preparation does not seem to be an adequate solution, since gaps less than 1 percent cause significant errors in the determination of the dielectric constant.

The modal-analysis method presented here is very complicated, and it does not provide a reasonable method for analyzing waveguide dielectric measurements to determine sample permittivity. It is hoped, however, that this analysis has shown that even very small air gaps must be taken seriously in any experiment of the type presented here.

Simple perturbation formulas that relate the "measured" dielectric constant to the "actual" dielectric constant often produce worse results than just accepting the measured value as the true value. The best way of correcting for the air gap, it would seem, is to fill the gap with a conducting paste. The resulting measurements produce inferred dielectric constants that are very close to the true values so long as the gap is small.

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Large-Signal Time-Domain Simulation of HEMT Mixers

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***Abstract*—A large-signal HEMT model and a time-domain nonlinear circuit analysis program have been developed. In this work a systematic method to simulate HEMT mixers and design them for maximum conversion gain is presented. The transconductance-compression effect reduces the mixer's conversion gain at high frequencies. Simulation results from mixers designed to operate at 10, 20, and 40 GHz show that a reduction in parasitic conduction in the AlGaAs layer significantly increases the conversion gain.**

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I. INTRODUCTION

The high electron mobility transistor (HEMT) is superior to the GaAs MESFET when used in low-noise microwave amplifiers. Recently the power performance of HEMT's with a single heterojunction [1] and double heterojunctions [2] have been investigated. Maas reported a HEMT mixer that operates at 45 GHz [3]. These promising results indicate that the HEMT will no longer be confined to small-signal applications.

For both the small-signal and the large-signal application of GaAs FET's, computer-aided design (CAD) is very valuable in the design process, especially in the design of monolithic integrated circuits, where tuning the circuits to optimize performance is impractical. The accuracy and efficiency of computer aided design and simulation rely on a well-developed device model. In this paper we present a large-signal HEMT model and a systematic CAD method in the time domain for the design and simulation of HEMT mixers. The effect on mixer performance by transconductance compression, which is a distinct feature of the HEMT [4], has also been studied.

II. A LARGE-SIGNAL MODEL OF THE HEMT

Under large-signal operation, the element values of the HEMT equivalent circuit vary with time and become dependent on the terminal voltages. A large-signal model can be derived by considering the main nonlinear elements of the equivalent circuit. As shown in Fig. 1, the elements of the large-signal HEMT model assumed to be nonlinear in this work are the gate-to-source capacitance, C_{gs}, and the drain current source, I_d. Other circuit elements are assumed to be linear.

The drain current source is represented by the current–voltage equations derived in [5]. From the experimental results, C_{gs} of a HEMT behaves quite similarly to that of a MESFET [6], [7], because considerable charges in addition to the two-dimensional electrons are modulated by the terminal voltages. Hence a valid approach is to use a Schottky diode equation and

$$C_{gs}(V_g) = \frac{C_{gs0}}{\left(1 - \frac{V_g}{V_{bi}}\right)^m} \tag{1}$$

where C_{gs0} and m are model parameters adjusted to fit the measured values. V_{bi} is the built-in voltage for the Schottky gate and V_g is the internal gate voltage.

Compared to the large-signal MESFET models given in [8] and [9], two diodes representing forward gate conductive current and gate–drain breakdown current are neglected in our HEMT model; hence this model is valid only when these two currents are absent in device operation, e.g., in a mixer. The nonlinearity of gate-to-drain capacitance, C_{dg}, is also neglected [7].

A GE HEMT with 0.25 μm gate length and 150 μm gate width was chosen [1]. Bias-dependent S parameters and dc characteristics were used to construct the large-signal model. Small-signal equivalent circuits were extracted from the measured S parameters at two bias points: $V_{GS} = -0.2$ V, $V_{DS} = 2.0$ V and $V_{GS} = -0.5$ V, $V_{DS} = 2.0$ V. C_{gs0} and m in (1) were determined by a

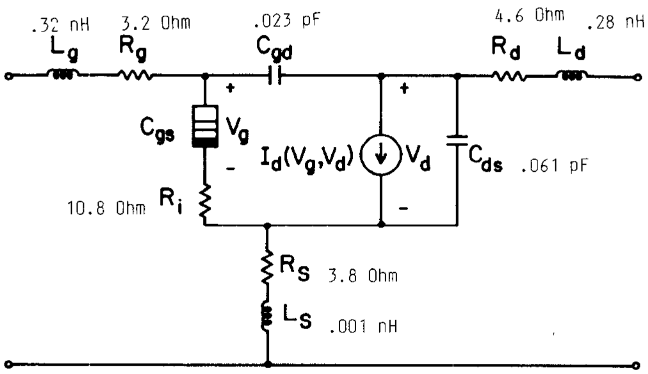

Fig. 1. Large-signal model of the HEMT with values of the linear elements.

