Traveling-Wave Amplifiers with Prescribed Frequency Response

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***Abstract*—This paper develops the design of microwave transistor amplifier combinations that have controlled frequency response over a specified bandwidth. Theoretical analysis of such an amplifier with an arbitrary number of sections is presented. The response of the amplifier is controlled by tapering the frequency selectivity or *Q* of each section of the amplifier. To verify the theory, a three-section amplifier with maximally-flat time delay response is designed, constructed, and evaluated. Existing traveling-wave amplifiers are modeled as lossy transmission lines. Although the amplifiers are relatively broadband, a prescribed frequency response is not achieved, and each transistor does not receive an equal portion of the signal power. Resistive elements are required for impedance matching at the input and output. The amplifier design developed in this work seeks to improve on existing techniques by trading bandwidth for controlled gain. By making the transmission line that connects the amplifier sections non-uniform, the frequency response is controlled over the design bandwidth. The impedance of the transmission lines is specified so that all of the input power is delivered to each transistor equally. No resistive elements are required for impedance matching, thus reducing the inherent noise of the amplifier. The designs are easily implemented using familiar components.**

I. Introduction

Many researchers have used traveling-wave power dividers and combiners [1] in microwave transistor amplifiers, situating the transistors at intervals along a uniform transmission line. In this configuration, the amplifier is modeled as a lossy transmission line, with the input and output impedances of the transistor as the lossy elements. No attempt is made to match the input or output impedance of the transistors. The first transistor receives most of the input power, and so limits performance. The last transistor has a low signal to noise ratio, so the overall noise performance of the amplifier is decreased. The power combining network on the output of the amplifier reflects power back into the outputs of the transistors. There are resistive elements terminating the input and output lines to provide impedance matching. These elements diminish the available power and add noise.

Despite these limitations, traveling-wave amplifiers do have some advantages over other microwave amplifiers. They are relatively broad-band, although the frequency response is not controlled. Since the signal reaches each amplifier at a different phase, the instantaneous power required from the signal source is diminished somewhat. There may also be some noise cancellation as a result of the phase difference of the signal [2]–[9]. A diagram of this amplifier configuration is shown in Fig. 1.

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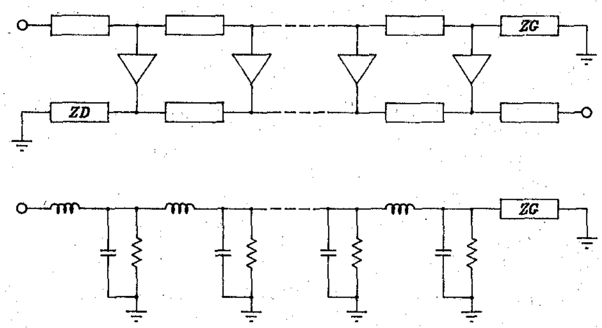

Fig. 1. The traveling-wave amplifier.

By making the transmission line in the traveling-wave power divider and combiner nonuniform, the frequency response of the magnitude or phase of the network can be controlled. At the same time, the impedance of the network can be specified so that the output power is delivered to each section of the amplifier equally, and each section produces an equal portion of the output power delivered to the load. Reflections in the output power combiner are minimized, because the connecting transmission lines show a lower impedance than the inputs to the transistors. No resistive elements are required for impedance matching, thus increasing the available power and reducing the inherent noise of the amplifier. Since the sections of the amplifier are separated by quarter-wavelength transformers, any noise that is common to all sections will be subject to some cancellation.

This work investigates the properties of these multi-section microwave transistor amplifiers that have a prescribed frequency response. The frequency response of the amplifier is specified by tapering the Q, or frequency selectivity, of each section of the amplifier as is done in multi-section microwave filters. Filter theory has been used to design broad-band matching networks for microwave amplifiers [10], but in general has not previously been incorporated into the overall amplifier design. A three-section amplifier with maximally-flat time delay (Thomson or linear-phase) response will be analyzed, designed, constructed, and evaluated.

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The design techniques presented in this paper represent a departure from existing techniques, in that the amplifiers share desirable characteristics of both series cascade and parallel amplifiers. Henceforth in this work, this distributed amplifier configuration will be called the *parallel-cascade* amplifier.

