. If three or less hits are registered at any of the ten taps for a given coarse range interval, then all the tap decoder chips are reset to zero and the coarse range is stepped

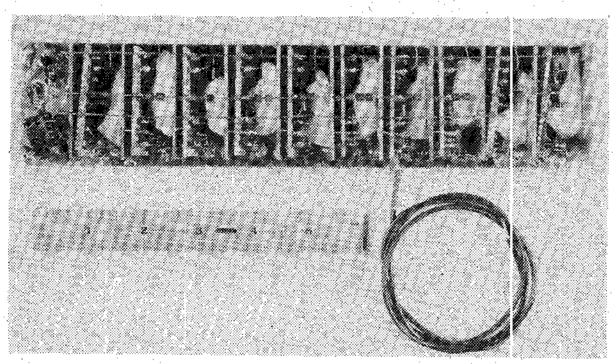

Fig. 4. Breadboard tap circuitry.

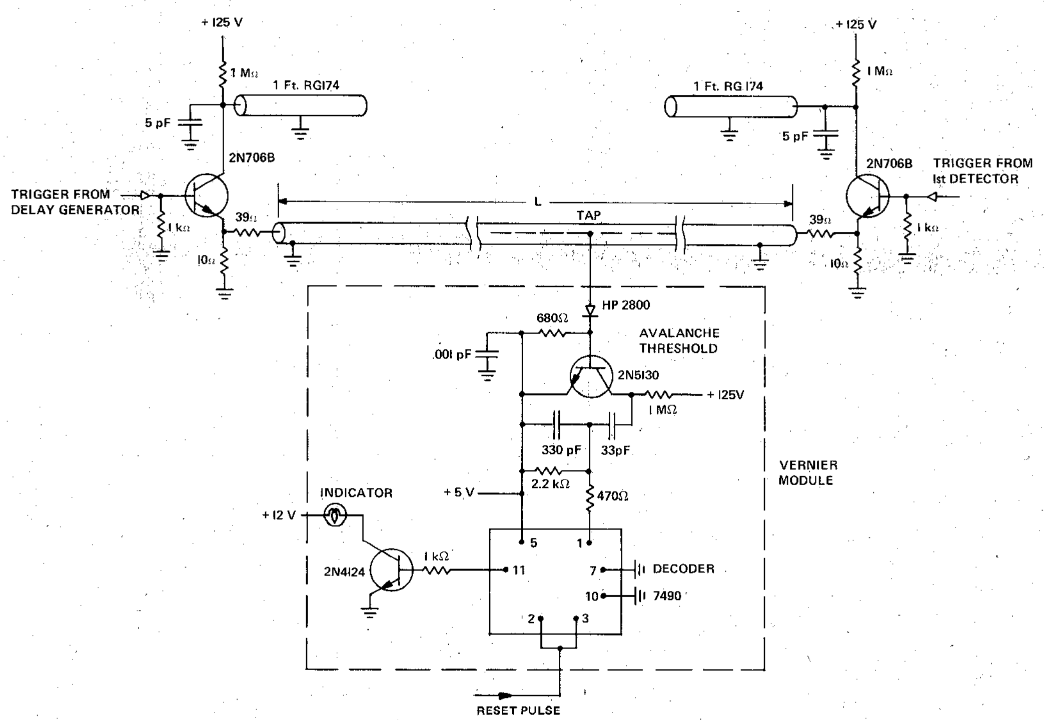

Fig. 5. Schematic diagram of local tap circuitry.

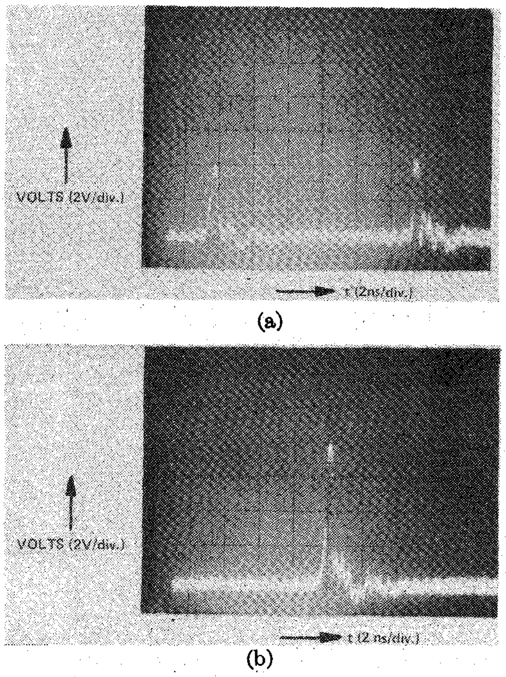

Fig. 6. Waveforms observed at a tap.

to the next increment each fifth transmitted pulse. In this manner, the entire range sector is scanned once every 50 transmitted pulses, a 200-Hz range-scan rate for a pulse-repetition frequency (PRF) of 10 kHz.

