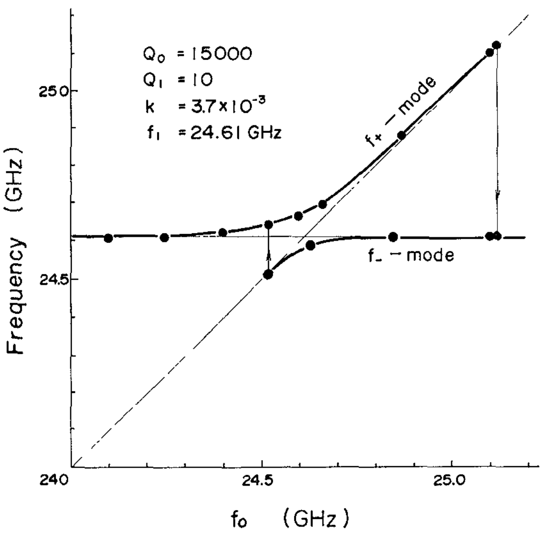

Fig. 4. Typical tuning characteristic of stabilized K_a-band Gunn oscillator as a function of f_0.

TABLE I
EXPERIMENTAL RESULTS

Operating frequency	25.01 GHz
output power	30 mW
Pushing figure	-0.7 KHz/mV
Frequency stability	-6 x 10^{-7} °C^{-1}
Bias voltage	-4 V
Dc current	500 mA

is negative at a frequency somewhat higher than the transit-time frequency f_T of the diode [7], [8]. We can obtain the condition $\tan\theta>0$ by decreasing the bias voltage V_0. The frequency f_T increases as V_0 decreases if the diode has the impurity concentration which is higher at the anode than at the cathode side [9]. In order to prevent mode jumping, the difference of the two frequencies $\Delta f(=f_0-f_1)$ should be smaller than the critical value [1], [2], which is 400 MHz in this oscillator. The results obtained in the stabilized oscillator are listed in Table I, where $\Delta f=400$ MHz. The frequency stability of 3×10^{-5} has been obtained over the temperature range from 0 to 50°C. The power loss has been minimized by adjusting the characteristic impedance of the transformer, and making the load conductance at $k=0$ smaller than G_1 [3]. The output power at $k=3.7\times10^{-3}$ was only 0.5 dB lower than the maximum generation power at $k=0$ and $f=25$ GHz. The frequency stability against the variation of the bias voltage is lower than 3×10^{-8}/mV. The oscillator can be operated in the frequency range up to 30 GHz by modifying the dimensions of the cavities. An output power of more than 20 mW has been obtained.

ACKNOWLEDGMENT

The authors wish to thank Y. Aono for the preparation of the Gunn diodes, H. Kondo and T. Itano for their help in experimental work, and Dr. H. Murakami, Dr. K. Ayaki, and Dr. K. Sekido for their encouragement and guidance.

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Broad-Band Parametric Amplifier Design

G. R. BRANNER, E. R. MEYER, AND P. O. SCHEIBE

***Abstract*—A computerized optimization technique is employed to provide design values for broad-band parametric amplifiers. Some results of the procedure are presented.**

Parametric amplifiers are of primary importance for design of low-noise microwave receivers. The importance of broad-band low-noise parametric amplifiers for use in satellite communication systems has been reemphasized recently [1], [2]; and it has also been suggested that frequencies above 10 GHz are quite attractive for commercial satellite systems since they are not so heavily utilized. In the past, bandwidths as high as 20 percent (for 20-dB gain stages) have been obtained at lower frequencies; however, at frequencies above X band, considerably smaller bandwidths have been reported.

It is the objective of this correspondence to present the results obtained from a computer-oriented optimization technique which is capable of providing realizable designs for extremely wide-band parametric amplifiers (40-percent 3-dB bandwidths have been obtained at low K band). The optimization procedure employs the Gauss–Newton iteration technique (requiring exact evaluation of all the essential partial derivatives of the objective function [3], [4]).

As is well known, the variation of parametric amplifier gain with respect to frequency is strongly dependent on the form of tuned circuits at signal and idler frequencies, and considerable effort has been expended toward the development of analytical techniques for selecting resonator parameter values to provide broad bandwidth [5]–[7].

The approach described in this correspondence is to utilize a prescribed amplifier topology (triple-tuned signal and single-tuned idler circuits) and employ computerized optimization to obtain realizable matching network element values which provide maximally flat bandwidths.

The nondegenerate parametric amplifier configuration which was employed is illustrated in Fig. 1(a). The varactor model [Fig. 1(b)] includes parasitic case capacitance (C_p), series lead inductance (L_0), and internal resistance (R_0). It was also assumed that signal and idler circuits are isolated by ideal bandpass filters centered, respectively, in the signal and idler bands of interest.

