ACCURATE PREDICTION OF THE SMALL-SIGNAL GAIN OF THE HBT DUAL-FED DISTRIBUTED AMPLIFIER

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ABSTRACT

The benefits of the MESFET dual-fed distributed amplifier over the conventional distributed amplifier (DA), both in small and large-signal operation, have previously been reported. Recent interest in the heterojunction bipolar transistor (HBT) DA has led to consideration of the HBT in the dual-fed configuration. It has been found that the previously proposed equation, developed to predict the small-signal gain of the MESFET dual-fed DA, is inaccurate when applied to the HBT dual-fed DA. This paper describes an expression which gives better agreement with the simulated performance of such an amplifier.

INTRODUCTION

The principle of the dual-fed DA is to feed the incoming signal to both ends of the input line, and combine the signals emerging from both ends of the output line, of the conventional DA. The benefits of the MESFET dual-fed DA over its conventional DA counterpart, both in small and large-signal operation, have previously been reported [1, 2, 3]. Measured results of two stage amplifiers [1, 2] have demonstrated improvements, across the lower 70% of the pass band, of 6dB in small-signal gain, 2dB in noise figure, 3dB in output power at 1dB gain compression and 7dB in third order intercept point, along with an approximate two to three-fold improvement in power added efficiency. Despite its bandwidth limitations, the technique has been used to produce 1W up to 12GHz [2].

The practical performance of DAs using the HBT rather than the MESFET as the active device has recently been reported [4, 5]. This has generated interest in the use of the HBT in the dual-fed DA configuration. The previously proposed equation for predicting small-signal gain [1], whilst valid in the case of the MESFET dual-fed DA, is inaccurate when applied to the HBT dual-fed DA. This is because implicit assumptions made in its derivation, relating to the structure of the active device and the terminations of the amplifier, are not valid in the case of the HBT. Hence, the aim has been to develop a more general expression for the small-signal gain of the dual-fed DA, which then gives an accurate prediction in the case of the HBT dual-fed DA.

DISCUSSION

Fig. 1 shows the schematic diagram of an n stage dual-fed DA, with 180° hybrids employed at input and output as a convenient means of splitting and combining the signal. It has been shown that this hybrid configuration is the optimum choice for this application [6].

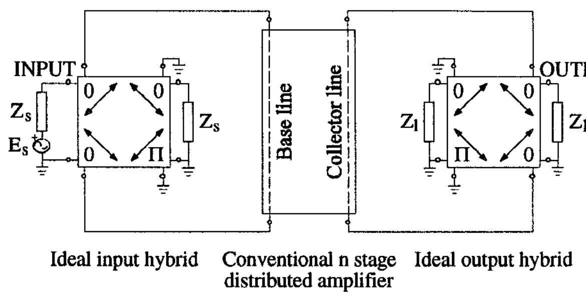

Fig. 1 An arbitrarily terminated n stage dual-fed distributed amplifier

Previously published analysis of this structure [1] assumed that intrinsic losses associated with the active device may be neglected, so that the active devices in the amplifier may be adequately represented by the model shown in Fig. 2. This assumption was found to be acceptable in the case of the MESFET dual-fed DA because, generally, this amplifier only contains two active devices. Thus, attenuation is not as significant as in the conventional DA, which usually contains four or more active devices. However, losses within the HBT are much greater than in the MESFET, so that even with only two stages, attenuation has been found to be significant in the HBT dual-fed DA. Therefore, intrinsic losses associated with the active device cannot be neglected when analysing such an amplifier. The model shown in Fig. 3 is used to represent the HBT, so that subsequent analysis to determine the small-signal gain of the dual-fed DA is based on this instead of Fig. 2. This model was first proposed in work concerning analysis of the conventional HBT DA [7, 8], where it was shown to be a valid representation of the HBT. Since the active device model of Fig. 3 is also valid for the MESFET, the derived gain expression will be general enough to be directly applicable to the MESFET dual-fed DA as well as the HBT dual-fed DA.

