C.A.D for Broad-Band Multistage MicrowaveTransimpedance Amplifier.

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ABSTRAT

In high data rate optical-fiber, it is necessary to employ an ultra broad-band transimpedance amplifier. In this paper, we present a technique for the design of a transimpedance amplifiers. It can be applied as well to the design of interstage equalizers for microwave transimpedance amplifiers. In the version described in this paper, the optimisation process is applied to the transimpedance gain and noise which is adjusted. Based on the load charge matching technique, a sequential procedure to design multistage microwave transimpedance amplifiers is developed. The synthesis of network is carried out with lumped elements. Then we give an example of design of a two stage 3-7 GHz transimpedance amplifier with HEMT transistor.

KEYWORDS : Transimpedance-amplifier, transistor, equalizer, gain, noise, optimisation.

INTRODUCTION

In very high data rate fiber-optic systems, such as an optical receiver, it is necessary to have a band with spanning usually three decades.Several circuits employing inductor and capacitor peaking to increase bandwith [1], [2] have been reported. In this paper, we will use inductor peaking to adjust input noise and we will introduce a new procedure of design (CAD) of transimpedance amplifiers. This one is similar in concept to simplified real frequency technique (SRFT) [3]. The major difference between the two methods is in the way of the direction of design.

Knowing that a photodiode is assumed to be a high impedance current source with a shunt capacitance, in the design of transimpedance amplifier, the goal is to attain a low noise, flat gain and bring back load charge impedance equal in general at 50Ω. Then, it will be easy to compute transimpedance gain as a function of load previously fixed to 50Ω. In consequence the unfolding of the design will be make from right to left.

After recalling the equations met in the SRFT, we will see that out of the scattering parameters which caracterise each lossless reciprocal two-port E we will use Y (or Z) matrix of equalizer. The synthesis of each equalizer is maked from the unit normalized $e_{11}(p)$ (fig.1) using the darlington theorem.

With this approach, a transimpedance amplifier has been designed.

II- TRANSIMPEDANCE AMPLIFIER DESIGN TECHNIQUE.

for synthesis of E.

III - DESIGN OF MULTISTAGE FET TRANSIMPEDANCE AMPLIFIERS.

Refering to fig.2 for the first k cascaded amplifier stages, the absolute value of transimpedance Z_{tk} is given by (k > 1) :

$$|Z_{t_k}| = |G_{v_{k-1}}| \left| \frac{Y_{21_k}}{Y_{22_k} + y_{in_k}} \right| \left| \frac{y_{21_k}}{y_{22_k} + Y_{in_{k-1}}} \right| \left| \frac{1}{Y_{in_k}} \right|, \quad (9)$$

where:

$|G_{v_{k-1}}|$: is the absolute value of voltage gain of the first (k-1) stages

$(Y_{ij})_k$: are Y parameters of kth transistor F_k

$(y_{ij})_k$: are Y parameters of kth equalizer E_k

y_{ink} : is input impedance of the kth equalizer E_k

Y_{ink} : is input impedance of the kth transistor F_k

Y_{ink-1} : is input impedance of the first (k-1) stages

and for the system shown in fig.2,

$$y_{in_k} = y_{11_k} - \frac{y_{21_k}\, y_{12_k}}{y_{22_k} + Y_{in_{k-1}}} \quad (11)$$

$$Y_{in_k} = Y_{11_k} - \frac{Y_{21_k}\, Y_{12_k}}{Y_{22_k} + y_{in_k}} \quad (12)$$

IV- OPIMISATION OF GAIN.

The optimisation has been built with goal to obtain flat gain on a given width bandpass. A simple approach to optimize transimpedance amplifier gain Zt and thereby could obtain the coefficients of h(p) may be formulated by using the least-square method.

At each stage, we compute the objective function as follows :

$$E^2 = \sum_{j=1}^{N_1} \left(\frac{Z_t(\omega_j)}{Z_{t_0}} - 1 \right)^2 \quad (13)$$

Where Z_{t0} is the desired flat gain level in the bandpass, and N_1 is the number of passband sampling frequencies.