II. Theory

Fig. 2 shows a symbolic diagram of the complete microwave circuit of a K-section parallel-cascade amplifier. Power supply and gate bias connections are omitted. Fig. 3 shows a photograph of the circuit realized in microstrip with SMA input and output connectors, including power and bias lines. Each section of the amplifier consists of a quarter-wavelength transformer of characteristic impedance Z_{0k}. At the outputs of the transformers are matching networks leading to the inputs of the transistors. The matching networks on the transistors transform the generally complex impedance of the transistor to a purely real value R_k at the design frequency ω_0 (radians per second). The outputs of the transistors also have matching networks leading to quarter-wavelength transformers at the output of the amplifier.

For analysis, the amplifier is divided across the centerline, with the input and output resistances of the matching networks for the transistors connected to a virtual ground. Each section is now modeled as a quarter-wavelength transformer with a resistive load, as shown in Fig. 4. The impedances at various points in this network are calculated using the transmission line equation [11] and basic rules for combining impedances.

The first design constraint is that each section must receive equal power. For an arbitrary section k in an amplifier with K sections, R_k will absorb a given amount of power. $(K - k)$ times as much power must flow into the remaining sections. Therefore, the source impedance of section $(k + 1)$ must be $R_k/(K - k)$. This impedance will combine in parallel with R_k such that the load on section k is

$$R_{Lk} = R_k/(K - (k - 1)). \tag{1}$$

To maintain the equal power constraint, the input impedance to the section must be the same as the source impedance

$$Z_{Sk} = R_{(k-1)}/(K - (k - 1)). \tag{2}$$

The line must match the input to the load. Using the transmission line equation,

$$Z_{0k} = \sqrt{\left(\frac{R_k}{(K - (k - 1))}\right)\left(\frac{R_{(k-1)}}{(K - (k - 1))}\right)} \tag{3}$$

or

$$Z_{0k} = \sqrt{\frac{R_k R_{(k-1)}}{(K - (k - 1))^2}}. \tag{4}$$

The second design constraint is that all of the input power must be transferred to the resistors with no loss, so the network must be matched to the characteristic impedance of the input transmission line. In other words, $Z_{IN} = Z_0$. Since there are K resistors in the network, the load on the first section will be R_1/K. This must match to Z_0 through Z_{01}. Therefore,

$$Z_{01} = \sqrt{\frac{Z_0 R_1}{K}}. \tag{5}$$

Each section of the amplifier will have a Q_k, or frequency selectivity, that has a specific numerical relation-

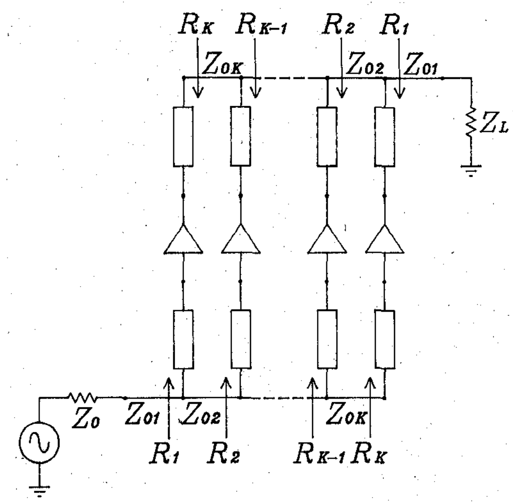

Fig. 2. Complete microwave circuit of a K-section parallel-cascade amplifier.

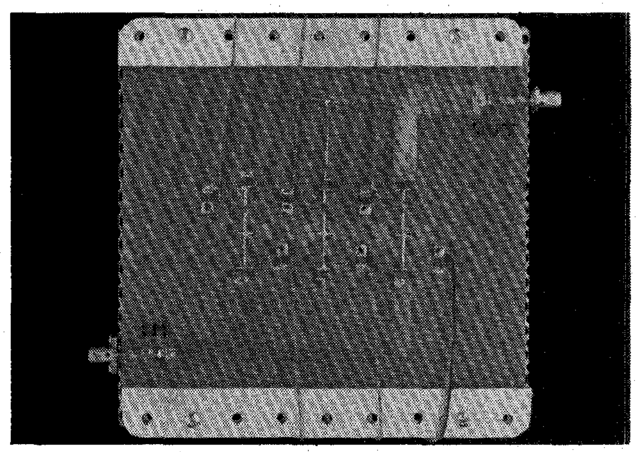

Fig. 3. Photograph of the circuit realized in microstrip with SMA input and output connectors. Power supply and gate bias leads shown.

