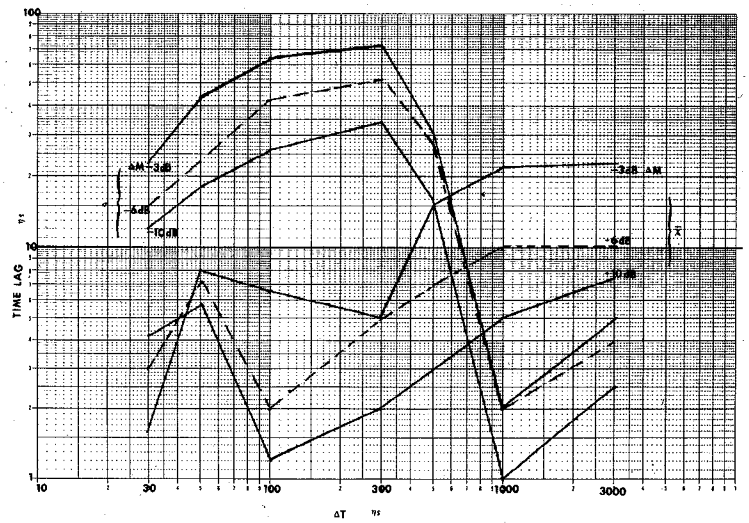

Fig. 10. Experimental mean ($\bar{x}$) and standard deviation (σ) per pulse bias error as a function of multipath time delay difference ΔT for several values of multipath amplitude difference ΔM.

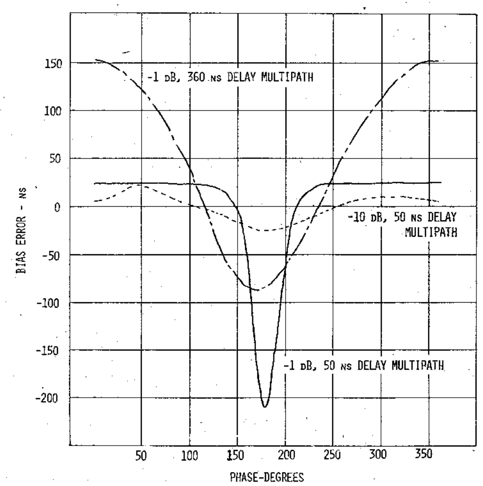

Fig. 11. Bias error versus multipath phase.

Additional benefits include a 30-dB adjacent channel selectivity rather than 8 dB and a requirement for a receiver with only 13 dB of dynamic range instead of 60 dB.

A detailed evaluation of system errors unique to pulse compression systems shows that a floating detection threshold system using alternate up and down chirp transmission will have no sources of error not present in the existing and proposed pulsed CW systems. TOA errors due to multipath have been experimentally determined and shown to be less than 24 ns (12 ft) for expected worst case conditions.

ACKNOWLEDGEMENT

The authors wish to thank J. Couvillon and B. Maher for the many technical contributions they made. The computer simulation was done by C. T. Cadenhead.

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Folded-Line and Hybrid Folded-Line Bandstop Filters

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***Abstract*—The feasibility of a compact bandstop-filter geometry is demonstrated. The filter geometry is particularly suited for stripline and microwave-integrated-circuit (MIC) fabrications in that grounding is not required for any part of the filter. Hybrid geometries allow the filter designer increased flexibility in choosing a suitable shape factor without significantly affecting the filter's electrical characteristics. Folded-line filter geometries are suitable for narrow bandwidths (provided capacitive-coupled stubs are used) to wide-bandwidth applications. Experimental confirmation is presented.**

INTRODUCTION

The fabrication of microwave components and systems in stripline and microwave integrated circuits (MIC) has advantages to alternative realizations with respect to size, weight, and (usually) reproducibility. To minimize the number of substrates in a given system or subsystem, the number of components per substrate should be maximized consistent with isolation requirements between components. The compact bandstop-filter geometry described in this

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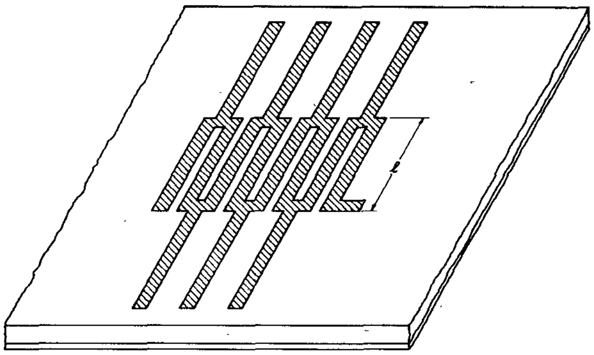

Fig. 1. Folded-line bandstop filter.

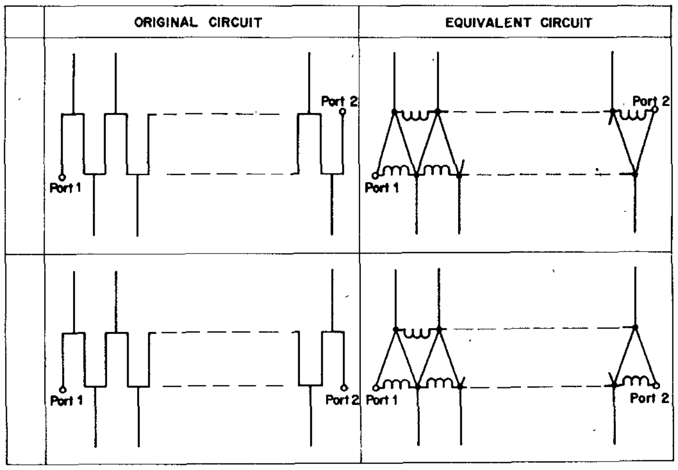

Fig. 2. Equivalent circuits for folded-line bandstop filters.

short paper, referred to as a folded-line geometry, utilizes the substrate surface area efficiently, and the design method allows for alternative hybrid geometries that give essentially the same electrical performance. The folded-line bandstop-filter geometry is shown in Fig. 1. Conceptually, it may be thought of as a meander line with quarter-wave stubs connected to the meander-line turns, or as a conventional quarter-wave bandstop filter (hereafter referred to as a linear-geometry bandstop filter) which has been folded in an accordian fashion. The distance between folded quarter-wave lines is sufficiently small so that coupling between turns is nonnegligible. This has the effect of moving the transmission zeros at $S = \pm 1$[1] to elsewhere in the complex plane. However, coupling between adjacent stubs is not allowed in the present design.[2]

TECHNICAL DESCRIPTION

The designs of the folded-line filters reported in this short paper were accomplished with computer-aided design methods based on an extension of techniques previously used for meander-line transformers [2]. Fig. 2 gives the physical form and equivalent circuits for folded-line filters [3], [4]. Those segments in the equivalent circuit shown as straight lines represent quarter-wave lines (unit elements) [5]. Inductor symbols represent short-circuited quarter-wave lines (S-plane inductors) [5]. Coupling between folded lines is proportional to the admittance of the S-plane inductors. The objectives of the computer design program were to obtain an equiripple passband response and maximum stopband selectively. This was accomplished by minimizing a least pth objective function of a weighted reflection coefficient in the passband and the reciprocal of a weighted reflection coefficient in the stopband. The network response was computed using the equivalent circuit in Fig. 2. The weighting function was adjusted so that the coupling between turns would be within limits set by the user prior to the running of the program. In the compilation of design tables described later in this short paper, $-12-18$-dB coupling was selected for all turns. The coupling $k_{i,i+1}$ between turns i and $i+1$ is defined as

$$k_{i,i+1} = 20 \log_{10} \frac{C_{i,i+1}}{[(C_{g_i} + C_{i,i+1})(C_{g_{i+1}} + C_{i,i+1})]^{1/2}} \text{ dB}$$

where

C_{g_i} capacitance to ground per unit length for the ith conductor;
$C_{i,i+1}$ mutual capacitance per unit length between the ith and ith $+ 1$ conductors.

Couplings between -12 and -18 dB are small enough to give practical physical realizations yet large enough to give compact filters. The computer program permits arbitrary choices of couplings between turns, including no coupling between any or all turns. Negligible or no coupling is specified in designing hybrid filters, a few examples of which are shown in Fig. 3. Hybrid filters consist of geometries of folded and straight lines. However, the filter is designed as an integral unit. Hybrid geometries permit a choice between a number of physical realizations each of which gives essentially the same electrical performance.

Initial parameter values used in the computer program were taken from the corresponding linear-geometry optimum designs. The couplings between turns were then gradually increased until $-12-$ -18 dB was obtained. Subsequent designs were obtained by stepping the bandwidth and using the final design values from a previous case as initial design values for the new case. Extrapolation of the stub values was found helpful. Design tables were obtained for symmetrical filters of degrees 5 and 9, where the filter degree is defined as the sum of the number of lines and stubs. The bandwidths obtained were

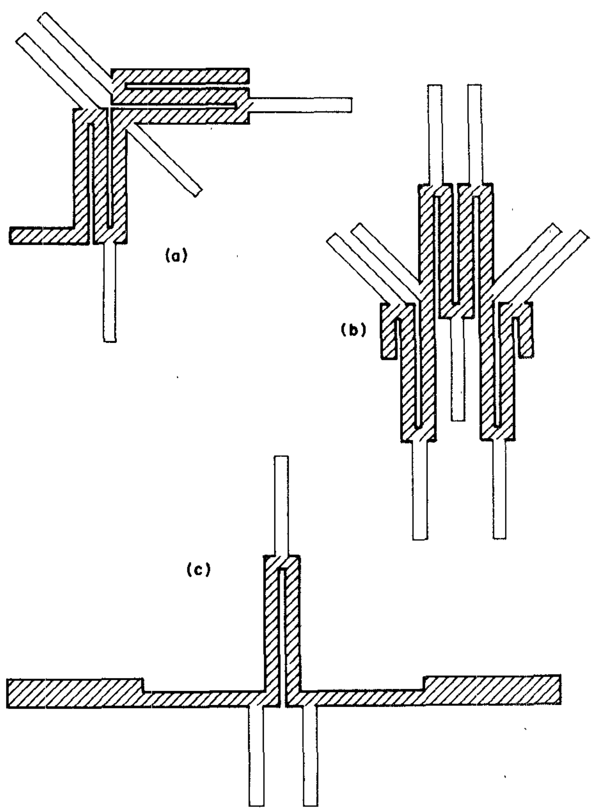

Fig. 3. Examples of hybrid folded-line bandstop filters.

[1] $S = j \tan(\theta)$ (Richards) transformation [1].
[2] Designs for which coupling also exists between stubs may yield superior filter characteristics. However, physical realizations would be quite difficult in most cases. It is worthwhile to point out that the computer design procedure described in this short paper could be easily extended to the latter case.

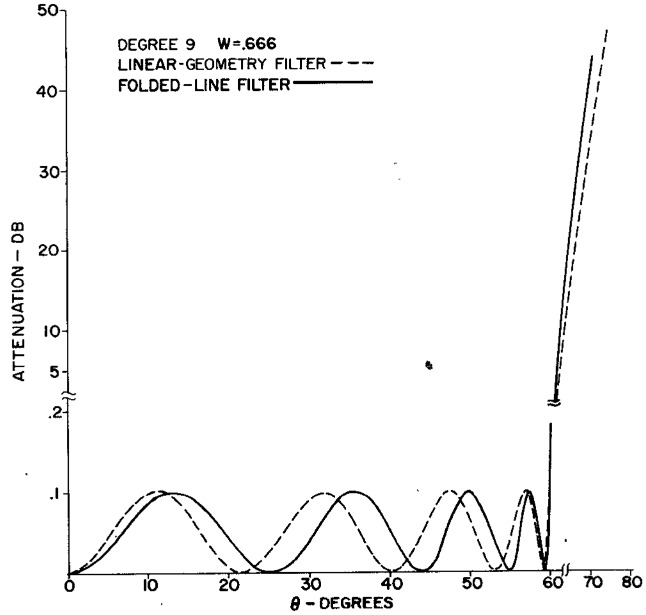

Fig. 4. Attenuations of linear-geometry and folded-line bandstop filters.

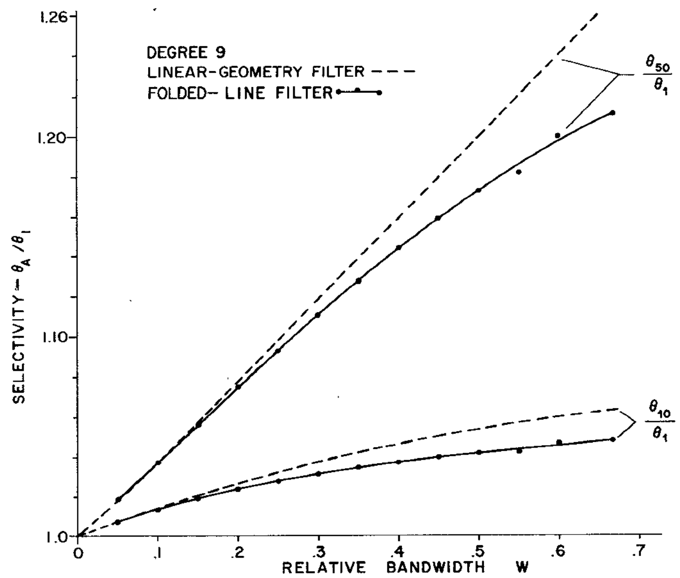

Fig. 5. Selectivities of linear-geometry and folded-line bandstop filters.

0.025–0.300 in steps of 0.025, and 0.050–0.600 in steps of 0.050 for the fifth- and ninth-degree filters, respectively. The octave bandwidth case was also included in the ninth-degree filter. Both tables were for 0.1-dB equiripple passbands.[3] The format and application of the tables are identical to those given in [2], and consequently a description will not be repeated here.

For a given filter degree, folded-line and linear-geometry bandstop filters have nearly identical electrical characteristics. Fig. 4 compares the theoretical attenuation responses for a linear-geometry and folded-line filter of the ninth degree and stop bandwidth of 67 percent. The principal differences are seen to be a shift in the ripple extremes in the passband and an increase selectivity for the folded-line filter. The shift in ripple extremes in the passband is, of course, inconsequential. The differences in stopband selectivity are small for narrow-band filters, but they become appreciable as the bandwidth increases. For example, in the degree-9 case a folded-line filter having a fractional bandwidth of a 0.50- and 0.1-dB ripple has approximately 14 percent greater selectivity than the corresponding linear-geometry filter at 50-dB attenuation.

Fig. 5 presents more detailed comparisons of selectivities between the linear-geometry and folded-line filters. The curves give the ratio of electrical degrees at 50-dB (and 10-dB) attenuation to the equiripple band edge.

EXPERIMENTAL RESULTS

To experimentally demonstrate the feasibility of the folded-line bandstop filter two trial designs of fifth and ninth degrees having differing bandwidths were constructed and tested. The first to be described is the ninth-degree filter which had a 0.1-dB passband

[3] See NAPS document 02339 for design tables. Order from ASIS/NAPS, c/o Microfiche Publications, 305 E. 46th St., New York, N. Y. 10017. Remit in advance for each NAPS accession number $1.50 for microfiche or $5.00 for photocopies up to 30 pages, and 15¢ for each additional page. Make checks payable to Microfiche Publications.

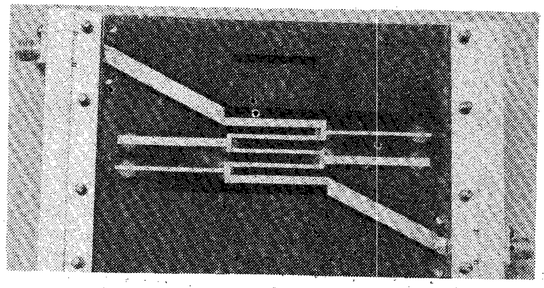

Fig. 6. Trial octave-bandwidth folded-line bandstop filter.

TABLE I
PARAMETER VALUES FOR NINTH-ORDER TRIAL FOLDED-LINE BANDSTOP FILTER

Line Number	Self-Capacitance Normalized to $376.7/\sqrt{\epsilon_r}$ $C_{g_i}/$ and $C_{g_{6-i}}/$	Mutual Capacitance Normalized to $376.7/\sqrt{\epsilon_r}$ $C_{i,i+1}/$ and $C_{5-i,6-i}/$	Stub Admittance Normalized to Terminating Admittance
1	0.6172	0.08834	0.5043
2	0.3945	0.06476	0.7066
3	0.4017		

ripple, a fractional bandwidth of 0.67, and a nominal center frequency of 1.5 GHz. The filter was constructed in stripline of copperclad rexolite ($\epsilon_r = 2.54$) and a 0.25-in ground-plane spacing [6]. A photograph of the filter with the top ground plane removed is shown in Fig. 6. Parameter values for the design are given in Table I. Definitions of the parameters are given in [2]. The filter was tuned to the center frequency using screws located at the ends of the stubs. Fine tuning for best return loss was achieved using tuning screws located at the junction of the stubs and folded lines. The attenuation and return loss were measured using swept-frequency sources from 0.2 to 4.0 GHz. The responses are shown in Fig. 7. It was not possible to achieve a 16.4-dB return loss (corresponding to 0.1-dB ripple) across the entire passband. However, a return loss of 14 dB (corresponding to 0.18-dB ripple) was easily achieved up to 3 GHz. This was felt to be good enough to confirm the design data. The skewing of the attenuation skirt above the center frequency was attributed to a slight mistuning.