The mathematical model for the amplifier was developed using standard analytical techniques (see [8] for example):

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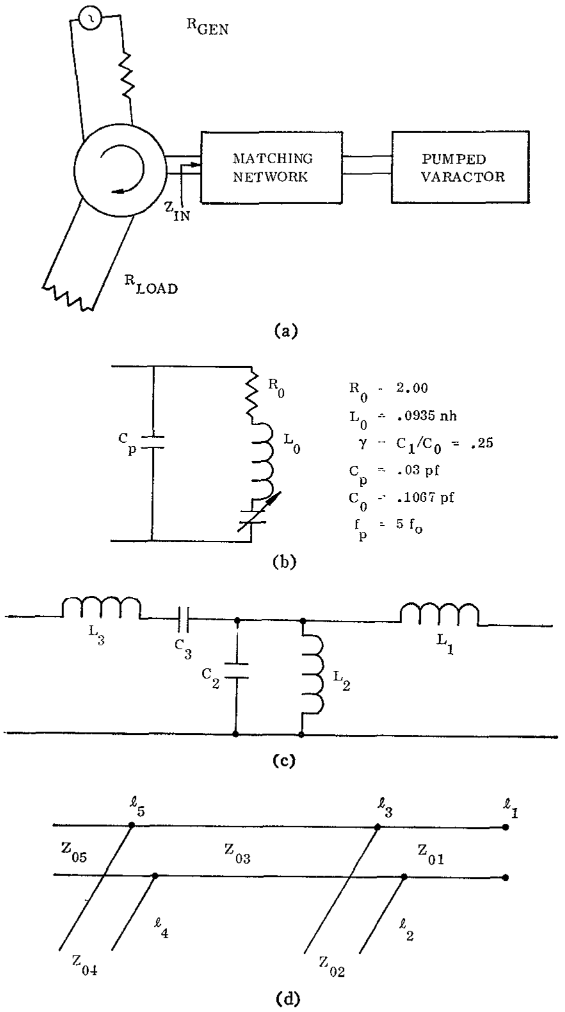

Fig. 1. Parametric amplifier and associated circuits. (a) Parametric amplifier. (b) Varactor model. (c) Lumped matching network. (d) Distributed matching network.

$$\text{transducer power gain} = G_t = \left|\frac{Z_{in} - Z_0}{Z_{in} + Z_0}\right|^2$$

where Z_{in} represents the input impedance of the matching network terminated in the pumped varactor as seen from the circulator see [Fig. 1(a)], and the circulator is replaced by its characteristic impedance Z_0.

A recently developed procedure for producing source code for algebraic partial derivatives [9] was employed to obtain the partial derivative of the transducer power gain expression with respect to the essential matching network parameters, and this overall mathematical model was subsequently utilized in the optimization routine (a more detailed discussion of the optimization routine and the aforementioned process will be presented in a forthcoming paper).

A lumped passive matching network [Fig. 1(c)] was employed initially to gain facility with the optimization strategy. Fig. 2 illustrates optimum responses obtained for a prescribed flat bandwidth of 30 percent employing the lumped matching network. (It should be mentioned that the responses for 14 and 15.5 dB were not optimized to be maximally flat since they were not of primary interest.) When the matching network of Fig. 1(c) was modified by exclusion of capacitor C_2, improved flat bandwidths were obtained. Fig. 3 represents the results obtained for prescribed 3-dB bandwidths of 40 percent. (Table I contains representative element values for several of the curves presented in Figs. 2 and 3.)

The optimization technique was then employed to obtain design

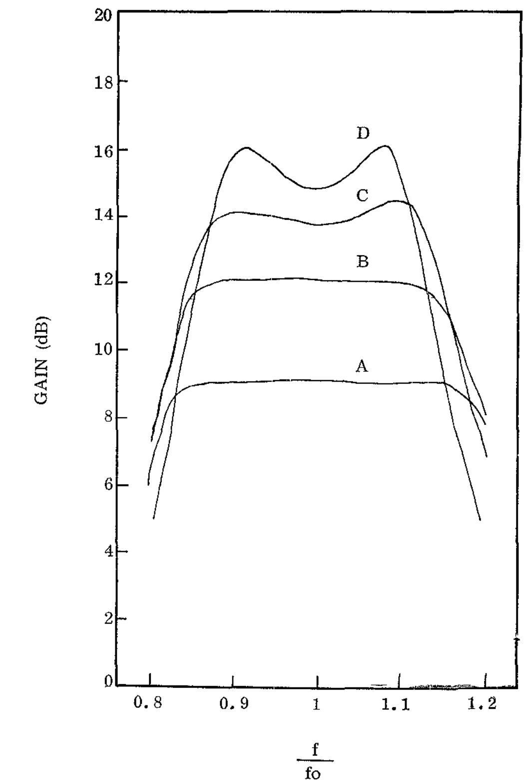

Fig. 2. Amplifier response for 30-percent flat bandwidth.