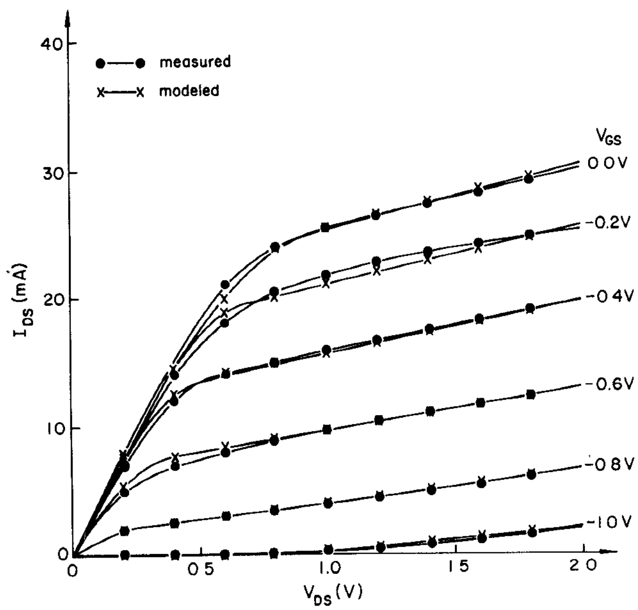

Fig. 2. Measured and modeled DC $I-V$ characteristics of the HEMT.

simple optimization program to fit (1) to the values of C_{gs} in the bias-dependent small-signal equivalent circuits. For $C_{gs0} = 0.178$ pF and $m = 0.67$, the model gives a close fit; therefore

$$C_{gs}(V_g) = \frac{0.178}{\left(1 - \dfrac{V_g}{0.8}\right)^{0.67}} \quad (\text{pF}). \tag{2}$$

Use was made of dc current–voltage characteristics to determine the parameters in the expression for $I_d(V_g, V_d)$. The modeled and measured $I-V$ characteristics are shown in Fig. 2. The modeling procedure and model parameters have been discussed in [5].

The values of the linear elements in the large-signal HEMT model are shown in Fig. 1. The S parameters calculated from the large-signal model are compared with the measured S parameters. The r.m.s. errors of the modeled and measured S parameters are 6.8 percent and 7.2 percent for the bias at $V_{GS} = -0.2$ V and $V_{GS} = -0.5$ V, respectively. These results are comparable to those obtained from using small-signal equivalent circuits to model the S parameters.

III. Design and Simulation

Owing to the nonlinear nature of the problem, to design a mixer with maximum conversion gain, a systematic computer-aided design method proves to be very helpful. A time-domain

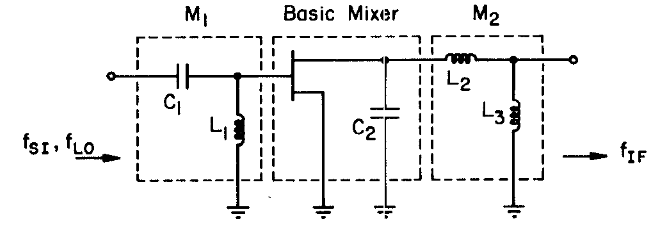

Fig. 3 Circuit diagram of the designed mixer.

TABLE I
Element Values of the Mixer Matching Networks.

Freq.	C_1(pF)	C_2(pF)	L_1(nH)	L_2(nH)	L_3(nH)
10	0.08	2.2	1.23	1.89	0.91
20	0.08	2.2	0.33	1.84	0.96
40	0.1	2.2	0.1	1.99	0.87

nonlinear circuit analysis program, CADNON [10], suitable for microwave FET circuit analysis, has been developed. This program utilizes Aprille and Trick's shooting algorithm [11] to reduce the convergence time for a nonlinear circuit to reach steady state. With this program, we can simulate the input and output impedances of the device at the desired radio frequency (RF) and intermediate frequency (IF) under large-signal local oscillator (LO) condition. The maximum conversion gain can then be obtained by simultaneous adjustment of the matching networks for both impedance matching and optimum load conditions. The RF power level is set at -30 dBm and the bias is chosen at $V_{GS} = -0.8$ V and $V_{DS} = 2.0$ V. Fig. 3 shows the circuit diagram of the mixers, and the element values of the matching networks are summarized in Table I for each operating frequency. The IF is 2 GHz in all cases.

For the mixer designed to operate at $f_{RF} = 10$ GHz and $f_{LO} = 8$ GHz, the optimal LO power is 2 dBm for a maximum conversion gain of 9 dB (Fig. 4). This result compares favorably with the maximum conversion gain on the order of 6 dB for MESFET mixers at X-band [12]–[14]. The input VSWR at 10 GHz and the output VSWR at 2 GHz are 1.1 at maximum conversion gain. These low VSWR's ensure that the mixer is optimized for conversion gain.

The conversion gain versus LO power of the mixer operating at $f_{RF} = 20$ GHz and $f_{LO} = 18$ GHz is also shown in Fig. 4. A maximum conversion gain of 3.6 dB with an input VSWR of 1.05 and an output VSWR of 1.1 is achieved at an LO power of -0.46 dBm.

To investigate the feasibility of millimeter-wave mixers using HEMT's, a HEMT mixer is designed and simulated at 40 GHz. Instead of conversion gain, the mixer shows a minimum conversion loss of 2 dB at LO power equal to 2 dBm, with input and output VSWR's of 1.1 and 1.15. The low VSWR's imply that we can hardly obtain more conversion gain from the HEMT mixer.

Conversion gain versus gate bias voltage of the 20 GHz mixer is shown in Fig. 5. The conversion gain shows a minimum at $V_{GS} = -0.4$ V, which is close to the bias point for maximum dc transconductance. This anomalous phenomenon has never been observed in MESFET mixers and is caused by transconductance compression.

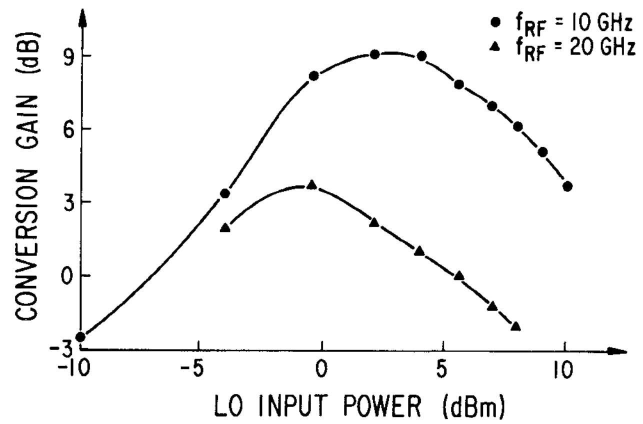

Fig. 4. Conversion gain versus LO power of the 10 GHz mixer and 20 GHz mixer.

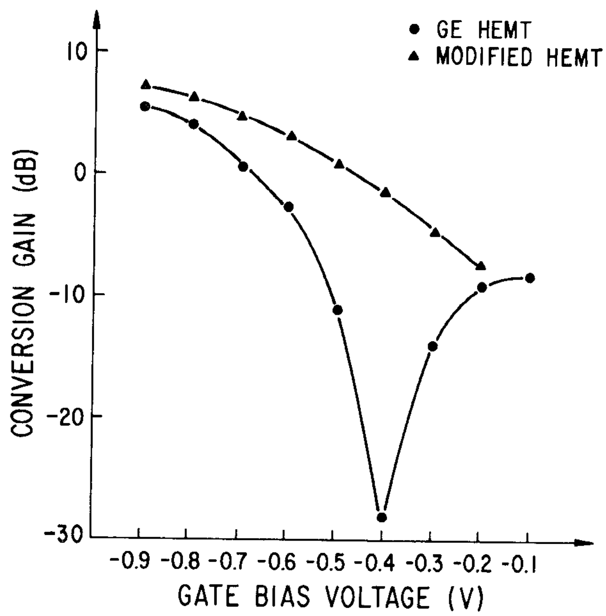

Fig 5 Calculated conversion gain versus gate bias of the 20 GHz mixer with the HEMT and the modified HEMT.

