A NEW GENERALISED APPROACH TO THE DESIGN OF MICROWAVE OSCILLATORS

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ABSTRACT

A new approach to the design of microwave oscillators is presented which allows both frequency and power output to be predicted. The main feature of this technique is that it allows the optimum performance to be obtained from the MESFET. This is achieved by employing a generalised substitution theorem.

A design example and experimental results are given for a J-band fixed-frequency MESFET oscillator. The experimental results were found to be within 9% of the predicted values for both frequency and output power, without any experimental adjustment.

1. Introduction

The design of oscillators using various transistors has been of considerable interest for many years. A number of techniques has been developed and reported by some authors [1,3,4,5].

Generally speaking, the design consists of two parts, namely, the characterisation of the active device and the determination of embedding elements. The basic philosophy of the present design technique is to match the passive embedding elements to the conditions required for optimal operation of the active device. This means that the limitations on the performance of the MESFET caused by the arbitrary connection of the embedding elements are reduced to a minimum.

The MESFET is characterised by applying two signal sources to the isolated transistor. The fact that there is no constraint placed on the voltages and currents associated with the sources guarantees that the results obtained are defined only by the properties of the device itself. In this way, the optimisation to obtain maximum power output at a given frequency, reflects the full potential of the active device.

A two-port technique is used in the embedding design so that all the circuit elements are determined simultaneously and the difficulties caused by having to choose some of the elements empirically are avoided .

2. The Basic Steps and Equations

The design technique can be viewed as consisting of two steps. The key principle is a generalised substitution theorem for the transistor network. With reference to Figure 1, the two basic steps are:

(STEP A). The device is characterised by taking the active device without embedding elements connected to it as a two-port network N, and applying two voltage sources $U_1(t)$ and $U_2(t)$ to port I and port II respectively. The device is then simulated and the current response at each port, $I_1(t)$ *and* $I_2(t)$ is calculated.

The variables in time domain are transformed into the frequency domain so that U_1^i, U_2^i, I_1^i and I_2^i are obtained, where the superscript i = 0, 1, 2 ... K denote the variables of ith harmonics and K is the number of significant harmonics.

Optimisations, for example that to maximise the power output at a given frequency, are performed at this stage.

(STEP B). The optimised port variables U_1^i, U_2^i, I_1^i and I_2^i are used to design the passive embedding 2-port network N_f.

Various equivalent circuit and physical device models can be employed in carrying out Step A) [2,4,5,6]. The former approach is easy to implement and analyse but is generally limited in its applicability because of the DC and frequency depen-

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dence and the non-linear behaviour of devices with respect to signal level. In contrast, physical models can provide greater insight into the detailed operation of the active device, encompassing a wide range of operating conditions but needing lengthy analysis.

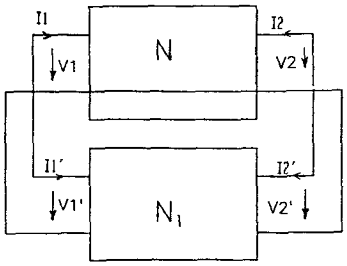

Fig.1

For the embedding network to be realised with passive elements, the real part of each of the elements (for example G_1, G_2 *and* G_3 in the following Eq. (11)) must be positive. This can be guaranteed so long as the net output power obtained in Step a) is positive, which should be noticed in carrying out the step.

Since the frequency of oscillation is determined mainly by the passive embedding network [7], higher frequency selectivity from the input port to the output port of the network is desirable. In addition to the optimisation of the power output, we took this an another criteria in carrying out Step a) to optimise Ψ_{12}, the phase difference between U_1(t) and U_2(t). In other words, we choose a value of Ψ_{12} such that the Q factor obtained for the passive network is as high as possible so as to achieve the oscillation at the specified frequency. Different emphasis can be given to the power output and the frequency selectivity to meet varied requirements of designs. Step B) was carried out by using the following equations among which (1) and (2) are always automatically satisfied, for the configurations shown in Fig 1., while (3) and (4) form the conditions for equilibrium oscillations:

$$U_1^i = U_1^{i'} \tag{1}$$

$$U_2^i = U_2^{i'} \tag{2}$$

$$I_1^i + I_1^{i'} = 0 \tag{3}$$

$$I_2^i + I_2^{i'} = 0 \tag{4}$$

It can be seen that the equations (1) to (4) reflect the relationships between the feedback power through the embedding circuit and that entering the MESFET required to maintain steady oscillations. To make this more clearer (3) and (4) can be written as

$$(I_1^i)^* = (-I_1^{i'})^* \tag{3.a}$$

$$(I_2^i)^* = (-I_2^{i'})^* \tag{4.a}$$

where "*" stands for the complex conjugate of the corresponding variables.