The error E is minimized using the values of $h_i + \Delta h_i$ given by the optimisation algorithm until a minimum is reached. This procedure is repeated for different sets of initi values of h_i, to ensure as far as possible that the global minimum has been attained. Tl optimisation algorithm that we have chosen is the J.MORE routine which is an improvem of Levenberg-Marquart algorithm. Vector Δh is given as [5] :

A- Formalism

It has been shown [3] that the scattering parameters of lossless reciprocal two-port E, or equalizer, can be completely determined from the numerator polynomial h(p) of the input reflexion coefficient $e_{11}(p)$. E is assumed to be a ladder network. Then he scattering parameters are given as (BELEVITCH representation) :

$$e_{11}(p) = h(p)/g(p) \; ; \; e_{21}(p) = e_{12}(p) = f(p)/g(p) \; ; \; e_{22}(p) = \pm h(-p)/g(p) \qquad (1)$$

with $f(p) = \pm p^k$ (2)

where k is an integer and specifies the order of the transmission zeros. The sign of e_{22} depends on parity of f(p), and is positive if "f" is odd, and negative if else. The polynomial h(p) $= h_0 + h_1 p + ... + h_n p^n$ is chosen as the unknown and assuming the equalizer to be lossless, the polynomial g(p) is generated from the Hurwitz factorisation of

$$g(p)\,g(-p) = h(p)h(-p) + (-1)^k\, p^{2k} \qquad (3)$$

B-New technique algorithm.

Suppose that the equalizer keeps all properties given in the formalism and that for the design example we take k = 0. From the polynomials h(p) and g(p) describe above, we can define the Y matrix of equalizer E [4]. we consider :

$$h(p) = h_1(p) + h_2(p) \qquad (4)$$

$$g(p) = g_1(p) + g_2(p) \qquad (5)$$

where, $h_1(p)$ and $g_1(p)$ are respectively the even parts of h(p) and of g(p), and $h_2(p)$ and $g_2(p)$ are respectively the odd parts of h(p) and of g(p).

The Y matrix of network E is given by :

$$(Y) = \begin{pmatrix} y_{11}(p) & y_{12}(p) \\ y_{21}(p) & y_{22}(p) \end{pmatrix} = \frac{1}{g_2(p) + h_2(p)} \begin{pmatrix} g_1(p) - h_1(p) & -f(p) \\ -f(p) & g_1(p) + h_1(p) \end{pmatrix} \qquad (6)$$

Knowing E={y_{ij}}, (i,j)=1,2, the absolute value of transimpedance Z_t of network E is given by (fig.1):

$$|Z_t| = \left| \frac{y_{21}}{y_{22} + \frac{1}{z_1}} \right| \left| \frac{1}{y_e} \right| = |G_v| \left| \frac{1}{y_e} \right| \qquad (7)$$

$$\text{where, } y_e = y_{11} - \frac{y_{21}\, y_{12}}{y_{22} + \frac{1}{z_1}} \qquad (8)$$

Knowing $h_1(p)$, $h_2(p)$ and $g_1(p)$, $g_2(p)$ imply the knowledge of h(p) and g(p), then of $e_{11}(p)$

$$\Delta h = - [J^T J + \alpha D^T D]^{-1} J^{-T} e_0 \tag{14}$$

Where, e_0 is initial error vector, J is the Jacobian matrix of e (where the elements are $\partial e_j/\partial h_i$, j=1, ... ,n), D is a diagonal matrix which takes into accout the scaling of the proble and α is the Levenberg-Marquart parameter. J.MORE introduces relationships between J, D and a which permit convergence, and this without initial guess on vector **h**.

VI- NOISE.

The noise equivalent circuit of the receiver front end is shown in fig.3. The intrinsic behavior of the transistor is described [6] by the drain noise, gate noise, and correlation coefficient, given by :

$$\overline{i_{nd}^2} = 4\,K\,T\,P\,g_m\,\Delta f \tag{15}$$

$$\overline{i_{ng}^2} = 4\,K\,T\,R\,\frac{(\omega C_{gs})^2}{g_m}\,\Delta f \tag{16}$$

$$\overline{i_{ng}^* i_{nd}} = j\,C_{cor}\sqrt{P\,R}\;4\,K\,T\,(\omega C_{gs})\,\Delta f \tag{17}$$

For the design example we take the noise coefficient values P = 1, R = 0.5, and C_{cor} = 0.9.