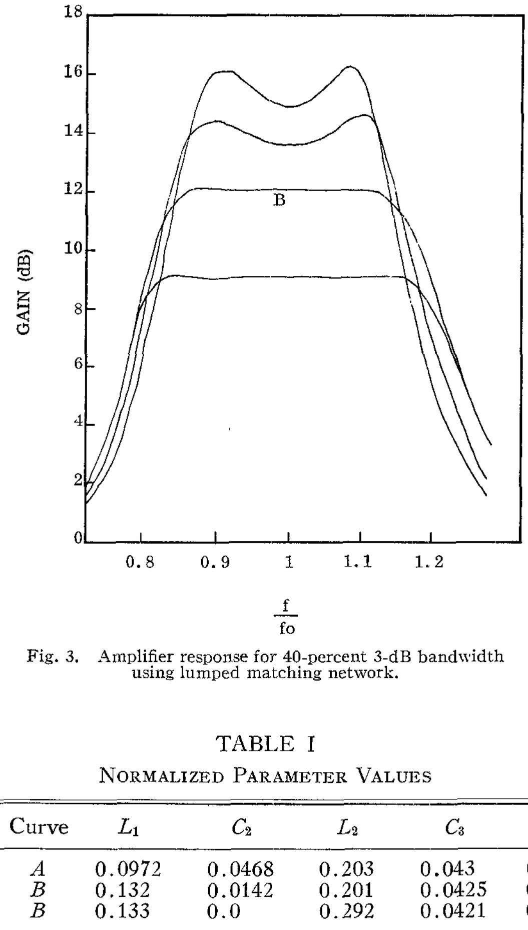

Fig. 3. Amplifier response for 40-percent 3-dB bandwidth using lumped matching network.

TABLE I
NORMALIZED PARAMETER VALUES

Figure	Curve	L_1	C_2	L_2	C_3	L_3
2	*A*	0.0972	0.0468	0.203	0.043	0.548
3	*B*	0.132	0.0142	0.201	0.0425	0.52
4	*B*	0.133	0.0	0.292	0.0421	0.525

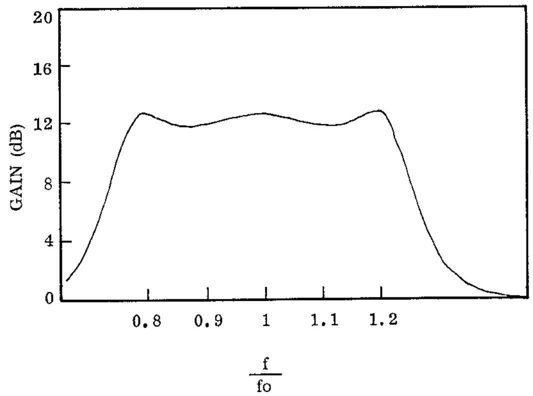

Fig. 4. Amplifier response for 40-percent 3-dB bandwidth using distributed matching network.

TABLE II

NORMALIZED PARAMETER VALUES

$Z01=2$	$l_1=0.0727$
$Z02=0.282$	$l_2=0.478$
$Z03=0.80$	$l_3=0.25$
$Z04=2.2$	$l_4=0.5$
$Z05=1$	$l_5=0.25$

values (see Table II) for 12-dB gain amplifier utilizing the distributed matching network illustrated in Fig. 1(d). The results obtained for a prescribed 3-dB bandwidth of 40 percent (± 0.5-dB ripple) are presented in Fig. 4.

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Loss Calculations for Coupled Transmission-Line Structures

V. K. TRIPATHI

***Abstract*—Expressions are derived for computing loss in coupled TEM structures in terms of complex even- and odd-mode propagation constants and characteristic impedances of the lines. The attenuation due to conductor (series) and dielectric (shunt) losses in a given structure can be determined utilizing these expressions. The results may be particularly useful for computing conductor loss in microwave circuits with a large number of sections used in traveling-wave devices and as microwave circuit elements. The method is applied to some typical structures for computing conductor loss.**