$$U_1^i\,(I_1^i)^* = -\,U_1^{i'}\,(I_1^{i'})^* \tag{3.b}$$

$$U_2^i\,(I_2^i)^* = -\,U_2^{\nu'}(I_2^{i'})^* \tag{4.b}$$

Under the conventional directions of the voltages and currents shown in Fig 1, these two equations mean that while the power leaving the second port of the device is equal to the power entering the second port of the passive circuit (Eq.(4.b), the power leaving the first port of the circuit is the same as that entering the MESFET (Eq (3.b)). In other words, the passive network designed according to (1) to (4) will provide proper feedback power to the device so that steady oscillations will be obtained.

From (1) to (4), we have

$$I_1^i + f_1^i\,(U_1^i, U_2^i, X, T) = 0 \tag{5}$$

$$I_2^i + f_2^i\,(U_1^i, U_2^i, X, T) = 0 \tag{6}$$

where X is a set of element parameters to be determined for N_ℓ, T is the topology of N_ℓ *and* f_1^i *and* f_2^i are linear functions of U_1^i *and* U_2^i for a certain topology.

As an example, if, for simplicity, the Π configuration shown in Fig. 2 is chosen and only the fundamental frequency is considered, it can be deduced from (5) and (6):

$$I_1 + U_1 Y_1 + (U_1 - U_2) Y_3 = 0 \tag{7}$$

$$I_2 + U_2 Y_2 + (U_2 - U_1) Y_3 = 0 \tag{8}$$

Note that the current and voltage variables in these two equations are known from the step a). Equations (7) and (8) are solved to obtain the admittances of the elements in N_e, namely, Y_1, Y_2, and Y_3.

For an admittance

$$Y_m = G_m + j\,B_m \tag{9}$$

$$m = 1, 2, 3$$

by inserting Eq. (9) into (7) and (8) then equating the real and imaginary parts of each equation to zero respectively, we obtain

$$\begin{cases} I_{1R} + U_{1R}\,G_1 - U_{1I}\,B_1 + (U_{1R} - U_{2R})G_3 \\ I_{1I} + U_{1I}\,G_1 + U_{1R}\,B_1 + (U_{1R} - U_{2R})B_3 \\ I_{2R} + U_{2R}\,G_2 - U_{2I}\,B_2 + (U_{2R} - U_{1R})G_3 \\ I_{2I} + U_{2I}\,G_2 + U_{2R}\,B_2 + (U_{2R} - U_{1R})B_3 \end{cases}$$

$$\begin{array}{l} -\,(U_{1I} - U_{2I})B_3 = 0 \\ +\,(U_{1I} - U_{2I})G_3 = 0 \\ -\,(U_{2I} - U_{1I})G_3 = 0 \\ +\,(U_{2I} - U_{1I})G_3 = 0 \end{array} \tag{10}$$

Here there are four equations with six unknown variables G_m and B_m, m=1, 2, 3, two of which, for exeample, G_1 *and* G_2, can be chosen while the remaining four are to be found from Eq.(10).

In this way, Eq. (10) becomes

$$\begin{bmatrix} U_{1R}-U_{2R} & -U_{1I} & 0 & -U_{1I}+U_{2I} \\ U_{1I}-U_{2I} & U_{1R} & 0 & U_{1R}-U_{2R} \\ U_{2R}-U_{1R} & 0 & -U_{2I} & U_{1I}-U_{2I} \\ U_{2I}-U_{1I} & 0 & U_{2R} & -U_{1R}+U_{2R} \end{bmatrix} \begin{bmatrix} G_3 \\ B_1 \\ B_2 \\ B_3 \end{bmatrix} = - \begin{bmatrix} I_1R + U_1R\, G_1 \\ I_1I + U_1I\, G_1 \\ I_2R + U_2R\, G_2 \\ I_2I + U_2I\, G_2 \end{bmatrix} \quad (11)$$

The voltages $U_1(t)$ and $U_2(t)$ applied to port I and port II should have different phases. This is achieved as follows. Denoting the coefficient matrix on the left hand side of Eq. (11) as C, the following condition should be satisfied:

$$|C| \neq 0 \quad (12)$$

where $|C|$ is the determinant of the matrix C.