A trial design of fifth-degree and 0.30 fractional bandwidth was also constructed and tested. It was found that the theoretical values of stub impedances were too large to realize directly. Instead, capacitive-coupled half-wave lines were substituted. An approximate equivalence between the theoretical stubs and the capacitive-coupled half-wave lines was established by requiring equality of resonance frequencies and equality of impedances at the lower frequency band edge. The impedances of the half-wave lines were arbitrarily chosen to be 80 Ω. It was next determined that coupling capacitors of 0.93 pF were required to satisfy the previously mentioned equalities.

The effects of the capacitive-coupled half-wave stubs on the filter responses were to contract the bandwidth and degrade the return loss above the stopband. However, the lower equiripple band edge was not displaced because of the manner in which the equivalence was specified. Computed responses for the exact and approximated filter given in Fig. 8 showed a bandwidth contraction of approximately 16 percent. Note, however, that the lower band edge and resonant frequency have been maintained quite accurately as anticipated.

The experimental fifth-degree filter was also constructed in stripline using copperclad rexolite and a 0.25-in ground-plane spacing.

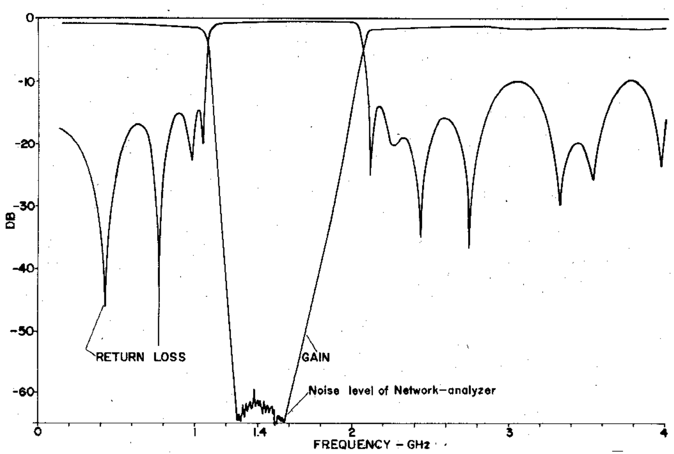

Fig. 7. Measured responses of the trial octave-bandwidth folded-line bandstop filter.

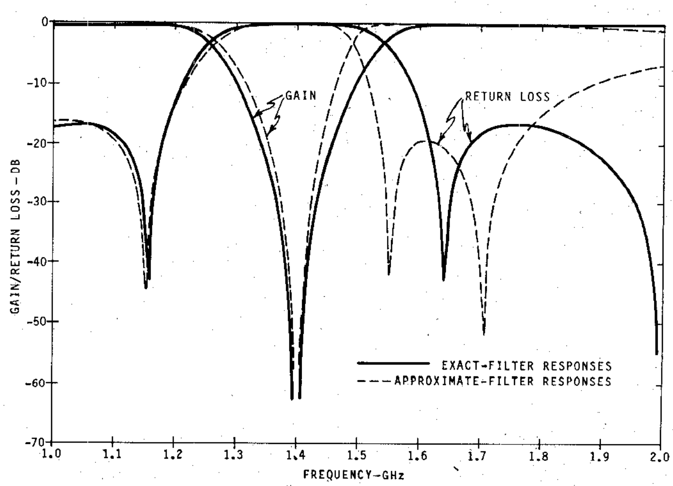

Fig. 8. Exact-filter and approximated-filter computed responses.

