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Optimal Design of Matching Networks for Microwave Transistor Amplifiers

***Abstract*—The design of input and output matching networks for transistor microwave frequency amplifiers has been optimized by the use of an efficient computer program. Power amplifiers capable of 2.5 watt peak power output with a 400 MHz bandwidth at 2.25 GHz have been fabricated. The matching network problem was reduced to an equivalent nonlinear programming problem by considering the *N*-filter elements as coordinates in a 2*N*-dimensional vector space. The optimal solution point in the vector space was found by the use of a "pattern search" routine which utilized randomly chosen orthogonal transformations of the search pattern to minimize an objective function. In this case, a suitable objective function was chosen to be the area under the curve of "reflection coefficient" vs. frequency for the filter input. By use of multiple data input as many as six designs have been achieved in less than nine minutes on the 7044 computer.**

Introduction

Construction of transistor amplifiers at microwave frequencies involves the design of suitable matching networks as shown in Fig. 1. Power amplifiers are further complicated by nonlinear operation to give efficiency as well as gain. The natural step to integrated microwave transistor amplifiers imposes a minimal size constraint upon the matching network design. Additional constraints such as transistor biasing networks and coupling capacitor dimensions must also be considered. The characterization techniques used to obtain the transistor input and output admittances have been previously described [1]. The admittances generally cannot be ascribed to a simple equivalent circuit. The technique described in this paper allows networks, consisting of lossless transmission lines, to be designed to match the measured admittances over a broad range of frequencies. An extension of the technique to lumped parameter elements also has been made.

We have characterized either port of a transistor as an admittance for the purposes of this discussion. This admittance is generally a function of both power level and frequency as shown in Fig. 2. Upon choosing an operating level, the admittance is given in terms of frequency alone. By least squares curve fitting, an admittance vs. frequency function may be empirically described. In general, two such admittance functions are given to be matched to one another by the use of a suitable network. The case described below utilizes lossless transmission line elements and capacitors to construct an admittance matching network over a broad band of frequencies.

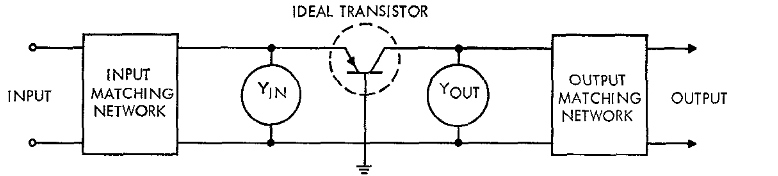

Fig. 1. Schematic diagram of a single stage transistor microwave amplifier. The output may be the input admittance of the next transistor stage.

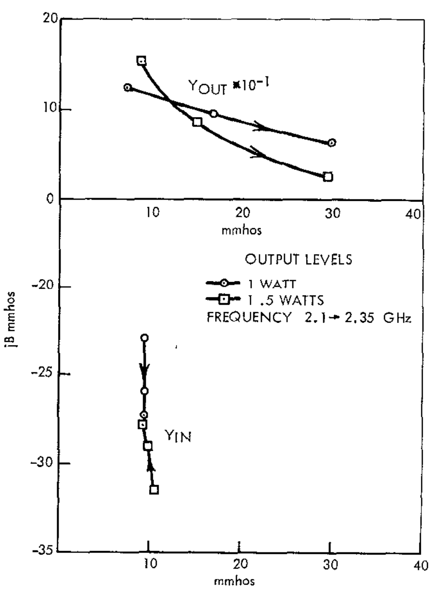

Fig. 2. Typical transistor admittance functions.

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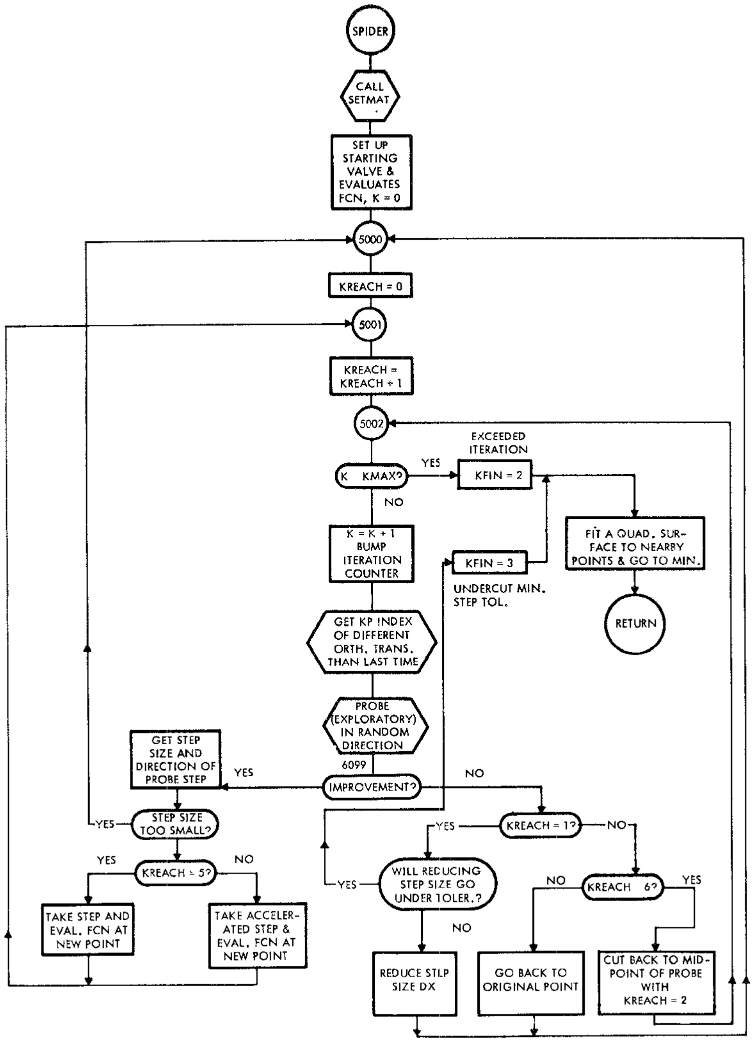