A sampling oscilloscope may be placed at any tap position and the resulting waveform may be observed as shown in Fig. 6. In Fig. 6(a) we observe the reconstituted baseband pulse return from a target located at an improper coarse range; the time scale is 2 ns/div and the ordinate is 2 V/div. At the appropriate coarse range and vernier tap, the two pulses coalesce as shown in Fig. 6(b) and exceed the threshold voltage of the avalanche transistor set at 6 V. The avalanche transistor is a monostable device which automatically resets a prescribed time after it discharges awaiting another trigger.

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A Transferred-Electron Frequency Memorizer

W. R. CURTICE, SENIOR MEMBER, IEEE

***Abstract*—Transferred-electron oscillators capable of single-frequency operation at any of as many as 20 closely spaced frequencies have been constructed for microwave-frequency-memory applications around 11 GHz. Switching between different frequency states has been achieved with a single RF pulse as short as 0.1 μs.**

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The author is with the RCA Laboratories, David Sarnoff Research Center, Princeton, N. J. 08540.

INTRODUCTION

In many microwave signal-processing systems, such as ECM systems, it is desirable to acquire an external input signal and to retain it for a long period of time. One technique for producing time delay is called the "loop" memory system; it consists of an amplifier with a delay line in the feedback path. A second method is to construct a multiposition frequency memory or register that will continue to oscillate at any one of a large number of assigned frequencies until shifted to another by a new input signal. Edson [1] has described the general properties of both systems in a classic article. He showed that the loop memory system is also a form of frequency register where the frequency separation is equal to the reciprocal of the loop time delay. Many of Edson's conclusions (e.g., system behavior in the presence of multifrequency signals, the number of modes possible in practice for each type of frequency register, etc.) appear to be valid at microwave frequencies.

The purpose of the present study is to investigate the feasibility of a frequency-memory register constructed at microwave frequencies using conventional transferred-electron devices (TED's). The TED frequency memorizer utilizes a multiresonant RF circuit to enable oscillation at as many as 20 RF frequencies. A range of over 2 GHz has been achieved with mode spacing of about 132 MHz. Memorizers with smaller mode spacing and the power requirements for switching with pulsed RF input have been studied. Magarshack [2] has previously demonstrated one type of TED frequency-memory oscillator. He points out that such devices may also have application in the field of telecommunications for switching between channels.

RF PERFORMANCE

Fig. 1 shows the equipment layout used for evaluating each frequency memorizer. The TED was operated with dc bias, and a pulsed RF input signal was used for switching. Tests were usually conducted with pulse repetition rates of 1 kHz and pulsewidths of 0.1–1.0 μs. Provision was made so that a single pulse of RF could be produced after setting the RF source to a new frequency. By this means it was verified that switching always occurred during the first pulse of the input RF signal.

Each memorizer tested was found to have a set of stable frequency states. A state was deemed stable if, after switching, the memorizer would continue to oscillate in that state when the input signal was removed. Some states exist only with the presence of additional signals at other states. These states have multiple frequency output.

The number of stable frequency states was affected by the bias voltage of the TED as well as by circuit tuning. The RF frequency and RF power output for each stable state was measured for fixed bias value and circuit tuning condition. This measurement shows the number, range, and separation of the frequency states instantaneously available for an input signal.

Figs. 2 and 3 show the frequency states measured for two experi-

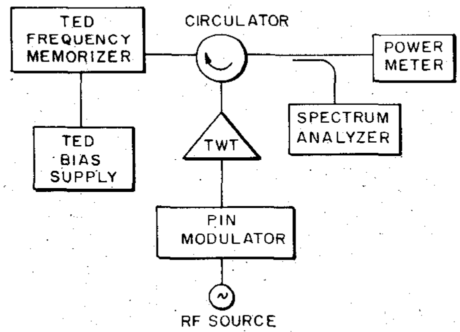

Fig. 1. Equipment layout for switching tests of a TED frequency memorizer.