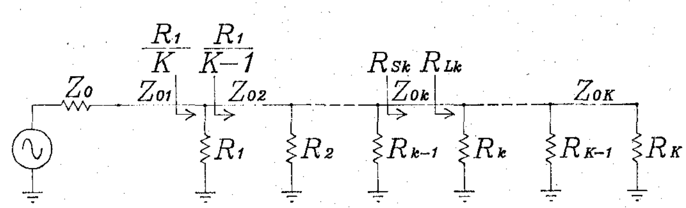

Fig. 4. Analytical model of parallel-cascade amplifier for impedance analysis.

ship to the Q of the other sections. The expression for the frequency selectivity of the matched quarter-wavelength transformer [12] is

$$Q = \frac{\pi}{8}\left|\frac{Z_{0t}}{R_L} - \frac{R_L}{Z_{0t}}\right|. \tag{6}$$

To determine numerical values for each R_k and Z_{0k}, the desired frequency response and the required total frequency selectivity Q_T of the amplifier must be determined. Q_T is defined as

$$Q_T = \frac{\omega_0}{(\omega_2 - \omega_1)} \tag{7}$$

where ω_0 is the design frequency in radians per second, and $(\omega_2 - \omega_1)$ is the 3 dB bandwidth.

A common requirement for the frequency response of an amplifier is maximally-flat time delay (Thomson or linear-phase response). Circuits having this type of response can be built with an arbitrary number of frequency-selective stages. The component values of each stage are computed from lowpass prototype filter tables [13] that give values in terms of a prototype component value g_k, shown in Table I. From these g_k, the Q of each section in the amplifier is determined by the equation

$$Q_k = \frac{g_k}{2} Q_T \tag{8}$$

Rewriting (6) for an arbitrary section of the circuit,

$$Q_k = \frac{\pi}{8}\left|\frac{Z_{0k}}{R_{Lk}} - \frac{R_{Lk}}{Z_{0k}}\right|. \tag{9}$$

The characteristic impedance of the line, Z_{0k}, is determined by the design constraints given earlier in this section. It must match the source impedance to the load on that particular section. When this condition is met,

$$Z_{0k} = \sqrt{Z_{Sk} R_{Lk}} \tag{10}$$

and

$$\frac{8Q_k}{\pi} = \left|\sqrt{\frac{Z_{Sk}}{R_{Lk}}} - \sqrt{\frac{R_{Lk}}{Z_{Sk}}}\right|. \tag{11}$$

Squaring both sides, multiplying through first by Z_{Sk}, and then by R_{Lk} gives

$$Z_{Sk}^2 - \left[2 + \left(\frac{8Q_k}{\pi}\right)^2\right] Z_{Sk} R_{Lk} + R_{Lk}^2 = 0 \tag{12}$$

a quadratic equation in R_{Lk}. Two roots are calculated using the quadratic formula. The value of R_{Lk} that gives the most reasonable value for Z_{0k} (usually between 15 Ω and 150 Ω) is used.

The first section has $Z_{Sk} = Z_0$ and $R_{Lk} = R_1/K$, yielding

$$R_1^2 - \left[2 + \left(\frac{8Q_1}{\pi}\right)^2\right] KZ_0 R_1 + (KZ_0)^2 = 0 \tag{13}$$

TABLE I
NORMALIZED ELEMENT VALUES (g_k) FOR MAXIMALLY FLAT DELAY (THOMSON OR LINEAR PHASE) RESPONSE AMPLIFIER HAVING n SECTIONS

n	g_1	g_2	g_3	g_4	g_5	g_6	g_7
2	0.5755	2.1478					
3	0.3374	0.9705	2.2034				
4	0.2334	0.6725	1.0815	2.2404			
5	0.1743	0.5072	0.8040	1.1110	2.2582		
6	0.1365	0.4002	0.6392	0.8538	1.1126	2.2645	
7	0.1106	0.3259	0.5249	0.7020	0.8690	1.1052	2.2659

Thus, Q_1, K, and the characteristic impedance Z_0 completely determine R_1.