Through simple manipulations of equation (11) and (12), we have

$$(U_{1R}\,U_{2I} - U_{1I}\,U_{2R}) \left[(U_{1R} - U_{2R}^2) + (U_{1I} - U_{2I}^2)\right] \neq 0 \quad (13)$$

This is equivalent to

$$U_1 \neq 0$$

$$U_{1R}\,U_{2I} - U_{1I}\,U_{2R} \neq 0$$

Or we can obtain from the above two inequalities

$$\frac{U_{1R}}{U_{1I}} \neq \frac{U_{2R}}{U_{2I}} \quad (14)$$

ie for the Eq. (11) to have meaningful solution, there must be a phase difference between $U_1(t)$ and $U_2(t)$.

The coefficient matrix C could be very ill-conditioned due to the intrinsic properties of the active device. For instance, for a MESFET circuit in common source configuration, the elements in the 2nd column of C could be very small compared with some others in the matrix. To solve this problem, scaling of some columns or rows of the matrix is sometimes necessary.

Another possible problem is that the diagonal element (3.3) in C is too small which adversely affects the conditions of the matrix to a considerable extent. This problem arises in cases where the phase angle of U_2 is close to 180°. Notice that it is the phase difference between U_1 and U_2 and not the absolute phases of the individual voltages which determins the performance of the device, we can, in general, overcome this difficulty by defining a suitable phase of U_1 so that the phase of U_2, which is determined by that of U_1 and the characteristics of the device, is around thirty degree away from 180°.

It is interesting to make a comparison between the new method and the well-established negative-impedance one-port design approach [1,3,5]. The two techniques are in principle very similar from the viewpoint that they both use an embedding circuit (or a load) with appropriate characteristics to replace the signal sources impressed on the device during the design process and simulation. On the other hand, there are several important differences. As mentioned in section 1, in addition to the fact that the "device surface" and the "load locus" of the one-port method [3,5] are generalised in a multidimensional space, the use of two signal sources in device characterisation and the fact that no element is chosen empirically makes this new approach an optimal design in the true sense.

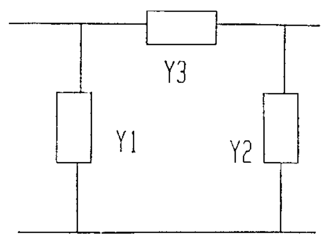

Fig.2

3. Design Example and Experimental Results

A Ku-band microwave microstrip fixed frequency oscillator was designed and fabricated using the techniques described in section 2. Fig.3 shows the circuit diagram.

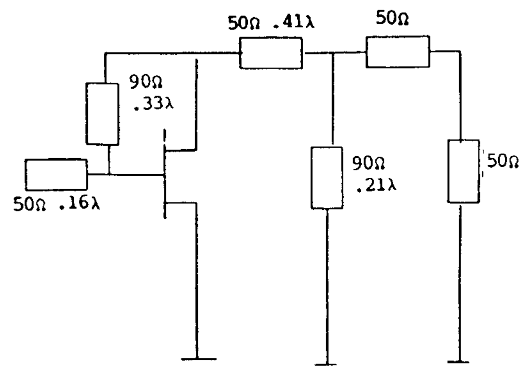

Fig.3

The device used is a NEC700 MESFET. The microstrip substrate had thickness of 0.254 mm and dielectric constant was 2.2. The measured power output and frequency were 23.7 mW and 12.7 GHz while the predicted values obtained using the above analysis technique were 21.6 mW and 12 GHz respectively. These experimental results were obtained without any empirical adjustment of the circuit. The measured spectrum of the oscillator is depicted in Fig. 4.

The whole embedding circuit was designed simultaneously. Several topologies were investigated. The Π configuration was chosen because the element values obtained with it are more feasible to be implemented with microstrip techniques. Careful consideration was also given to the layout of the microstrip to minimise parasitic effects due to discontinuities and couplings. For example, a half-turn microstrip line was used in the section between the gate and drain of the MESFET which acts as an inductor to reduce the parasitic capacitance associated with right angle corners.

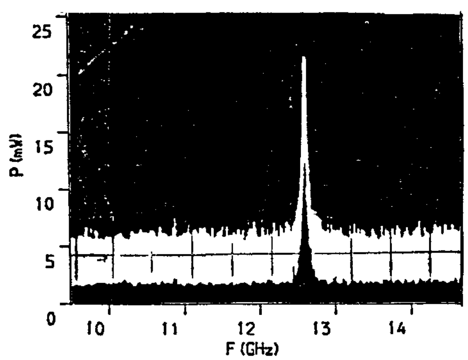

Fig.4

4. Conclusion

An efficient oscillator design approach is described. The method allows us to maximize the power output of the oscillator and is easy to be implemented. Both mathematical and physical aspects of the embedding circuit design are dicussed. The method is verified by experiments of which design examples and measued results are given. Very good agreement between the predicted and measured results has been obtained.

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