TABLE II
PARAMETER VALUES FOR FIFTH-ORDER TRIAL FOLDED-LINE BANDSTOP FILTER

Line Number	Self-Capacitance Normalized to $376.7/\sqrt{\varepsilon_r}$ C_{g_i}/ε and $C_{g_{4-i}}/\varepsilon$	Mutual Capacitance Normalized to $376.7/\sqrt{\varepsilon_r}$ $C_{i,i+1}/\varepsilon$ and $C_{3-i,4-i}/\varepsilon$	Theoretical Stub Normalized to Terminating Admittance	Coupling Capacitance (80Ω Stub)
1	0.7052	0.1121	0.1464	0.93pF
2	0.5423		0.1464	0.93pF

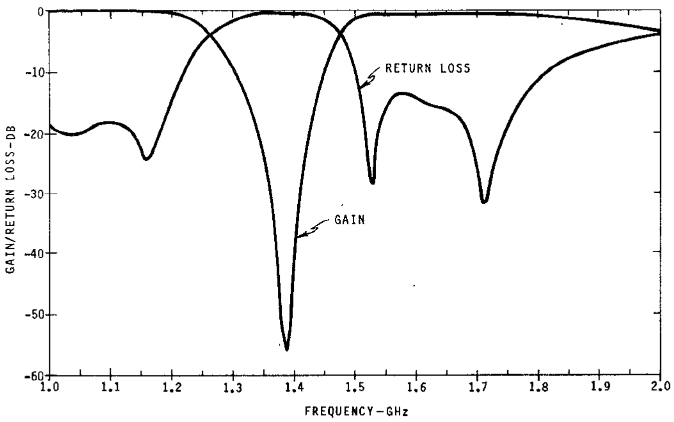

Fig. 9. Measured responses of a fifth-degree folded-line bandstop filter with capacitive-coupled half-wave stubs.

The nominal center frequency was chosen to be 1.4 GHz. Filter parameter values are summarized in Table II. The filter was tuned in the manner previously described, resulting in the measured responses given in Fig. 9. Optimum tuning was accomplished slightly below 1.4 GHz, presumably because the added length of the turns between lines was not accounted for in the stripline realization. The measured fractional bandwidth was approximately 0.24 which compares favorably with the expected value of 0.25. In fact, a comparison of the computed data of Fig. 8 and measured data of Fig. 9 shows generally excellent agreement between them.

CONCLUSIONS

Design tables for folded-line bandstop filters of fifth and ninth degrees and a 0.1-dB passband ripple were compiled for a range of bandwidths. Folded-line bandstop filters have greater selectivity than their linear-geometry counterparts of the same degree. The increased selectivity is small for narrow bandwidths but appreciable for wide bandwidths. Experimental results on two trial folded-line filters confirmed the designs and feasibility of the filter geometry. Folded-line and hybrid geometries together with linear geometries will give the filter designer increased flexibility in selecting stripline and MIC realizations.

ACKNOWLEDGMENT

The authors wish to thank Dr. C. K. Campbell for the use of his laboratory facilities and G. Kappel for drafting a number of the figures used in this short paper.

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An Automated General Purpose Test System for Solid-State Oscillators

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***Abstract*—An automatic test system for determination of the tuning and output characteristics of voltage-controlled solid-state oscillators is described. Hardware, software, and functional capabilities of the broad-band modular system are presented. The utilization of a minicomputer for data collection with on-line curve fit and rapid CRT display of test and analysis results facilitates device tuning and streamlines traditionally tedious measurement tasks.**

I. INTRODUCTION

There exist numerous measurement problems associated with the medium scale production of today's microwave systems which demand consideration in terms of test automation. The demand is economic in nature, for automation can improve both people and asset effectiveness.

Current technology supports microwave measurement automation in two ways. Complete general purpose network analyzers equipped to solve a wide class of problems such as broad-band measurement of S parameters are available [1]–[3]. However, there are many measurement requirements not covered by one of such proffered general purpose machines. For some of these problems there are programmable instrumentation components available which when innovatively integrated may constitute a test system of sufficient power to justify development costs [4].

The decision to "custom automate" has been a gamble of sorts because effort estimates are often difficult and results may not smoothly fulfill expectation and need. However, careful planning with a wary eye on modularity, flexibility, and utilization optimization can go far toward ensuring the success of the effort.

II. MANUAL TECHNIQUES—TIME CONSUMING, INEFFICIENT

The system to be described was designed to test solid-state voltage-controlled oscillators (VCO's) which serve as local oscillator (LO) elements in radar and guidance systems. Device types

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