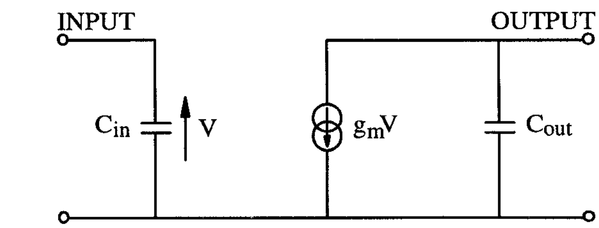

Fig. 2 Simplified loss-less active device model used in previous work [1]

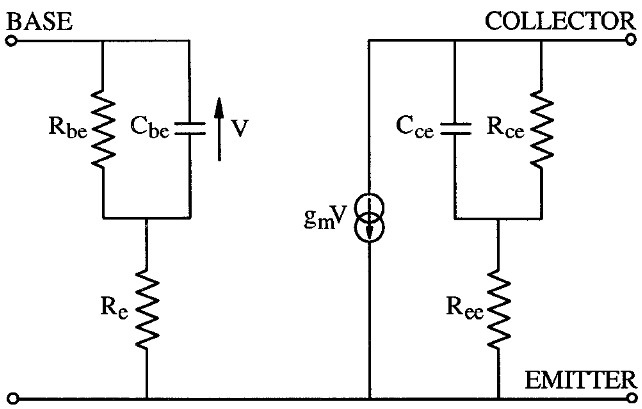

Fig. 3 Simplified lossy active device model applicable to the HBT (and MESFET)

The other major simplifying assumption made in previous work [1] is that the amplifier is ideally terminated. That is, Z_s and Z_l, shown in Fig. 1, are assumed to be equal to the T-section characteristic impedances of the lines, Z_{OTb} and Z_{OTc} respectively (defined in (8)). In practice, such terminations are unachievable, as they generally behave as a frequency dependent resistance and reactance. Fixed resistance terminations are used in practice, usually 50Ω, and so mismatch is inevitable. The degree of this mismatch can be lessened by placing *m*-derived half-sections at the ends of the lines [9], which effectively transform the constant-resistance terminations to impedances closer to the characteristic impedances of the lines.

Conventionally, design of a dual-fed DA, including its *m*-derived half-sections, is based on standard DA design methodology, which assumes a simple loss-less model of the active device [10]. This assumption is made because lossy equivalent device models, such as the one shown in Fig. 3, are not directly amenable to conventional design techniques based on artificial transmission line theory. It has been found that a dual-fed DA designed in this manner behaves less like it would if ideally terminated when using the HBT as the active device than if using the MESFET. This is because the *m*-derived half-sections used, transform the constant-resistance terminations to impedances which are not as close to the characteristic impedances of the lines in the case of the HBT dual-fed DA, as in the case of the MESFET dual-fed DA. By eliminating the assumption of ideal terminations, allowing the amplifier to be arbitrarily terminated, a more accurate prediction of the small-signal gain of the HBT dual-fed DA is obtained.

ANALYSIS RESULTS

The small-signal transducer gain, G_T, of an arbitrarily terminated n stage dual-fed DA, as shown in Fig. 1, is given by:

$$G_T = \frac{4g_m^2 Z^2 A(\omega)\left|B(\omega)\right|^2\left|C(\omega)\right|^2}{\left|e^{n\gamma_b} - \Gamma_b^2 e^{-n\gamma_b}\right|^2\left|e^{n\gamma_c} - \Gamma_c^2 e^{-n\gamma_c}\right|^2} \qquad (1)$$

where the model of Fig. 3 has been used to represent the active device. The symbols used in (1), and their constituent parts where necessary, are defined by equations (2) to (13) below.

$$A(\omega) = \frac{\left(\dfrac{R_{be}}{R_{be}+R_e}\right)^2}{1+\left(\dfrac{\omega}{\omega_b}\right)^2} \quad ; \quad \omega_b = \frac{1}{C_{be}\left(R_{be}//R_e\right)} \qquad (2)$$

and:

$$B(\omega) = e^{\frac{1}{2}n(\gamma_b+\gamma_c)} + \Gamma_b\Gamma_c e^{-\frac{1}{2}n(\gamma_b+\gamma_c)} + \Gamma_c e^{\frac{1}{2}n(\gamma_b-\gamma_c)} + \Gamma_b e^{-\frac{1}{2}n(\gamma_b-\gamma_c)} \qquad (3)$$

and:

$$C(\omega) = \frac{\sinh\left[\frac{1}{2}n(\gamma_b-\gamma_c)\right]}{\sinh\left[\frac{1}{2}(\gamma_b-\gamma_c)\right]} + \frac{\sinh\left[\frac{1}{2}n(\gamma_b+\gamma_c)\right]}{\sinh\left[\frac{1}{2}(\gamma_b+\gamma_c)\right]} \qquad (4)$$

The contribution of impedance terms to the final gain expression (1) is given by:

$$Z^2 = \frac{\mathrm{Re}(Z_s)\mathrm{Re}(Z_l)|Z_{0Tb}Z_{0\Pi b}||Z_{0Tc}Z_{0\Pi c}|}{|Z_s + Z_{0Tb}|^2 |Z_l + Z_{0Tc}|^2} \tag{5}$$

The reflection coefficients at the ports of the amplifier are defined by:

$$\Gamma_b = \frac{Z_s - Z_{0Tb}}{Z_s + Z_{0Tb}} \quad ; \quad \Gamma_c = \frac{Z_l - Z_{0Tc}}{Z_l + Z_{0Tc}} \tag{6}$$

The propagation constants on the base and collector lines are given by:

$$\gamma_i = \alpha_i + j\beta_i = 2\sinh^{-1}\left(\sqrt{\frac{Z_{1i}}{4Z_{2i}}}\right) \quad ; \quad i = \{b,\ c\} \tag{7}$$

The T and Π-section characteristic impedances of the lines are given by:

$$Z_{OTi} = \sqrt{Z_{1i}Z_{2i}\left(1 + \frac{Z_{1i}}{4Z_{2i}}\right)} \ ; Z_{O\Pi i} = \sqrt{\frac{Z_{1i}Z_{2i}}{1 + \frac{Z_{1i}}{4Z_{2i}}}} \ ; i = \{b,\ c\} \tag{8}$$

where:

$$Z_{1b} = j\omega L_b \quad ; \quad Z_{2b} = R_e + \frac{R_{be}}{1 + j\omega C_{be}R_{be}} \tag{9}$$

and:

$$Z_{1c} = j\omega L_c \quad ; \quad Z_{2c} = \frac{\left(R_{ee} + \frac{R_{ce}}{1 + j\omega C_{ce}R_{ce}}\right)\frac{1}{j\omega C}}{R_{ee} + \frac{R_{ce}}{1 + j\omega C_{ce}R_{ce}} + \frac{1}{j\omega C}} \tag{10}$$

and L_b and L_c represent the series inductance separating the active devices on the base and collector lines respectively.

In practice, it is generally arranged by adding external padding capacitance, C, that:

$$C_{in} = C_{out} + C \quad ; \quad L_b = L_c = L = C_{in}R_0^2 \quad ; \quad R_0 = 50\Omega \tag{11}$$

where C_{in} and C_{out} are conventionally chosen to be the capacitances shown in Fig. 2 [10], rather than those shown in Fig. 3.

As a specific example of the application of the above expression, the results of analysis of a dual-fed DA with 50Ω terminations and with added loss-less *m*-derived half-sections on all four ports of the DA are presented in the next section. In this application, Z_s and Z_l are not simply equal to the terminations of the amplifier, as indicated in Fig. 1, but are defined by:

$$Z_s = Z_l = \frac{1}{2}mZ_1 + \frac{R_0\left[\left(\frac{1-m^2}{2m}\right)Z_1 + \frac{2}{m}Z_2\right]}{R_0 + \left(\frac{1-m^2}{2m}\right)Z_1 + \frac{2}{m}Z_2} \tag{12}$$

where:

$$Z_1 = j\omega L \quad ; \quad Z_2 = \frac{1}{j\omega C_{in}} \quad ; \quad m = 0.6 \tag{13}$$

This redefinition of Z_s and Z_l is made in this case purely for convenience, because it avoids having to alter any of the preceding equations in order to incorporate the effect of the loss-less *m*-derived half-sections.

Note that if an equivalent unilateral model different from that shown in Fig. 3 was chosen to represent the active device, only equations (2), (9) and (10) would need to be altered. It should also be pointed out that if lossy *m*-derived half-sections were used instead of loss-less ones in the application example given below, equation (5) would need to be slightly modified.

CALCULATION AND SIMULATION COMPARISON

The small-signal transducer gain of a two stage HBT dual-fed DA predicted using the above expression has been compared to that predicted from previous work [1], which was derived specifically for the MESFET dual-fed DA. Both these predictions in turn have been compared to the simulated performance of the same amplifier, with 50Ω terminations and with added loss-less *m*-derived half-sections on all four ports of the DA, using Touchstone®. For simulation purposes, the HBT model used was based on a linearised version of a model previously published by Teeter *et al.* [11]. For the calculations, the simplified equivalent models of Figs. 2 and 3 were fitted to this full model.

Fig. 4 indicates the predictive accuracy of the two expressions with frequency, by plotting the gains calculated from them alongside the performance of the amplifier simulated using Touchstone®.

Fig. 4 demonstrates that the proposed expression (1) leads to a significantly better prediction of the small-signal gain of the HBT dual-fed DA than that given by previous work [1], reducing a maximum discrepancy between calculated and simulated gain of over 3dB to one of less than 0.5dB.

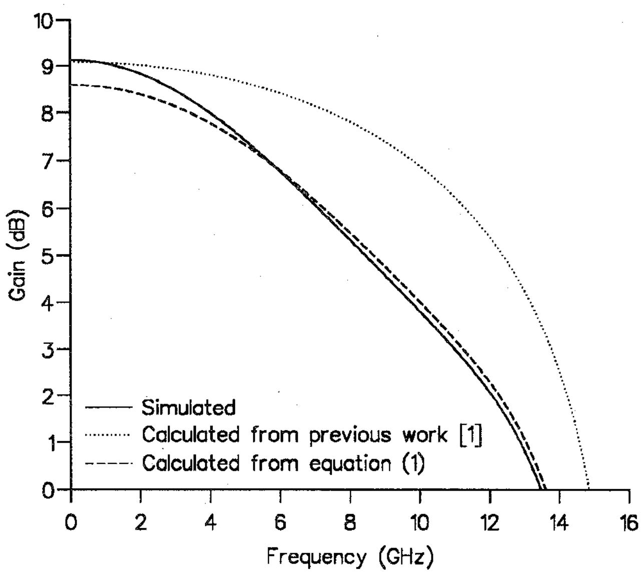

Fig. 4 Calculated and simulated small-signal transducer gain of a two stage HBT dual-fed DA (showing good agreement between our equation (1) and the simulation, but poor agreement between the earlier expression [1] and the simulation)

The gain roll-off evident from Fig. 4 is a characteristic of the dual-fed DA, which restricts its useful bandwidth. However, it can be reduced by optimising the line inductors and padding capacitors employed in the amplifier. The unoptimised performance shown in Fig. 4 indicates a 3dB bandwidth of approximately 7GHz.

CONCLUSIONS

An expression for the small-signal gain of the dual-fed distributed amplifier has been developed, which is significantly more accurate than anything previously published in the case of the HBT dual-fed DA. This is due to both a more accurate representation of the active device, and the elimination of the assumption that the amplifier is ideally terminated. Further, the resulting expression is general enough to also be directly applicable to the MESFET dual-fed DA.

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