Design approach and performance analysis of optical receivers based on input matching network

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ABSTRACT

The performance of bandpass optical receivers can be improved using a lossless equaliser between the photodiode and the amplifier, as a consequence of the mismatch between the source and the input section of the amplifier. We propose two design approaches based on noise matching and on power matching, as well as a performance evaluation criterion to make comparisons between these different techniques. We have found the limit noise performance of the approaches at a single frequency, and a upper bound, derived from the Bode-Fano limit, in a wide-band application for power matching approach.

2. INTRODUCTION

The role of narrow band optical receivers is becoming more and more important owing to the presence of distribution services which use optical signals[1,2] to carry IF signals as AM-SCM and FM-SCM for CATV transmission systems. This kind of application uses a bandpass transmission channel and a subcarrier multiplexing technique. The bandwidth requirement is quite moderate (0.5-2GHz) with respect to the possibility of the optical fiber, while severe constraints are set for nonlinear distortions. The design of receivers for this kind of application can be greatly optimised taking into careful consideration the impedance matching conditions at the input interface. In section 3 the techniques to be used are presented; in section 4 a performance evaluation criterion will be set; in section 5 the wide band synthesis will be presented and in section 6 a case study will be shown. Concluding remarks will be given in section 7.

3. DESIGN APPROACHES

The basic structure of this type of optical receiver is shown in Fig. 1 and it consists of a P-I-N photodiode, a lossless equaliser and a bandpass amplifier. To introduce the design approaches we have regarded the amplifier as if it were fixed and consequently we set our attention on the lossless equaliser.[3,4] In the following section we will specify suitable properties for the amplifier.

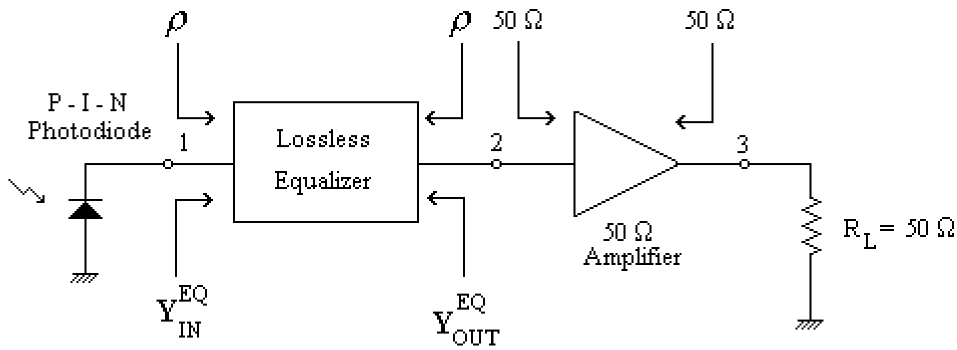

Fig. 1. Input matched optical receiver.

There are two design approaches in the synthesis of the lossless equaliser which can be justified by evaluating the signal and noise matching conditions at the interface between photodiode and amplifier.

There are two different impairments at this interface which contribute to reduce SNR at the output of the amplifier: firstly a large amount of the available power of the photodiode is wasted in the coupling between high-impedance signal source and low-impedance input port of the amplifier; secondly the amplifier is coupled to a signal source which is very far from the noise matching condition. Two different approaches are possible to remedy these impairments.

Power Matching Approach

To design the equaliser using a power matching approach (PM) which sets as main goal the condition shown in Eq. 1:

$$Y_{IN}^{EQ} = Y_G^*. \tag{1}$$

This technique takes advantage of the high impedance of the photodiode by obtaining an impedance transformation between photodiode and the input port of the amplifier. The equaliser transfers the available power of the photodiode P_{AV} to the input port of the amplifier, maximising the transimpedance gain $T=V_{out}/I_G$ of the receiver[5]

$$P_{AV} = \frac{|I_G|^2}{8} R_P \tag{2}$$

$$\left|T_{EQ}^{MAX}\right| = \frac{1}{2}\sqrt{R_P 50} \tag{3}$$

where I_G is the peak value of signal current, and where R_P is the output resistance of the photodiode, $\left|T_{EQ}^{MAX}\right|$ is the maximum modulus of the transimpedance of the equaliser. In Eq. 3 we have assumed that the amplifier is a 50Ω module and its input impedance is 50Ω at the frequency of interest. In principle the low conductance of the photodiode (R_P>1MΩ) can make high transimpedance gain available at a single frequency

$$\left|T_{EQ}^{MAX}\right| > 60\,\mathrm{db}\Omega. \tag{4}$$

From the noise performance point of view the constraint set by Eq. 1 implies a noise figure higher than NF_{MIN}.

$$NF = NF_{MIN} + \frac{R_N}{G_{OUT}^{EQ}}\left|Y_{OUT}^{EQ} - Y_{OPT}^{AMP}\right|^2 \tag{5}$$

Power matching at input port of equaliser and its lossless characteristic set Eq. 6 for output port

$$Y_{OUT}^{EQ} = \left(Y_{IN}^{AMP}\right)^* \tag{6}$$

and consequently

$$NF = NF_{MIN} + \frac{R_N}{G_{IN}^{AMP}}\left|\left(Y_{IN}^{AMP}\right)^* - Y_{OPT}^{AMP}\right|^2 . \tag{7}$$

Noise Matching Approach

Setting the condition of noise matching (NM) given by Eq. 8 as a goal , it becomes possible to get the minimum of the noise figure of the amplifier, with a limitation in the transducer power gain of the lossless equaliser

$$Y_{OUT}^{EQ} = Y_{OPT}^{AMP} . \tag{8}$$

Using scattering parameters with complex normalisation[6] it is possible to obtain transimpedance gain of the equaliser in the NM condition

$$\left|S_{22}\right| = \left|S_{11}\right| = \left|\frac{\left(Y_{OPT}^{AMP}\right)^{*} - Y_{IN}^{AMP}}{Y_{OPT}^{AMP} + Y_{IN}^{AMP}}\right| \tag{9}$$

$$\left|T\right| = \left|T_{MAX}\right|\sqrt{1-\left|S_{11}\right|^{2}}\,. \tag{10}$$

To compare and clarify the meaning of the different approaches, we have to point out the consequences of the different matching conditions: in the NM approach the design goal is equivalent to set the minimum input equivalent noise current independent of the transducer power gain of the lossless equaliser. As a consequence it becomes necessary to increase the overall gain of the amplifier, or increase the number of stages which are cascaded. In the PM approach the design goal maximises the input signal at the input port of the amplifier, but increases the noise contribution.

Eq. 8 relative to PM approach shows that the noise performance is strictly related to R_N and to the parameter Δ_{PN} defined as follows

$$\Delta_{PN} = \left(Y_{OPT}^{AMP}\right)^{*} - Y_{IN}^{AMP}\,. \tag{11}$$

The Eq. 10 relative to the NM approach also shows a dependence on Δ_{PN}. The two parameters, R_N and Δ_{PN} can be considered as figures of merit of the amplifier for this specific application: by reducing R_N and Δ_{PN} there will be an optimisation of the performance. It is important to note that a technique has been presented[7] which provides a block which satisfies the condition $\Delta_{PN} = 0$ at a single frequency and for a given load. This can be achieved using a combination of two feedback networks (shunt and series feedback) to transform the noise parameters set of a transistor.[8]

Our point of view focuses the attention of the designer on the equaliser, after having obtained a full characterisation of the photodiode and of the amplifier. In the frequency range 0-4GHz a lot of gain blocks are available which can be used for this application. It should be noted that these modules aren't usually fully characterised with the four noise parameters.

4. PERFORMANCE EVALUATION

In the previous section the different techniques shown led to results which are difficult to compare, because the PM technique increases signal and noise power with respect to the NM approach. To make some comparisons we have proposed to use the input equivalence noise current, which can be deduced from the noise figure under the constraint of defined overall transimpedance gain.

Firstly we will explain a single frequency performance comparison and later, in section 5, we will extend the analysis to wideband (WB) applications.

To offset the lower transimpedance gain of the equaliser in the NM approach, we have chosen to use a cascade of gain blocks. Each module is basically a 50Ω cascadable gain block and in our analysis it will be characterised by noise parameters {N} and admittance matrix {Y} as follows

$$\{N\}=\{NF_{OPT},\ Y_{OPT},\ R_N\} \tag{12}$$

$$\{Y\}=\{Y_{11}=Y_{11}(\omega),\ Y_{12}=0,\ Y_{21}=Y_{21}(\omega),\ Y_{22}=1/50\Omega\}. \tag{13}$$

A multistage monolithic small signal amplifier can be easily matched at its output port to 50Ω using a buffer stage, and its reverse parameter is small enough to be approximated to zero in a theoretical analysis.

Power Matching Approach

The lossless equaliser is designed to get the power matching condition (Eq. 1) at the input and output ports, consequently its transimpedance gain will be maximised

$$\left|T_{EQ}^{MAX}\right| = \frac{1}{4G_{11}G_S} = \frac{R_P}{4G_{11}}\,. \tag{14}$$

The condition set for the overall transimpedance gain specifies voltage gain of the cascade and the number of stages

$$\left|\frac{T_{RX}}{T_{EQ}^{MAX}}\right| = \left(A_V\right)^{N_S^{PM}} \quad . \tag{15}$$

The noise figure of the receiver is given by

$$NF = NF_1 + \frac{NF_2 - 1}{G_{A1}} + \frac{NF_3 - 1}{G_{A2}G_{A1}} + \ldots = NF\left(Y_{11}^*\right) + \frac{NF\left(Y_{11}^*\right)}{G_A^{MAX}} + \frac{NF\left(Y_{11}^*\right)}{\left(G_A^{MAX}\right)^2} + \ldots$$

$$NF = 1 + \left[NF\left(Y_{11}^*\right) - 1\right] \sum_{K=0}^{N_S^{PM}-1} \frac{1}{\left(G_A^{MAX}\right)^K} \quad . \tag{16}$$

Noise Matching Approach

The equaliser is designed to fulfil Eq. 8, so its transimpedance gain will be smaller than the maximum:

$$\left|T_{EQ}^{NM}\right|^2 = \left|T_{EQ}^{MAX}\right|^2 \left(1 - \left|S_{22}\right|^2\right) \tag{17}$$

the number of stages is found as in Eq. 15

$$\left|\frac{T_{RX}}{T_{EQ}^{NM}}\right| = \left(A_V\right)^{N_S^{NM}} \quad . \tag{18}$$

The noise figure of the receiver will be given by Eq. 19

$$NF = NF\left(Y_{OPT}\right) + \frac{NF\left(Y_{11}^*\right)}{G_A^{MAX} F} + \frac{NF\left(Y_{11}^*\right)}{\left(G_A^{MAX}\right)^2 F} + \ldots \tag{19}$$

where the term F is due to power mismatch at the output of the equaliser (see appendix 1). We can make a initial comparison between the noise figures obtained following the different approaches:

NM $$NF^{NM} = NF\left(Y_{OPT}\right) + \frac{NF\left(Y_{11}^*\right) - 1}{G_A^{MAX} F} \tag{20}$$

PM $$NF^{PM} = NF\left(Y_{11}^*\right) + \frac{NF\left(Y_{11}^*\right) - 1}{G_A^{MAX}} \tag{21}$$

the contribution of the first amplifier is greater for the PM approach since it is unmatched for noise, while the contribution of the second amplifier is greater for the NM condition because the available gain of the first stage is smaller.

As a first approximation it is possible to proceed to a comparison

$$NF^{PM} - NF^{NM} = NF\left(Y_{11}^*\right) + \frac{NF\left(Y_{11}^*\right) - 1}{G_A^{MAX}} - NF\left(Y_{OPT}\right) - \frac{NF\left(Y_{11}^*\right) - 1}{G_A^{MAX} F} =$$

$$= \frac{R_N|\Delta_{PN}|^2}{G_S^{PM}} + \frac{NF_{OPT} - 1 + \frac{R_N|\Delta_{PN}|^2}{G_S^{PM}}}{G_T^{MAX}} - \frac{NF_{OPT} - 1 + \frac{R_N|\Delta_{PN}|^2}{G_S^{PM}}}{G_T^{MAX} F} \tag{22}$$

we can separate the terms which contain the factor $R_N|\Delta_{PN}|^2$ to obtain a condition on the overall noise figure:

$$\frac{R_N|\Delta_{PN}|^2}{G_{11}} > \frac{(NF_{OPT} - 1)(G_T^{MAX} - G_T^{MAX} F)}{G_T^{MAX} F + F - 1} \qquad \Rightarrow \qquad NF^{PM} > NF^{NM} \tag{23}$$

This result confirms the critical role of the parameters R_N and Δ_{PN} for this kind of receiver. Eq. 23 can be justified considering that as the product $R_N\Delta_{PN}$ increases, the solution PM is progressively worsened by noise detuning. On the other hand the NM approach becomes less effective when the available gain of the single amplifier is small and/or the power reflection is high.

5. WIDE BAND CONSIDERATIONS

To extend these considerations to wideband applications it is necessary to introduce a further constraint to the characteristic of the amplifier. For the sake of simplicity we set Y_{11} = 1/50Ω so we can directly use the result of the Bode-Fano theory which sets the upper bound for the product Gain-Bandwidth of a lossless matching network. The matching of a resistive source to a load made by a resistor shunted by a capacitor is limited by[9]

$$\Delta\omega \ln \frac{1}{\rho_{MAX}} \leq \frac{\pi}{RC}. \tag{24}$$

It is possible to use this limit also for the matching of a photodiode to the amplifier where the time constant RC is given by the product of the low frequency resistance of the photodiode multiplied by its total capacitance.

It should be noted that in WB applications, the performance of the receiver is strongly affected by the capacitance of the photodiode. So a preliminary characterisation of the photodiode will be necessary to determine, as accurately as possible, the stray capacitance. Flip-chip techniques,[10,11] which are often used in ultrabroadband receivers, can be useful to obtain the best performance for this kind of receiver.

Assuming that $|S_{22}|$ is constant in the frequency range of interest, we can find an upper bound to the noise figure of the receiver. The locus of the points in the plane Y_{OUT}^{EQ} which corresponds to a constant $|S_{22}|=|\rho|$, is a circle that is defined by its centre G_C, B_C and radius r:

$$G_C(|r|) = Y_{11}\frac{1+|r|^2}{1-|r|^2} \tag{25}$$

$$B_C(|r|) = 0 \tag{26}$$

$$r(|r|) = \frac{4|r|^2}{\left(1-|r|^2\right)^2}. \tag{27}$$

Each point which belongs to the circle relative to a given |ρ| is defined by

$$G_C(|r|, f) = Y_{11}\frac{1+|r|^2}{1-|r|^2} + \frac{4|r|^2}{\left(1-|r|^2\right)^2}\cos(f) \tag{28}$$

$$B_C(|r|,f) = \frac{4|r|^2}{\left(1-|r|^2\right)^2} sin(f) \qquad -\pi \le \phi \le \pi. \tag{29}$$

Using Eqs. 29 - 30 it is possible to find the upper and lower bounds to the noise figure of the first amplifier:

$$NW_1^{WB} = NF_{OPT} + \frac{R_N}{G_{OUT}^{EQ}(|r|,f)}\left[\left(G_{SOPT} - G_{OUT}^{EQ}(|r|,f)\right)^2 + \left(B_{SOPT} - B_{OUT}^{EQ}(|r|,f)\right)^2\right] \tag{30}$$

and the noise figure of the receiver

$$NF_{TOT}^{WB} = NF_1^{WB} + \frac{NF\left(Y_{11}^*\right)}{G_A^{MAX} F} + \frac{NF\left(Y_{11}^*\right)}{\left(G_A^{MAX}\right)^2 F} + \dots \tag{31}$$

where

$$\mathrm{F} = 1 - |\mathrm{S}_{22}|^2 \tag{32}$$

Choosing the couple of values of ϕ which minimises and maximises $\mathrm{NF}_1^{\mathrm{WB}}$ for each frequency, it is possible to find a range where the noise performances of the receiver are confined.

6. A CASE STUDY

To check the results of the previous sections we present some comparisons between the PM and NM approaches at a single frequency. To make up for the lack of noise parameters of MMIC amplifiers, we have used a single MESFET, lossy matched gain block whose parameters are shown in Table 1.

f [GHz]	RE[Y11]	IM[Y11]	RE[Y21]	IM[Y21]	RE[Y12]	IM[Y12]	RE[Y22]	IM[Y22]
1,5	0,023	-0,005	0,122	-0,023	-5,20E-06	-0,001	0,022	-0,004
1,6	0,022	-0,005	0,12	-0,025	-6,10E-06	-0,001	0,021	-0,004
1,7	0,021	-0,005	0,118	-0,026	-7,00E-06	-0,001	0,021	-0,004
1,8	0,021	-0,005	0,117	-0,028	-7,70E-06	-0,001	0,02	-0,004
1,9	0,02	-0,005	0,115	-0,03	-8,50E-06	-0,002	0,02	-0,005
2	0,02	-0,005	0,114	-0,031	-9,20E-06	-0,002	0,019	-0,005
2,1	0,019	-0,004	0,113	-0,032	-7,50E-06	-0,002	0,019	-0,005
2,2	0,018	-0,004	0,111	-0,033	-5,70E-06	-0,002	0,019	-0,005
2,3	0,018	-0,004	0,11	-0,034	-3,70E-06	-0,002	0,018	-0,005
2,4	0,017	-0,004	0,109	-0,034	-1,60E-06	-0,002	0,018	-0,005
2,5	0,017	-0,004	0,108	-0,035	7,60E-07	-0,002	0,017	-0,005

Table 1a. Admittance parameters of the gain block.

f [GHz]	NFMIN	MAG[GMN]	ANG[GMN]	RN/50
1,5	5,606	0,437	-170,634	0,375
1,6	5,452	0,434	-170,288	0,364
1,7	5,297	0,432	-169,988	0,353
1,8	5,141	0,429	-169,736	0,342
1,9	4,984	0,427	-169,533	0,331
2	4,828	0,424	-169,383	0,32
2,1	4,728	0,418	-169,13	0,316
2,2	4,628	0,412	-168,911	0,313
2,3	4,529	0,405	-168,726	0,309
2,4	4,431	0,399	-168,579	0,305
2,5	4,333	0,392	-168,47	0,301

Table 1b. Noise parameters of the gain block.

6.1 Single frequency results

A lossless equaliser has been optimised to set the condition described by Eq. 6 (PM) and Eq. 8 (NM). The results of the frequencies 2, 4 GHz are shown in Tab. 2, 3 as a function of the number of amplifier stages.

f=2 GHz	T(0)	T(1)	T(2)	T(3)	NF(1)	NF(2)	NF(3)	ieq(1)	ieq(2)	ieq(3)
				[dB]					[pA/Hz^0.5]	
UNM	33,91	43,41	52,90	62,40	46,33	46,53	46,53	26,72	27,33	27,33
PM	70,40	79,99	89,49	98,98	6,37	6,74	6,78	0,24	0,25	0,25
NM	68,07	77,53	87,02	96,52	6,06	6,72	6,79	0,22	0,25	0,25

Table 2. Transimpedance and noise performance as a function of the number of stages (n) @ 2 GHz.

f=4 GHz	T(0)	T(1)	T(2)	T(3)	NF(1)	NF(2)	NF(3)	ieq(1)	ieq(2)	ieq(3)
				[dB]					[pA/Hz^0.5]	
UNM	33,72	42,12	50,53	58,93	45,68	45,80	45,91	24,78	25,12	25,44
PM	69,62	78,03	86,43	94,84	6,48	6,94	7,00	0,24	0,26	0,26
NM	67,52	75,93	84,33	92,74	6,02	6,82	6,94	0,22	0,25	0,26

Table 3. Transimpedance and noise performance as a function of the number of stages (n) @ 4 GHz.

The effectiveness of the matching network is confirmed by an improvement in the sensitivity which is greater than 20dB. For this amplifier the NM has given the best performance for both frequencies.

6.2 Wide band results

To check the previous results in a wide-band application we have optimised two equalisers, made by a ladder of lossless LC elements, with the goal of setting a PM condition on a given bandwidth and to reach the Bode-Fano limit. The results are shown in Fig. 2, 4 and the penalty found with respect to this limit is found to be less than 1 dB. The amplifier is a single stage module as previously shown optimised to be 50Ω matched on the bandwidth. The noise performances are shown in Fig. 3, 5.

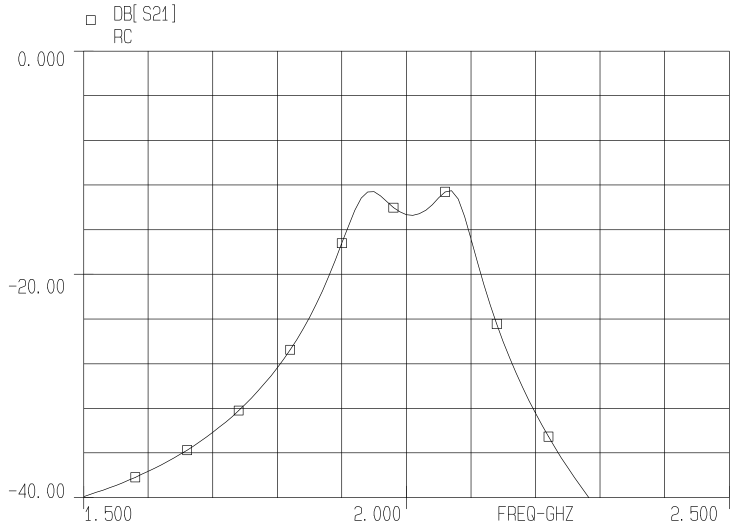

Fig. 2. S21 parameter of lossless equaliser optimised for BW=0.2GHz.

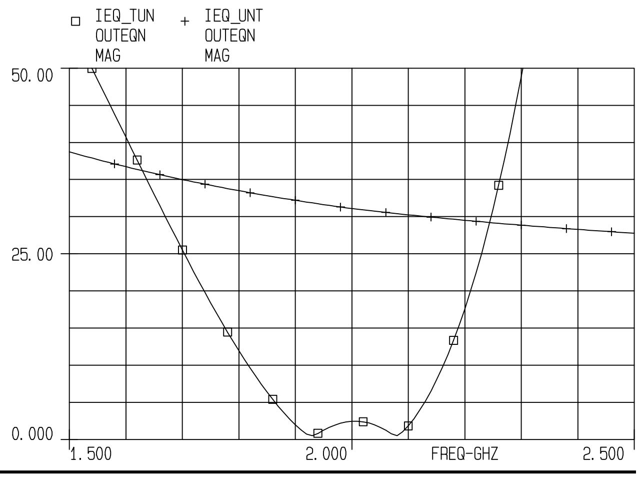

Fig. 3. Input equivalent noise current of the WB receiver optimised for 0.2 GHz (IEQ_TUN) and input equivalent current of the unmatched receiver (IEQ_UNT).

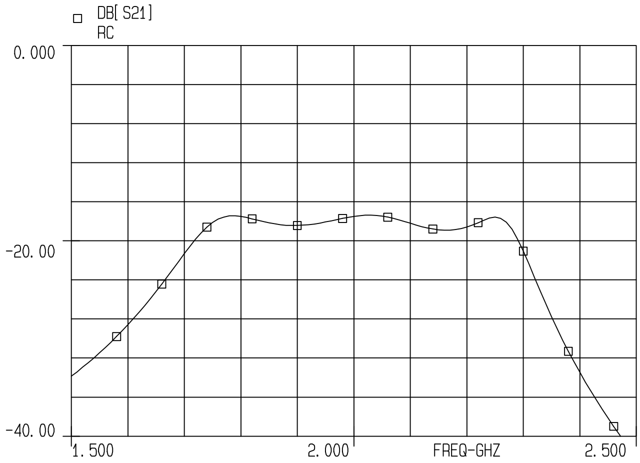

Fig. 4. S21 parameter of lossless equaliser optimised for BW=0.6GHz.

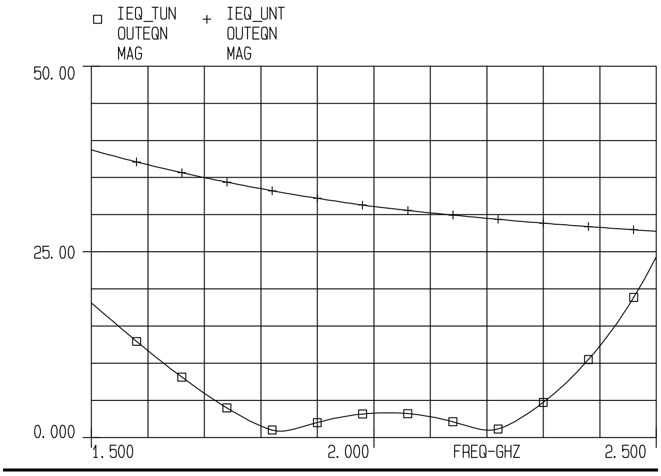

Fig. 5. Input equivalent noise current of the WB receiver optimised for 0.6 GHz (IEQ_TUN) and input equivalent current of the unmatched receiver (IEQ_UNT).