IV. Effects of Transconductance Compression on Mixer Performance

A distinct feature of the HEMT as compared to the GaAs MESFET is the strong transconductance compression frequently observed in the normal bias range, because of the parasitic conduction in the AlGaAs layer [4].

Fig. 6 shows the calculated g_m versus V_{GS} of the HEMT. Transconductance compression is observed when the gate biasing voltage exceeds -0.5 V. Since g_m is nearly symmetrical with V_{GS}, when the gate is biased at -0.4 V, the odd components of the Fourier series expansion of the $g_m(t)$ will be reduced. The conversion gain of the HEMT mixer decreases accordingly.

To study the transconductance compression effects, the same HEMT is used in the 20 GHz mixer circuit. But the parasitic conduction is removed from the $I_d(V_g, V_d)$ model by setting the threshold voltage for parasitic conduction much larger than the maximum gate voltage during mixer operation. The conversion gain as a function of gate bias is shown in Fig. 5 for this modified HEMT and all the other operating conditions unchanged. The dip in the original conversion gain characteristics shown in Fig. 5 disappears, and the maximum conversion gain increases by 2.5 dB. With the modified HEMT in the 40 GHz mixer, an increase of the maximum conversion gain from -2 dB to 0.75 dB is observed. This suggests that a mixer for millimeter-wave application, which provides conversion gain instead of conversion loss, is possible by using properly designed HEMT's. This result also agrees with the experimental result by Maas, who obtained 1.5 dB conversion gain in a 45 GHz HEMT mixer [3].

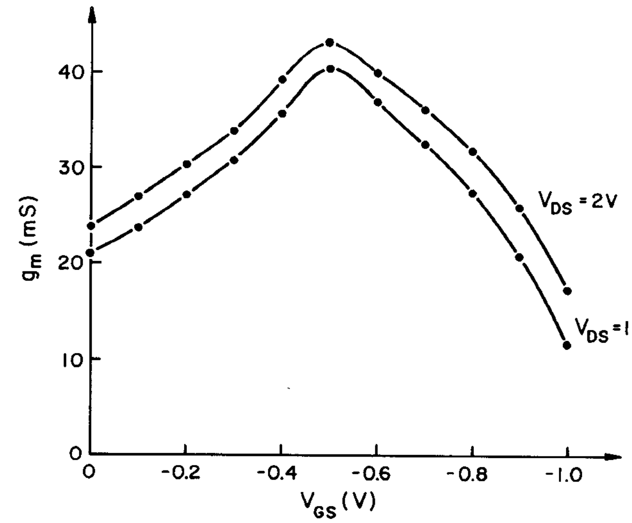

Fig. 6. Calculated transconductance versus gate bias of the HEMT

V. Conclusions

HEMT mixers have been designed and analyzed effectively in the time domain with a CAD technique. The HEMT shows potential for application as a millimeter-wave mixer which offers conversion gain. Transconductance compression is found to degrade the performance of the HEMT mixer. In addition, to avoid the reduction of conversion gain, the gate bias has to be properly chosen for HEMT's that exhibit transconductance compression.

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Design of Combline and Interdigital Filters with Tapped-Line Input

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Abstract **—Explicit design equations for combline and interdigital filters with tapped-line inputs are presented. The equations are based upon a new equivalent circuit for a tapped-line input filter, derived from the open-wire-line equivalent circuit given by Cristal [1]. Using the new equivalent circuit, explicit expressions are given for all parameters of the circuit. The derivation of the design equations from the equivalent circuit is similar to that described by Matthaei *et al.* [2]. The design equations are checked by an analysis program. The results are compared to the data given by Dishal [3] and Cristal [1].**

I. Introduction

Realizations of combline and interdigital filters with tapped-line inputs have advantages over filters with the conventional transformer input [3]. Dishal [3] describes a method for tapped interdigital filter design over a narrow bandwidth. Cristal [1] provides an exact open-wire-line equivalent circuit for the tapped-line input stage. This circuit can be utilized to perform an exact analysis of the filter. For synthesis, Cristal [1] establishes an equivalence between the transformer input and tapped-line input coupling circuits and utilizes this equivalence to apply standard interdigital filter design techniques.