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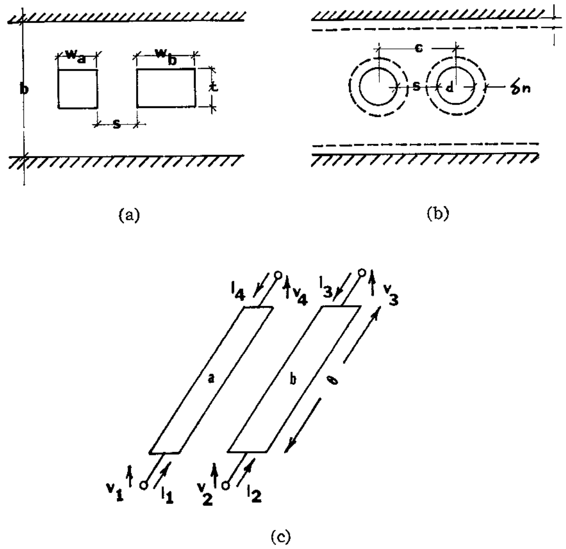
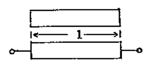
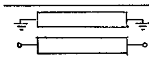
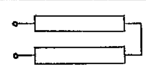
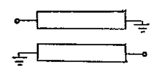
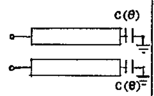


Fig. 1. (a) Rectangular coupled bars between parallel ground planes. (b) Circular coupled bar cross section showing an inward perturbation of all metal surfaces by δn. (c) Schematic of a parallel coupled line four-port.

TABLE I

UNIT SECTIONS	ATTENUATION PER SECTION, NEPERS
 1- ALL PASS	$\alpha p = \dfrac{Z^r_{oe}\,\alpha_e l + Z^r_{oo}\,\alpha_o l}{Z^r_{oe} + Z^r_{oo}}$; $\beta p = \theta$
 2- ALL PASS	$\alpha p = \dfrac{Y^r_{oe}\,\alpha_e l + Y^r_{oo}\,\alpha_o l}{Y^r_{oe} + Y^r_{oo}}$; $\beta p = \theta$
 3- ALL PASS (Meander)	$\alpha p^{*} = \dfrac{\sqrt{Z^r_{oe}\,Z^r_{oo}}\,(\alpha_e l + \alpha_o l)}{Z^r_{oe}\cos^2\theta + Z^r_{oo}\sin^2\theta}$; $\beta p = \mathrm{Cos}^{-1}\left(\dfrac{Z^r_{oe} - Z^r_{oo}\tan^2\theta}{Z^r_{oe} + Z^r_{oo}\tan^2\theta}\right)$
 4- BAND PASS (Interdigital)	$\alpha p^{\dagger} = \dfrac{\left[\sin\theta\,(Y^{r2}_{oo}\,\alpha_o l - Y^{r2}_{oe}\,\alpha_e l) + \left(\dfrac{1+\cos^2\theta}{\sin\theta} \mp 2\,\dfrac{\cos\theta}{\theta}\right) Y^r_{oo}\,Y^r_{oe}\,(\alpha_e l - \alpha_o l)\right]}{\sin\beta p\left[(Y^r_{oo} - Y^r_{oe})^2\right]}$ where, $\beta p = \cos^{-1}\left(\dfrac{Y^r_{oo} + Y^r_{oe}}{Y^r_{oo} - Y^r_{oe}}\cos\theta\right)$
 5- BAND PASS	$\alpha p^{\dagger} = \dfrac{\left[\sin\theta\,(Z^{r2}_{oe}\,\alpha_e l - Z^{r2}_{oo}\,\alpha_o l) + \left(\dfrac{1+\cos^2\theta}{\sin\theta} \pm 2\,\dfrac{\cos\theta}{\theta}\right) Z^r_{oo}\,Z^r_{oe}(\alpha_o l - \alpha_e l)\right]}{\sin\beta p\left[(Z^r_{oe} - Z^r_{oo})^2\right]}$ where, $\beta p = \cos^{-1}\left(\dfrac{Z^r_{oe} + Z^r_{oo}}{Z^r_{oe} - Z^r_{oo}}\cos\theta\right)$
 6- BAND PASS (Comb-Lines)	$\alpha p^{\dagger} = \dfrac{\left[2\,Y^r_{oo}\,Y^r_{oe}\,(\alpha_e l - \alpha_o l)\left(\dfrac{2}{\sin 2\theta} \mp \dfrac{1}{\theta}\right) + \omega C(\theta)\,(Y^r_{oo}\,\alpha_o l - Y^r_{oe}\,\alpha_e l)\left(\dfrac{1}{\cos^2\theta} \mp \dfrac{\tan\theta}{\theta}\right)\right]}{\sin\beta p\left[(Y^r_{oo} - Y^r_{oe})^2\right]}$ where, $\beta p = \cos^{-1}\left(\dfrac{\cot\theta(Y^r_{oo} + Y^r_{oe}) - \omega C(\theta)}{\cot\theta\,(Y^r_{oo} - Y^r_{oe})}\right)$

* This is the same expression as obtained by Kolker [1].
† In these expressions the upper sign is for conductor (series) loss and the lower sign is for dielectric (shunt) loss.