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New design method of the single stage distributed amplifier

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ABSTRACT

This article describes a new design method of the single stage distributed amplifier, based on the approximation of the amplifier transducer gain by the Chebyshev polynomial. This method allows making a very significant improvement in bandwidth, compared with the conventional distributed amplifier with a single field effect transistor, and a significant reduction in ripple ratio with a slight broadening of the bandwidth compared with an identical amplifier topology designed with another method that we had developed. These improvements are of 128% in bandwidth compared with the conventional distributed amplifier. In addition, we can, with this method, design an amplifier so that it has the same performance as the conventional distributed amplifier with 4 transistors.

1. Introduction

The conventional distributed amplifier (CDA) consists of N field effect transistors connecting two artificial transmission lines [1], called gate line and drain line. The gate and drain lines are made out of series inductances L_g and L_d coupled with the transistors' gate and drain capacitances C_{gs} and C_{ds}. These lines are designed so that they have the same characteristic impedance, and the load Z_0 connected to each of their termination is equal to this characteristic impedance (Fig. 1).

The lines are then matched, and the loads connected to the termination 2 and 3 will absorb a certain amount of the total power, implying a reduction in power delivered to the load. This is therefore a disadvantage. So what is the goal of matching the lines? The goal is to obtain a constant gain in the bandwidth. By mismatching lines by opening the ends 2 and 3 (Fig. 1), the gain will improve but its curve will present ripple within the bandwidth. It is this type of assembly with a single transistor, based on the high-gain microwave amplifier (HGMA) [2], that we propose with a new design methodology which improves the single stage distributed amplifier (SSDA) [3]. The SSDA is a single-stage amplifier wherein the input and output load's have been removed, the fact of removing these loads will create a mismatch at the gate and drain lines, this causes fluctuation of the gain and decreases bandwidth; to overcome this problem the gain of the circuit was forced to have the same response as the Chebyshev polynomial, this approach led to a significant improvement in the amplifier's performance but does not allow controlling the rate of ripple and bandwidth, this constraint led us to propose a new approach that will design an amplifier where these two parameters become controllable. This structure (Fig. 1), which is called the mismatched single stage distributed amplifier (MSSDA), allows to have a transducer gain in low frequencies greater of 12 dB compared with the same structure but matched.

In [2], by setting the lines characteristic impedance to 50 Ω, we impose the cutoff frequency for a given transistor; it means that the bandwidth is imposed. The impedance characteristic and the lines cut-off frequency determine the band-width but have no influence on the amplifier gain. In fact, the amplifier is low pass type, its gain, therefore, is set at zero frequency and at this frequency the lines have a neutral behavior.

By considering the characteristic impedance or the cut-off frequency, or both, such as design parameters, one can control the bandwidth and also the ripple ratio. Thus in selecting only the characteristic impedance as design parameter, the SSDA [3] gave an improvement of 98.9% in bandwidth comparing with the HGMA whose lines have characteristic impedance equal to 50 Ω, but with a ripple ratio fixed at 0.37.

In the method developed in this paper, we consider the characteristic impedance and the cutoff frequency as design parameters. The result is that we can control the bandwidth and the ripple ratio according to the values of the characteristic impedance and the cutoff frequency; this is not the case of the SSDA [3], where both parameters are not controllable.

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This ability to control these two parameters allows us to make an improvement of 2% in bandwidth and a reduction of 15% in the ripple ratio compared to the results of [3].

The determination of the characteristic impedance and the cutoff frequency can be done if we define the approximation function. The most suitable approximation function is expressed as a function of the Chebychev polynomial as waves that propagate along the lines of the amplifier are stationary because of their mismatch.

2. Problem formulation

To express the amplifier transducer gain, we use the simplest model of unilateral field effect transistor (Fig. 2), where C_{gs} and C_d are respectively the transistor gate and drain capacities, and g_m its transconductance.

The transducer gain G_T is given by

$$G_T = \frac{-\frac{1}{2} R_e\left(V_2 i_2^*\right)}{\frac{|E_g|^2}{8Z_0}} \tag{1}$$

where V_2 and i_2 are respectively the output voltage and current, E_g and Z_0 the voltage and the internal impedance of the generator.

According to Eq. (1), G_T of the MSSDA is

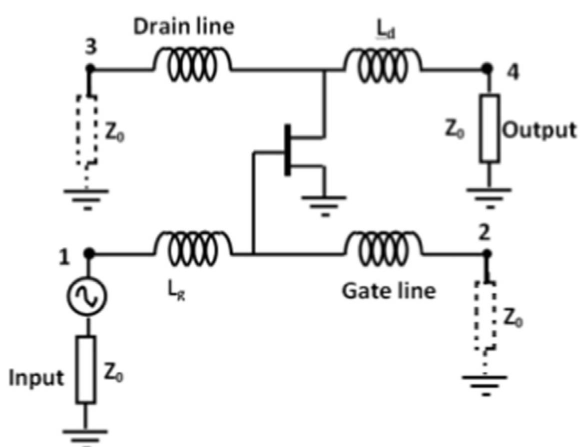

Fig. 1. Distributed amplifier (with dashed Z_0 CDA; without dashed Z_0 MSSDA).