Fig. 3. Spider pattern search routine.

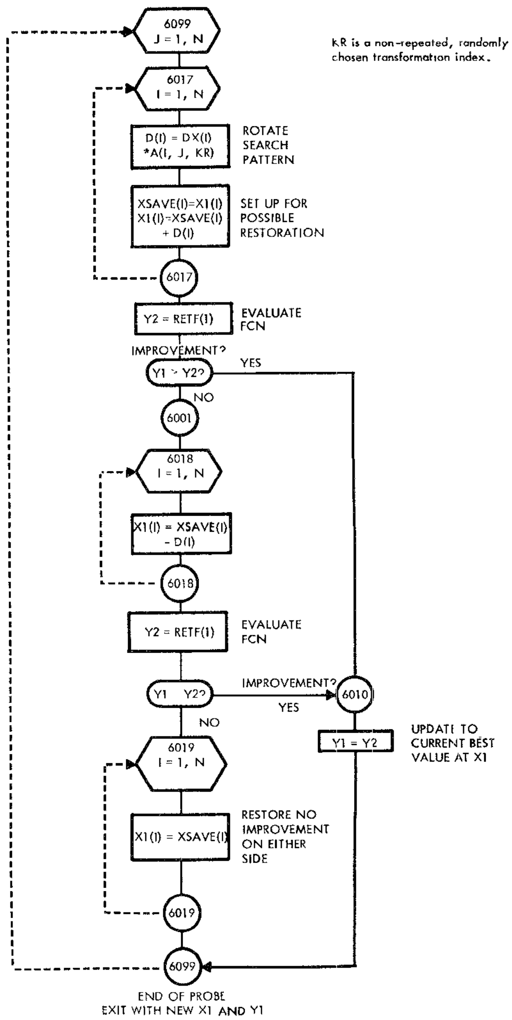

Fig. 4. Search routine program.

Network Design

The lossless line elements used are series sections, open stubs, and shorted stubs; the characteristic impedance and the individual line lengths, the capacitance and series resistance, are varied in the optimization procedure. A configuration of N elements thus describes the coordinates of a $2N$ vector space. By using one admittance as a load, the "reflection coefficient" between the transformed load admittance and the second (source) admittance may be calculated. Use of this function follows Fano [2], but the "reflection coefficient" actually employed is that function defined by Youla [3]. For a point in the vector space the "reflection coefficient" may be calculated as a function of frequency over the range of interest. Using a finite difference scheme and effectively integrating the resulting "reflection coefficient" function over this range results in an objective function to be minimized over the vector space.

The inclusion of constraints (upper and lower bounds) on the magnitudes of the $2N$ variables, imposed by practical design considerations, poses this as a special form (linear constraints with a nonlinear objective function) of a general nonlinear programming problem [4], [5].

The small number of elements ($N \leq 10$) allowed us to include the design constraints in the objective function by means of penalty functions. This, in turn, allowed us to utilize an efficient pattern search optimization routine called "SPIDER" for the mimimization process [6]–[8].

From the initial point a successful direction is determined by searching a fixed distance from the base point in a number of random orthogonal directions. If a successful direction is determined, a series of accelerating moves are made in this direction by a speed factor. If no successful direction is found with the current incremental values, then the increments are reduced. The exploratory moves are repeated with new increment values. Increments are successively reduced until the lower limit on the first variable is undercut, which initiates an exit routine. The maximum number of iterations may be specified to cause a separate (possibly nonoptimal) exit. Figures 3 and 4 diagram the search routine program.

At the present time the network configuration is at the discretion of the designer; an initial point and step sizes must also be given to the program. Boundary values for the transmission line elements are specified by the designer. This approach has been tested by designing transistor microwave power amplifiers. Table I gives a sample configuration, starting values, and final values for a network matching the output admittance of an amplifier to a 50 Ω load. Figure 5 shows a computer plot of the "reflection coefficient" vs. frequency curve achieved by the program. Figure 6 shows a 2.5 W peak-power amplifier designed using the computer program; Fig. 7 gives the frequency performance of the amplifier. The matching networks operate satisfactorily within the accuracy of the admittance measurements and the knowledge of the transmission-line design parameters.

TABLE I
NETWORK MATCHING OUTPUT IMPEDANCE OF AN AMPLIFIER TO A 50 Ω LOAD

Element No.		Initial Value Y_0 (mmhos)	Initial Value $\beta\ell$ at 2.4 GHz (degrees)	Final Value Y_0 (mmhos)	Final Value $\beta\ell$ at 2.4 GHz (degrees)
1	Series	15	45	12	14
2	Open Stub	15	45	20	62
3	Series	15	45	13	18
4	Shorted Stub	15	45	28	20
5	Series	15	45	23	15

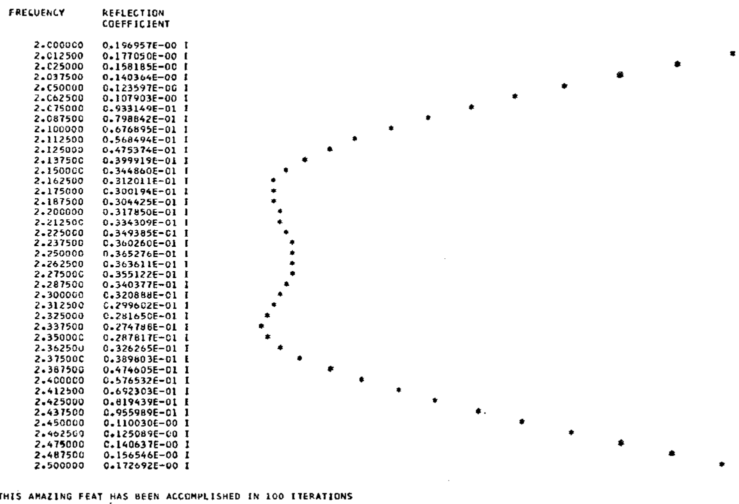

Fig. 5. Reflection coefficient vs. frequency.

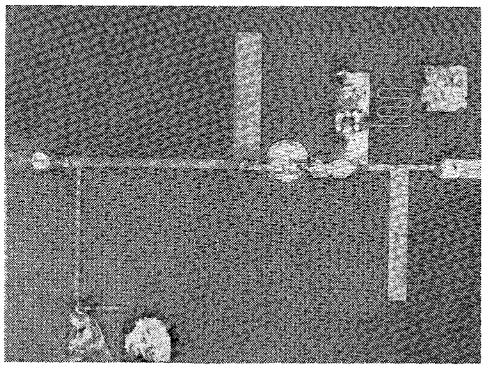

Fig. 6. Peak power amplifier (2.5 watts).

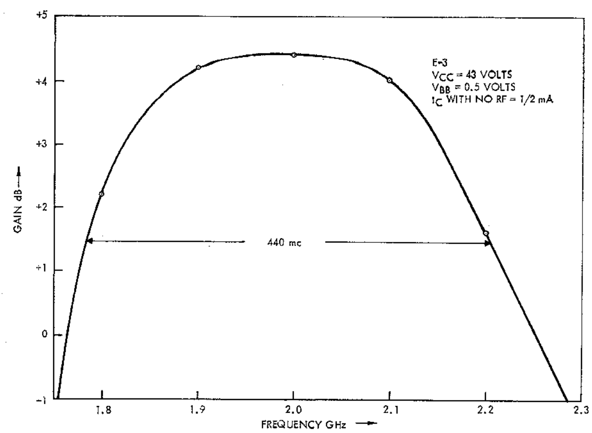

Fig. 7. Frequency performance of amplifier.

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Confocal Resonator Band-Pass Filters

Direct-coupled confocal[1] resonators are considered for use as band-pass filters at millimeter wavelengths in this correspondence. In previous work on band-pass filters for millimeter wavelengths [1], two flat reflectors were used to form resonators; these resonators could not produce high unloaded Q values because of the critical tolerances of maintaining parallelism between reflectors [2]. To overcome these difficulties, resonators with curved spherical surfaces have been used at millimeter wavelengths to achieve high Q_u [3]–[5]. Single-resonator Fabry-Perot interferometers and absorption wavemeters were considered (in these references), and the possibility of using them as band-pass filter elements was suggested [5]. One- and two-resonator band-pass filters are described herein together with experimental data. Emphasis is placed on types of coupling structures, reduction of spurious responses, and an extension to filters of arbitrary numbers of resonators.

Figure 1 shows a tunable, single-resonator, band-pass filter constructed of brass and operating in the 40 GHz region. Each reflector is machined into a cylindrically shaped piece which has a concentric thread. One end reflector has a right-hand thread, the other a left-hand thread. Rotation of the cylinder tunes the filter. Two steel guide rods are used to maintain reflector alignment.

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[1] The filters described are confocal and nonfocal; for conciseness, "confocal" will be used to describe either situation.