6.3 Further investigations

The effectiveness of these techniques has to be proven taking into consideration the limits imposed by the technology chosen. The performance of an equaliser which has to perform a strong impedance transformation can be severely degraded by losses and parasitic elements.

The design of the whole receiver can take advantage of a MHMIC technology which makes available libraries of lumped, high-Q passive elements. The peculiarity of MHMIC technology is to allow a easy integration of elements from different technology as MMIC modules, chip-on-carrier photodiode and its own passive elements as MIM capacitors and suspended inductors.

The design approaches presented in sections 3, 4 seem to be quite general and well defined, but they don't solve at the moment the problem of the determination of the optimum reflection coefficient which has to be set at the output of the lossless equaliser.

The technique presented by Engberg[7] seems to be very promising for this application and it should be used with microwave devices and technology to evaluate its effectiveness in terms of stability, bandwidth and reduction of gain of the gain block.

Transmission systems which use SCM techniques can take advantage in terms of non linear distortion from the input matching, since the part of the transimpedance gain due to the lossless equaliser is free from distortions.

7. CONCLUSION

Two design approaches have been presented to increase the sensitivity of optical receiver for bandpass type systems. A performance evaluation criterion has been defined and comparative results, for a gain block made by a single stage amplifier, have shown the effectiveness of these techniques. Further investigations should be performed to specify a design technique in wide band applications.

APPENDIX 1

Let consider a cascade of blocks made by amplifiers and lossless matching network as shown in Fig. 6.

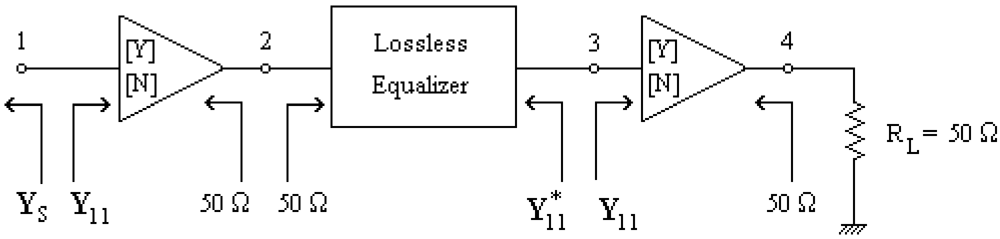

Fig. 6. Basic structure of the main amplifier used in the performance evaluation criterion.

Each amplifier is characterised by the parameter set 12-13 and is output matched to 50Ω. If the amplifier is assumed to be unilateral, its available gain can be split in two contributions[12] as in Eq. a1

$$G_A = G_A^{MAX} F(Y_G, Y_{11}) \tag{a1}$$

where

$$G_A^{MAX} = \frac{|Y_{21}|^2}{4G_{11}G_L} \tag{a2}$$

is the maximum available gain, and F

$$F(Y_G, Y_{11}) = \frac{4G_G G_{11}}{|Y_G + Y_{11}|^2}. \tag{a3}$$

is the reflection loss. The overall available gain G_A^N of a cascade of N gain blocks and matching networks where each amplifier is output matched is given by

$$G_A^N = \left(G_A^{MAX}\right)^N F(Y_G, Y_{11}) \tag{a4}$$

and its noise figure is[13]

$$NF = NF_1 + \frac{NF_2 - 1}{G_{A1}} + \frac{NF_3 - 1}{G_{A2} G_{A1}} = NF\left(Y_G\right) + \frac{NF\left(Y_{11}^*\right)}{G_A^{MAX} F(Y_G, Y_{11})} + \frac{NF\left(Y_{11}^*\right)}{\left(G_A^{MAX}\right)^2 F(Y_G, Y_{11})} \quad \text{(a5)}$$

The noise figure of the amplifier at its input port relative to Y_{OUT}^{EQ} is the same of the receiver at the input port of equaliser evaluated with respect to Y_G.[14] Noise figure can be used to evaluate input equivalent noise spectral density using the Eqs. a6, a7[15]

$$NI_{AMP}=(NF-1)N_S \quad \text{(a6)}$$

$$N_S=4kTRe[Y_G]. \quad \text{(a7)}$$

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