This method, however, has two shortcomings:

1) An exact equivalence between transformer and tapped-line filters is not possible over a broad band of frequencies [1].
2) Graphs, rather than explicit analytical expressions, are provided. The method is therefore somewhat cumbersome from a design point of view.

In this paper, the open-wire-line equivalent circuit suggested in [1] is used as a basis for another equivalent circuit. This new circuit permits the derivation of explicit expressions for the tapped-line input parameters. The subsequent design procedure is very similar to that used by Matthaei *et al.* [2] in solving the conventional problem. The validity of the design equations is checked by using the exact equivalent circuit given in [1] in an analysis program. Combline filters up to 15 percent bandwidth and interdigital filters up to an octave bandwidth are synthesized and analyzed with good agreement. These bandwidth limits are the limits of the conventional filters' design techniques. For narrow bandwidths, the expression derived for the tapping location (Φ in Fig. 1) for the interdigital filter reduces to Dishal's expression.

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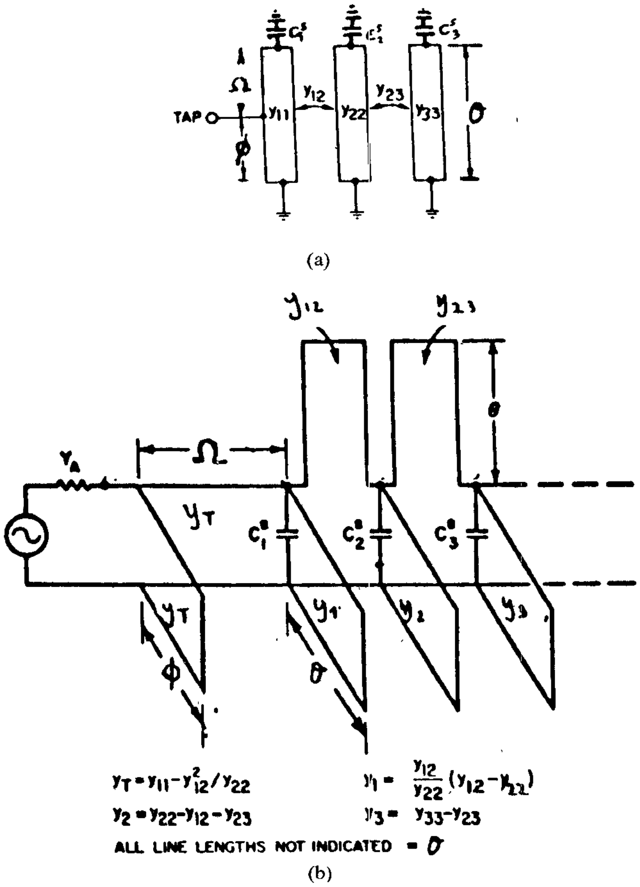

Fig. 1. Eqivalent cırcuit for tapped-lıne input combline filter. (a) Geometry and parameters. (b) Equivalent circuit.

II. Method

A. Combline Filter

The equivalent circuit suggested by Cristal [1] for the tapped-line input combline filter is shown in Fig. 1(b). The admittance inverter used to derive the equations for the combline filter [2] consists of a pi configuration, i.e., a series-shorted stub of characteristic admittance y_{12} and two shunt-shorted stubs of characteristic admittance $-y_{12}$. Therefore y_1 is split into two parts, with $y_L = (y_{12}^2/y_{22})$ added on the left to the input stage and $y_R = -y_{12}$ forming the shunt stub of the admittance inverter (J_{12}). The input stage of the filter now consists of four elements, as shown in Fig. 2.

1) New Equivalent Circuit for Input Stage: The *ABCD* matrix of the first two elements is

$$\begin{bmatrix} 1 & 0 \\ -jy_T\cot(\Phi) & 1 \end{bmatrix}\begin{bmatrix} \cos(\Omega) & jz_T\sin(\Omega) \\ jy_T\sin(\Omega) & \cos(\Omega) \end{bmatrix} = \begin{bmatrix} \alpha & \beta \\ \Gamma & \delta \end{bmatrix} \tag{1}$$

where

- θ electrical length of resonators,
- Φ electrical length from ground to input line,
- Ω $= \theta - \Phi$,
- $z_T(y_T)$ impedance (admittance) of the first and second elements.