The p-i-n diode is described by the shot noise current source and the series thermal resistor noise, given by :

$$\overline{i_{sh}^2} = 2\,q\,I_{dc}\,\Delta f \tag{18}$$

$$\overline{i_{nR_D}^2} = \frac{4\,K\,T}{R_D}\,\Delta f \tag{19}$$

Where I_{dc} is the average detected photocurrent. Defining, $Y_D = j\,\omega C_D$, $Z_1 = R_D + j\,\omega L_D$, $R_i = R_c//R_p$, where, R_c and R_p are respectively the feedback and polarisation resistor. Additional components in the equivalent input noise current arise from :

$$\overline{i_{inR_D}^2} = 4\,K\,T\,R_D\,|Y_D|^2 \tag{20}$$

$$\overline{i_{inR_{in}}^2} = \frac{4\,K\,T}{R_{in}}\,(1 + |Y_D|^2\,|Z_1|^2 + 2\,R_e\{Y_D Z_1\}) \tag{21}$$

$$\overline{i_{inFET}^2} = \frac{4\,K\,T}{g_m}\,[(|C|^2 + |Y_D|^2\,|A|^2 + 2\,R_e\{Y_D Z_1\})\,P +$$

$$(|D|^2 + |Y_D|^2\,|D|^2 + 2\,R_e\{Y_D\,B\,D^*\})\,R\,(\omega C_{gs})^2 +$$

$$2\,R_e\{[C\,D^* + |Y_D|^2 A\,B^* + Y_D\,(A\,D^* - B\,C^*)]\,j\,C_{cor}\sqrt{P\,R}\,(\omega C_{gs})\} \tag{22}$$

Where A, B, C, D are elements of chain matrix by cascading Z_1, Rin, L_f and $C_{gs} + C_{gd}$. i_{inFET} is adjusted by L_f. The total circuit noise current referred to the input is given by :

$$\overline{i_{in}^2} = \overline{i_{inFET}^2} + \overline{i_{inR_D}^2} + \overline{i_{inR_{in}}^2} + \overline{i_{sh}^2} \qquad (23)$$

VII- EXAMPLE OF DESIGN.

We give an example of transimpedance amplifier.All transistor are identical.
Load : $Z_l = 50\Omega$, Passband : 3 Ghz $\leq f \leq$ 7 Ghz

Device : FUJITSU HEMT chip (FHX 04X) (C_{gs} = 0.2 pF, C_{gd} = 0.025 pF, g_m = 50 ms, g_d = 5.3 ms, C_{ds} = 0.049 pF, R_i = 2.5 Ω, τ = 0.85 psec, $R_G = R_S = R_D$ = 1.3 Ω, $L_G = L_D$ = 0.1 nh, L_S = 0.08 nh)

The transimpedance gain (Z_t) was 44.4 ± 0.3 dBΩ (fig.4) which corresponds to preamplifier gain (S_{21}) of 8.5 ± 1.3 dB. S_{22} was less than - 13 dB. The input noise current density (I^2_{in}) was calculated for the first stage of amplifier and we have $I^2_{in} \leq$ 1.5.1E-22 $A^2.(Hz)^{-1}$. We observed that noise diminish with inductor peaking, L_f (fig.5). The circuit obtained from the synthesis, in lumped elements, is given in figure6.

V- CONCLUSION.

It has been shown how a technique taking as a pattern of the simplified real frequency method may be used to synthesise a multi-stage transimpedance amplifier network. Results show that this technique has wide flexibility in its range of application. Synthesis is possible in distributed elements.

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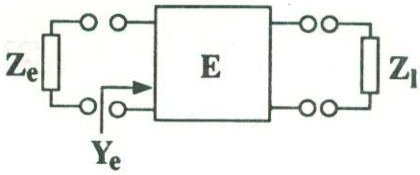

Figure 1

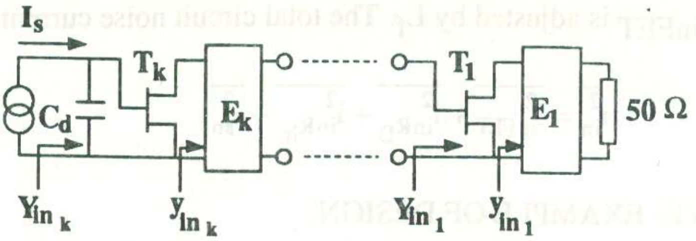

Figure 2. Computation steps for designing broad-band multistage microwave FET transimpedance amplifier.

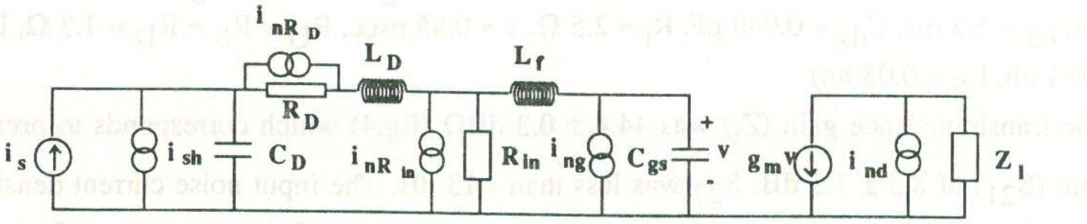

Fig.3 : Receiver front end noise equivalent circuit ($R_D = 10\ \Omega$, $C_D = 0.4$ pF, $L_D = 2$ nH, $L_f = 0.5$ nH)

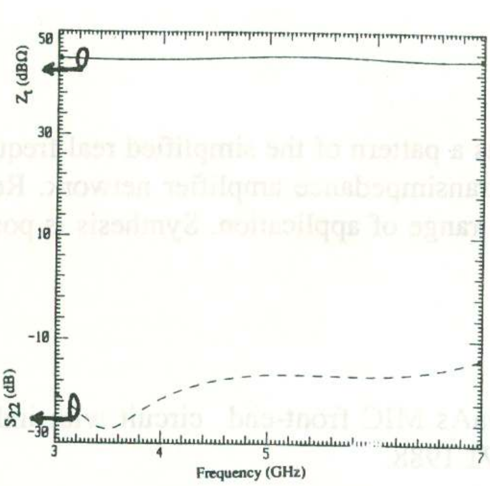

Fig 4 : Reponse of 3-7 Ghz two stages transimpedance amplifier.

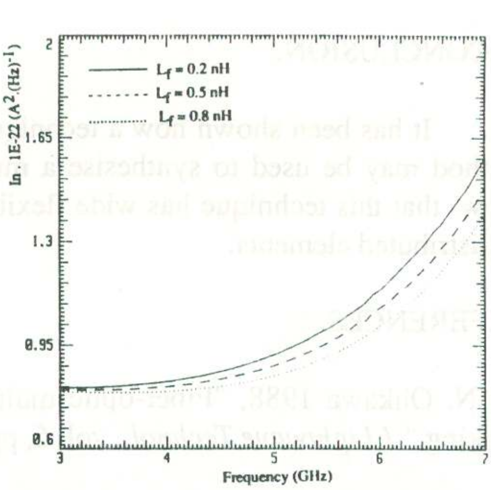

Fig 5 : Front end input noise density as function of inductor peaking.

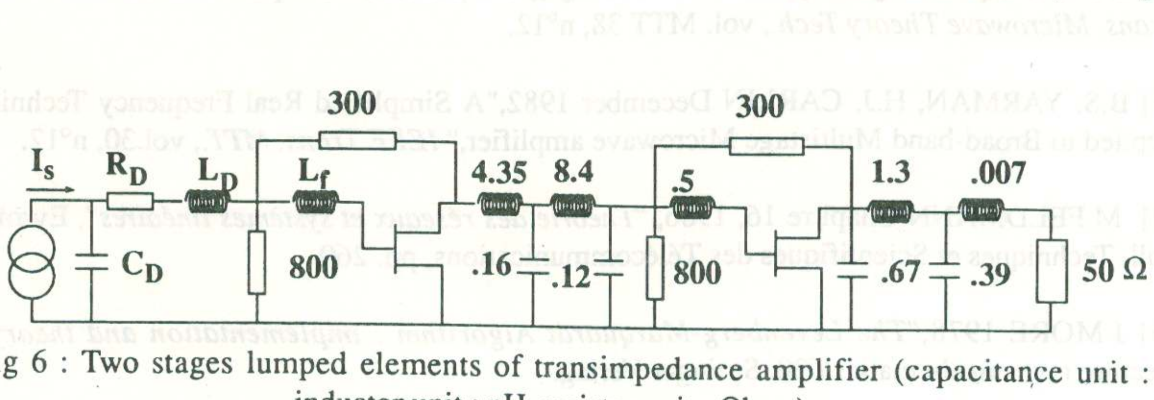

Fig 6 : Two stages lumped elements of transimpedance amplifier (capacitance unit : pF, inductor unit : nH, resistor unit : Ohms)