For an arbitrary section k, the source impedance Z_{Sk} and the load resistance R_{Lk} are given by (2) and (1), respectively. Substituting these expressions into (12) yields

$$\left(\frac{R_{k-1}}{(K-(k-1))}\right)^2 - \left[2 + \left(\frac{8Q_k}{\pi}\right)^2\right] \frac{R_{k-1}R_k}{(K-(k-1))^2} + \left(\frac{R_k}{(K-(k-1))}\right)^2 = 0 \tag{14}$$

or

$$R_k^2 - \left[2 + \left(\frac{8Q_k}{\pi}\right)^2\right] R_{k-1}R_k + R_{k+1}^2 = 0. \tag{15}$$

Given R_{k-1} and Q_k, pick the value for R_k that gives a reasonable value for Z_{0k}. Once R_1 is determined, (15) is used with Q_2 to determine R_2; then R_2 and Q_3 determine R_3, and so on until the resistor values for all K sections are calculated.

For a circuit with many sections, K will be a large value. The load on the first section is then a small resistance R_1/K. Depending on Q_1, Z_{01} will have to be small to give a match at the input. If the power is to be divided equally between a large number of resistors, each individual resistor must be large to produce a reasonable value for the input impedance. Then, the last section of transmission line will have a high value for Z_{0K} to match to R_K.

In microstrip circuits, low impedance lines are fairly wide and high impedance lines are fairly narrow. In practical circuits, it is difficult to join transmission lines of different widths at a common point. The range of widths leads to difficult layout and production of the circuit, and less predictability in its behavior. Fringing capacitance at line junctions changes the effective electrical length of the lines. Dispersive effects are non-negligible at lower frequency in lines with extremely low impedance. Practical circuits are effectively limited to two or three sections.

The equations for the three-section amplifier are shown in the following section. These equations will be used to design a parallel-cascade amplifier with $Q_T = 2.0$, built to operate at $f = 2$ GHz with a maximally-flat phase response in the designed passband. This amplifier will be constructed, and its measured performance will be compared with its simulated performance in the Results section.

III. Three-Section Amplifier

Fig. 5 shows the equivalent circuit of a three-section parallel-cascade amplifier. Looking away from the source,

$$Z_A = \frac{Z_{03}^2}{R_3} \tag{16}$$

$$Z_B = R_2 \| Z_A \tag{17}$$

$$Z_C = \frac{Z_{02}^2}{Z_B} \tag{18}$$

$$Z_D = R_1 \| Z_C \tag{19}$$

$$Z_{IN} = \frac{Z_{01}^2}{Z_D}. \tag{20}$$

For an impedance match at the input, $Z_{IN} = Z_0$. For equal power to each section, $Z_A = R_2$ and $Z_C = R_1/2$. Then $Z_B = R_2/2$ and $Z_D = R_1/3$.

Using (5) and (3),

$$Z_{01} = \sqrt{Z_{IN} Z_D} = \sqrt{\frac{Z_0 R_1}{3}} \tag{21}$$

$$Z_{02} = \sqrt{Z_C Z_B} = \sqrt{\frac{R_1 R_2}{4}} \tag{22}$$

$$Z_{03} = \sqrt{Z_A R_3} = \sqrt{R_2 R_3}. \tag{23}$$

Now

$$\begin{aligned} &Z_{S1} = Z_0 && Z_{S2} = \tfrac{1}{2}R_1 && Z_{S3} = R_2 \\ &R_{L1} = \tfrac{1}{3}R_1 && R_{L2} = \tfrac{1}{2}R_2 && R_{L3} = R_3 \\ &Z_{01} = \sqrt{\tfrac{1}{3}Z_0 R_1} && Z_{02} = \sqrt{\tfrac{1}{4}R_1 R_2} && Z_{03} = \sqrt{R_2 R_3} \end{aligned} \tag{24}$$

in accordance with the general equations given in Section II.

Substituting appropriate values for each section into (12),

$$\begin{aligned} R_{L1}^2 - \left[2 + \left(\frac{8Q_1}{\pi}\right)^2\right] Z_{S1} R_{L1} + Z_{S1}^2 &= 0 \\ (R_1/3)^2 - \left[2 + \left(\frac{8Q_1}{\pi}\right)^2\right] Z_0 (R_1/3) + Z_0^2 &= 0 \\ R_1^2 - \left[2 + \left(\frac{8Q_1}{\pi}\right)^2\right] 3Z_0 R_1 + (3Z_0)^2 &= 0 \end{aligned} \tag{25}$$

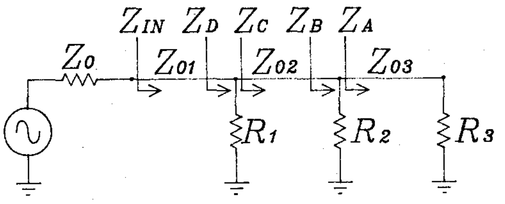

Fig. 5. Equivalent circuit of a three-section parallel-cascade amplifier.

TABLE II
Component Values for 3-Section Maximally Flat Delay (Thomson or Linear Phase) Response Amplifier*

k	1	2	3
Q_k	0.3374	0.9705	2.2034
Z_{0k}	32.94	91.96	89.84
R_k	65.10	519.64	15.53

*k is the section number of the amplifier. Q_k is the frequency selectivity, Z_{0k} is the characteristic impedance, and R_k is the shunt resistance at the output of the kth section of transmission line.

$$\begin{aligned} R_{L2}^2 - \left[2 + \left(\frac{8Q_2}{\pi}\right)^2\right] Z_{S2} R_{L2} + Z_{S2}^2 &= 0 \\ (R_2/2)^2 - \left[2 + \left(\frac{8Q_2}{\pi}\right)^2\right] (R_1/2)(R_2/2) + (R_1/2)^2 &= 0 \\ R_2^2 - \left[2 + \left(\frac{8Q_2}{\pi}\right)^2\right] R_1 R_2 + R_1^2 &= 0 \end{aligned} \tag{26}$$

$$\begin{aligned} R_{L3}^2 - \left[2 + \left(\frac{8Q_3}{\pi}\right)^2\right] Z_{S3} R_{L3} + Z_{S3}^2 &= 0 \\ R_3^2 - \left[2 + \left(\frac{8Q_3}{\pi}\right)^2\right] R_2 R_3 + R_2^2 &= 0 \end{aligned} \tag{27}$$

Q_1 and the characteristic impedance Z_0 completely determine R_1, which then, with Q_2, determines R_2. Q_3 and R_2 give R_3.

Using these equations to design an amplifier with a Thomson response, operating at $f = 2$ GHz with $Q_T = 2.0$, gives the component values shown in Table II. To verify that simulations and practical circuit measurements agree, the amplifier was designed around the NE72084, a low-cost, general-purpose GaAs MESFET manufactured by NEC [14, pp. 3-117–3-123]. The input and output impedances of the transistor were conjugately matched to their respective R_k values using series reactance elements and quarter-wavelength transformers.

IV. Results

The circuit designed in the previous section was constructed on 1/32 in thick polyethylene circuit board with

$\epsilon_r = 2.32$, and 1 oz copper cladding on both sides. The circuit layout was etched on one side of the board. The copper cladding on the other side of the board was left as a ground plane.

The input reflection coefficient s_{11}, the forward transmission coefficient s_{21}, the output reflection coefficient s_{22}, and the reverse transmission coefficient s_{12} were obtained in two ways. The circuit was simulated using the SPICE [15]–[17] circuit analysis program. Since SPICE does not provide a device model for the MESFET, a simple unilateral MESFET transistor equivalent circuit ([18, p. 35]) was used. The S-parameters of the circuit were extracted from the simulations using the techniques in [19]. The S-parameters of the practical circuit were measured on a Hewlett-Packard HP8753A network analyzer.

The component values for the MESFET equivalent circuit were calculated from the input and output impedances of the transistor at the design frequency, using S-parameter data provided by the manufacturer. If the transistor is unconditionally stable at the design frequency, a simultaneous solution for these impedances can be found that takes into account the bilateral nature of the transistor. The choice of design frequency for the circuit considered in this paper was constrained by the measurement range of the network analyzer. The transistor was required to operate in a frequency range where it is potentially unstable, and so the input and output impedances were calculated with the necessary assumption that the transistor is unilateral.

The matching networks at the input and output of the transistors were designed based on the calculated input and output impedances. Any discrepancy between the calculated impedances and the impedances of the physical transistor will produce a shift in the frequency where the best match occurs. This shift in the frequency of the individual matching networks will be reflected in the overall response of the amplifier.

Measured and simulated data was compared using the Graph Tool scientific graphing package [20]. In most cases, the magnitude of the S-parameter is of main interest. The phase of the S-parameter is shown for the forward transmission coefficient of the Thomson amplifier, as linear phase is a primary design goal. In all figures, measured data is shown by a solid line, and simulated data is shown as a dashed line.

The measured curve for the magnitude of the input reflection coefficient in Fig. 6 follows the general shape of the simulated curve. The center frequency is offset to approximately 2.1 GHz.

The magnitude of the forward transmission coefficient in Fig. 7 is surprisingly flat in the simulation, considering that the primary design criterion is linear phase. The measured data has about the same bandwidth, with reduced gain in the passband. This reduced gain occurs because the combining network is reflecting some power back to the output of the transistors. This reflected power does not contribute to the forward gain of the amplifier.

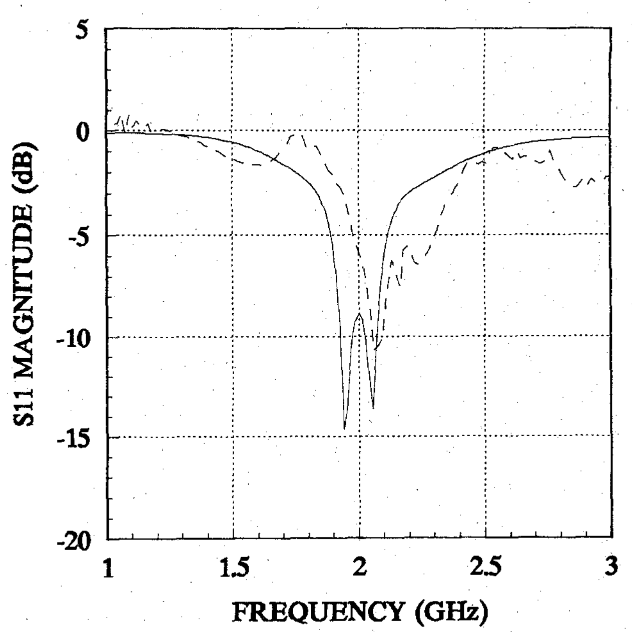

Fig. 6. Input reflection coefficient magnitude ($|s_{11}|$) for the Thomson amplifier with $Q_T = 2.0$. Measured data shown as a solid line; simulation results shown as a dotted line.

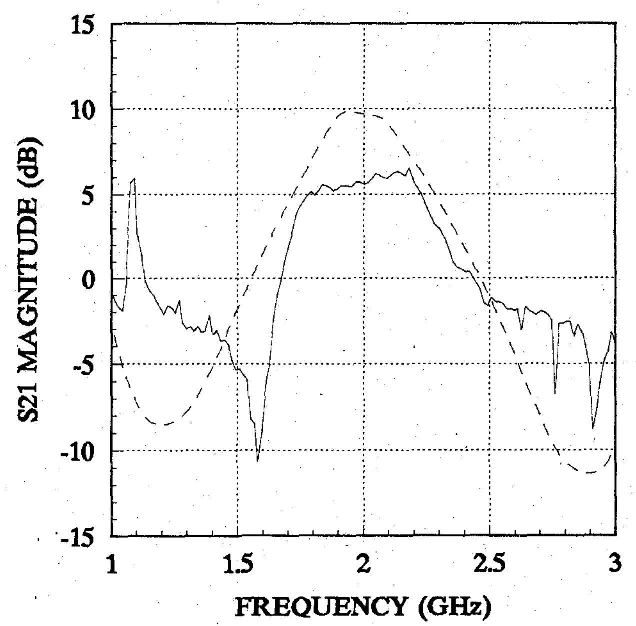

Fig. 7. Magnitude of forward transmission coefficient ($|s_{21}|$) of the Thomson amplifier with $Q_T = 2.0$. Measured data shown as a solid line; simulation results shown as a dotted line.

The forward transmission coefficient in Fig. 8 shows excellent phase response; it is quite linear in the passband for both the measured and simulated curves. The measured curve is again offset in frequency for this parameter.

The simulated curve for the output reflection coefficient in Fig. 9 shows a good impedance match at the design frequency. The measured data shows a fairly good match at about 2.4 GHz, but outside of the passband the measured and simulated values do not agree well. This is

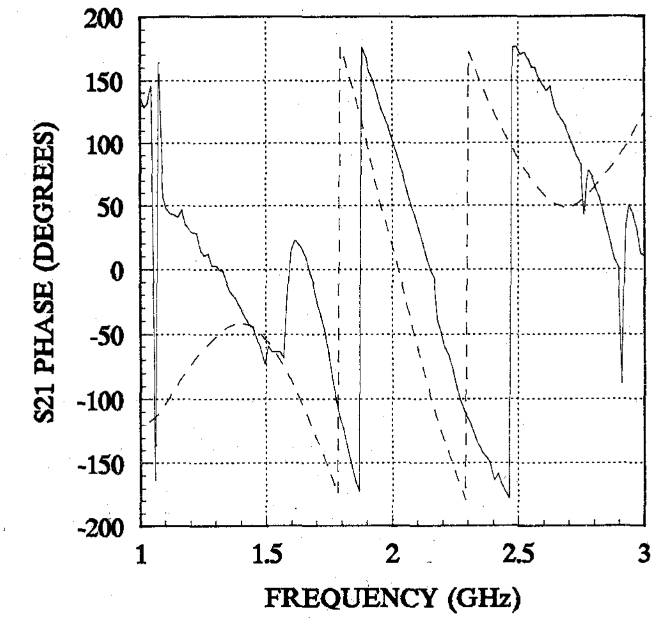

Fig. 8. Phase of forward transmission coefficient ($\angle s_{21}$) of the Thomson amplifier with $Q_T = 2.0$. Measured data shown as a solid line; simulation results shown as a dotted line.

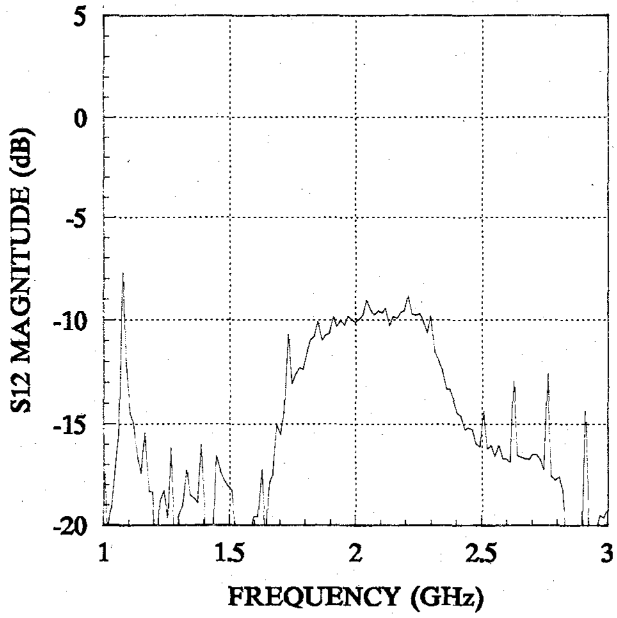

Fig. 10. Reverse transmission coefficient magnitude ($|s_{12}|$) for the Thomson amplifier with $Q_T = 2.0$.

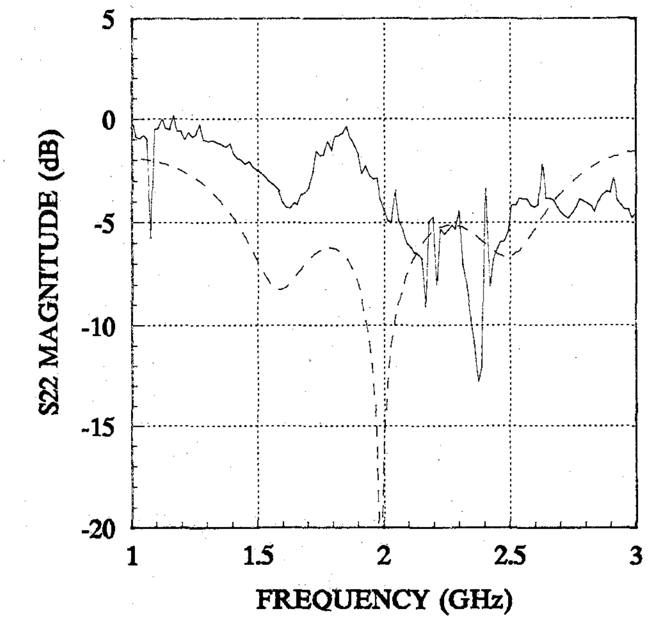

Fig. 9. Output reflection coefficient magnitude ($|s_{22}|$) for the Thomson amplifier with $Q_T = 2.0$. Measured data shown as a solid line; simulation results shown as a dotted line.

largely because the transistor model is unilateral, whereas the transistor itself is bilateral. The reverse gain of the transistor affects the output impedance match much more strongly than the input impedance match.

The measured reverse transmission coefficient is shown in Fig. 10. The simulation has no transmission in the reverse direction, because the transistor model is unilateral. Reflected power at the amplifier output is transmitted back to the input within the designed passband.

V. Conclusion

This work has shown that the parallel-cascade microwave amplifier with prescribed frequency response using traveling-wave power dividers and combiners produces improved performance over a specified bandwidth. A mathematical treatment of the underlying theory has aided in the development of effective design techniques that insure design goals can be met. In spite of the fact that an approximate equivalent circuit was used to model the transistor, practical circuits built using modern GaAs MESFET microwave transistors have shown that measurements agree reasonably well with prediction.

The theory can be generalized to K-sections, but constraints on microstrip line widths limit the number of sections in practice. Experimental results agree well with simulations for the three-section amplifier. The Thomson amplifier shows linear phase response over the designed passband. Since maximally-flat magnitude is not the primary requirement for linear phase circuits, performance can be considered satisfactory. Further research into extended applications of these design techniques is warranted. Monolithic implementations on GaAs substrates could prove particularly interesting for communications applications.

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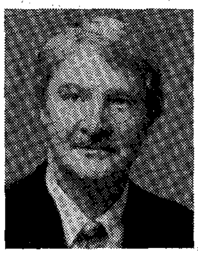
From 1982 to 1987, he was a Research Electronics Technician for the Department of Psychiatry at Duke University Medical Center, where he participated in medical research and product development. He is presently holding a postdoctoral position with the Microelectronics Center of North Carolina, where he is engaged in research on high-frequency amplification using microvacuum transistors. Dr. Palmer is a member of Sigma Xi.



William T. Joines (M'61) was born in Granite Falls, NC, on November 20, 1931. He received the B.S.E.E. degree (with high honors) from North Carolina State University, Raleigh, in 1959, and the M.S. and Ph.D. degrees in electrical engineering from Duke University, Durham, NC, in 1961 and 1964, respectively.

From 1959 to 1966, he was a member of the Technical Staff at Bell Laboratories, Winston-Salem, NC, where he was engaged in research and development work on microwave components and systems for military applications. He joined the faculty of Duke University in 1966, and is currently a Professor of Electrical Engineering. His research and teaching interests are in the area of electromagnetic wave interactions with materials.