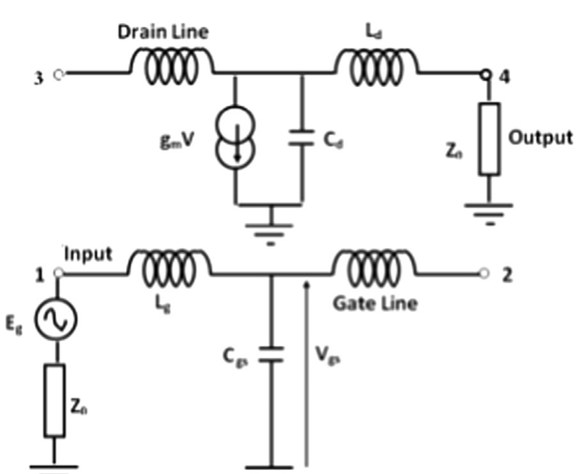

Fig. 2. Equivalent circuit of the single stage distributed amplifier.

$$G_T = \frac{4Z_0^2 g_m^2}{\left[\left(1-4\zeta^2 x^2\right)^2 + \zeta^2 \alpha_1^2 x^2\right]\left[\left(1-4a^2x^2\right)^2 + \alpha_2^2 a^2 x^2\right]} \tag{2}$$

where $x=\frac{\omega}{\omega_{c1}}$ is the normalized pulsation (frequency) compared to the cutoff pulsation (frequency) of the grid line $\omega_{c1}=\frac{2\zeta}{\sqrt{L_g C_{gs}}}$, ζ is a parameter that we can vary, $\alpha_1 = 2\frac{Z_0}{Z_{c1}}$, $\alpha_2 = 2\frac{Z_0}{Z_{c2}}$, and $a=\frac{\omega_{c1}}{\omega_{c2}}$ where $Z_{c1}=\sqrt{\frac{L_g}{C_{gs}}}$ and $Z_{c2}=\sqrt{\frac{L_d}{C_d}}$ are respectively characteristic impedances, at relatively low frequencies, of k-constant circuits constituting the grid and drain lines, and $\omega_{c2}=\frac{2}{\sqrt{L_d C_d}}$ is the cutoff pulsation of the drain line.

The choice between ω_{c1} and ω_{c2} as cutoff pulsation must relate to the one that has the lowest value so that the approximation of G_T by the polynomial of Chebyshev will be possible. ω_{c1} could be smaller in value because, in a transistor, C_{gs} is always higher than C_{ds}. Then we choose ω_{c1}.

By normalizing the gain G_T to the quantity $4Z_0^2 g_m^2$, we obtain the following expression:

$$g_T = \frac{G_T}{4Z_0^2 g_m^2} = \frac{1}{\left[\left(1-4\zeta^2 x^2\right)^2 + \zeta^2 \alpha_1^2 x^2\right]\left[\left(1-4a^2x^2\right)^2 + \alpha_2^2 a^2 x^2\right]} \tag{3}$$

Eq. (3) contains dimensionless quantities; so it can be applied independently of the transistor and the lines characteristics; this constitute an important advantage.

By developing the denominator of (3), we obtain the following result:

$$g_T = \frac{1}{\left(1 + A_2 x^2 + A_4 x^4 + A_6 x^6 + A_8 x^8\right)} \tag{4}$$

where

$$A_2 = a^2\left(\alpha_2^2 - 8\right) + \zeta^2\left(\alpha_1^2 - 8\right)$$

$$A_4 = 16a^4 + \zeta^2 a^2\left(\alpha_1^2 - 8\right)\left(\alpha_2^2 - 8\right) + 16\zeta^4$$

$$A_6 = 16\zeta^2 a^4\left(\alpha_1^2 - 8\right) + 16\zeta^4 a^2\left(\alpha_2^2 - 8\right)$$

$$A_8 = (16)^2 \zeta^4 a^4$$

The approximation by the polynomial of Chebyshev will be made on the denominator of (4). This approximation forces us to initially write the denominator of g_T in the $D = 1 + Q_n(x)$ form, which can also be expressed as

$$D = \left(1-\delta^2\right)\left[1 + \varepsilon^2\left(1 + \frac{Q_n(x)}{\delta^2}\right)\right] \tag{5}$$

where

$$\varepsilon^2 = \frac{\delta^2}{1-\delta^2} \tag{6}$$

represents the ripple ratio.

We approximate then a part of the denominator D by the polynomial of Chebyshev $T_n^2(x)$ namely:

$$1 + \varepsilon^2\left(1 + \frac{Q_n(x)}{\delta^2}\right) = 1 + \varepsilon^2 T_n^2(x) \tag{7}$$

Then

$$T_n^2(x) = 1 + \frac{Q_n(x)}{\delta^2} \tag{8}$$

As $Q_n(x)$ is of degree 8, $T_n^2(x)$ must also be of degree 8 [4]: