26PA17

MODELLING OF THE CROSS TALK IN A PHASE-SHIFT LASER RANGEFINDER

Thierry Bosch
École des Mines, Département Automatique et Productique
3 rue Marcel Sembat, 44049 Nantes Cedex 04 France
Phone: 40 44 83 19 Facsimile: 40 71 97 40
E-mail: bosch@auto.emn.fFr
Marc Lescure
ENSEEIHT, LEN7, 2 rue Camichel, 31071 Toulouse Cedex France
Phone: 33 61 58 82 83 Facsimile: 33 61 58 82 37

I INTRODUCTION

The accuracy of a laser phase-shift range finder can be limited by the influence of the cross talk between the current source used to modulate the intensity of the light beam and the photoelectric signal at the same frequency f.

For a given signal to induction ratio of 30, the maximal corresponding error is about 5 cm at 10 m. In these conditions, if the signal to noise ratio is of 10, we can not guaranteed the distance measurement, the induction being masked by the noise.

As the influence of this cross talk cannot be removed with shielding techniques, different methods have been proposed. The gain switch of a laser diode (LD) is used as a light comb generator, an avalanche photodiode selecting the 2nd harmonic of the photoelectric signal [1]. A Pockels cell is also used to obtain a measured signal at a frequency multiple of the modulation frequency [2]. But these techniques present some difficulties to be implemented.

In this paper, we propose a modelling of fictitious sources of disturbance at the input of the transimpedance to determine the rough estimate of the cross talk between the electronic circuit associated with the LD and the one associated with the photodiode (PD), according to the nature of the inductance or capacitive coupling.

II MODELLING OF THE DISTURBANCE SOURCES

Let us consider the transimpedance amplifier associated with the PD (Fig.1). The time constant τ_2 associated with the PD and the input circuit is given by the relationship:

$$\tau_2 = \left(R_1 // \frac{R_f}{1+A_0} \right) \left[C_T + C_0 + C_{in} + C_f (1 + A_0) \right]$$

where R_1 is the static load resistance of the PD and R_f the feedback resistance. $[R_f / (1 + A_0)]$ represents the input resistance of the closed loop circuit. $[C_{in}+C_f\ (1+ A_0)]$ is the input capacitor. C_T is the transition capacitor of the PD. C_0 is a parasitic capacitor.

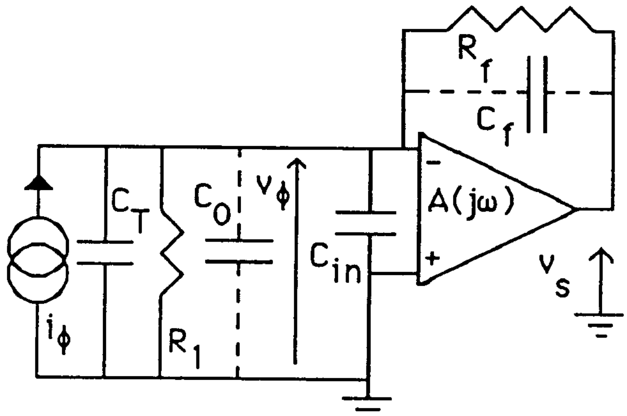

Fig. 1: the transimpedance circuit.

Let r be equal to $[R_f / (1 + A_0)]$. Let C_T' be equal to $C_T + C_0 + C_{in}$. By supposing $R_1 >> r$ we obtain the relationship:

$$\tau_2 = r \left[C_T' + C_f \left(1 + A_0\right) \right]$$

If the shielding associated to the PD is not perfect, the driver circuit of the LD, at the voltage v_3 with the current i_3, induces in the input circuit of the PD both voltages v_c and v_i respectively generated by a capacitive coupling and an inductance coupling.

Figure 2 represents the equivalent circuit at the transimpedance input. Inductance and capacitive couplings are represented by respectively: the mutual capacitor C_{30} and the mutual inductance M_{30}. Now, we consider both kinds of couplage separately.

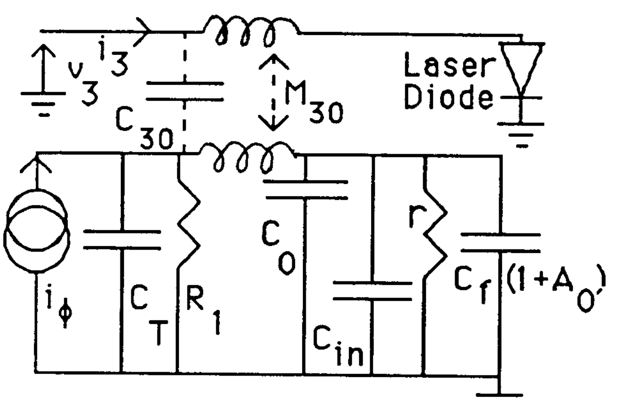

Fig. 2: Modelling of the inductance and capacitive coupling.

II-1 CAPACITIVE COUPLING

By supposing $C_{30} << C_T + C_0 + C_{in}$, the v_c voltage at the amplifier input is given by the relationship:

$$v_c = \frac{C_{30}\left[j\,\tau_2\,\omega\,v_3/(1+j\,\omega\,\tau_2)\right]}{C_T + C_0 + C_{in} + C_f\,(1+A_0)}$$

II-2 INDUCTANCE COUPLING

The voltage source of disturbance v_{30} is given by the relationship: $v_{30} = -j\,M_{30}\,\omega\,i_3$.

The v_i voltage applied to the amplifier input is obtained by the relationship:

$$v_i \approx -jM_{30}\omega \frac{1}{(1+R_1/r)}\frac{1+j\,R_1\,C_T\,\omega\,i_3}{1+j(\tau_2+rC_T)\omega}$$

As the rough estimate is the same for R_1 and R_f, C_T' and $C_f\,[1 + A_0]$, we can write:

$$r\,C_T' \approx \tau_2 / 2.$$

As the frequency of the working signal is close to the high cut-off frequency, we have the relationship: $R_1\,C_T\,\omega >> 1$, that is to say:

$$v_i \approx \frac{M_{30}\,(\tau_2/2)\,\omega^2\,i_3}{1+j\,(3/2)\tau_2\,\omega}$$

II-3 SIGNAL TO INDUCTION RATIO

The useful photoelectric signal v_ϕ at the input of the transimpedance amplifier is generated by the photoelectric current i_ϕ:

$$v_\phi = \frac{R_1 // r}{1 + j\,\tau_2\,\omega}\,i_\phi$$

The capacitive induced signal to photoelectric signal ratio is:

$$\frac{v_c}{v_\phi} = j\,C_{30}\,\omega\,\frac{v_3}{i_\phi}$$

The inductive induced signal to photoelectric signal ratio is:

$$\frac{v_i}{v_\phi} \approx \frac{M_{30}\,C_T\,\omega^2(1+j\tau_2\omega)}{1+j\,(\tau_2+r\,C_T)\,\omega}\,\frac{i_3}{i_\phi}$$

$$\frac{v_i}{v_\phi} \approx M_{30}\,\omega\,C_T\frac{i_3}{i_\phi}$$

As i_3, v_3 and i_ϕ represent small-signal modulations, we can suppose the ratios v_c / v_ϕ and v_i / v_ϕ are respectively independent of the ratios v_3 / i_ϕ and i_3 / i_ϕ. By considering the impedance $(1 / C_{30}\,\omega)$ is higher than the others, the current source of disturbance can be represented by: $i_{30} = j\,C_{30}\,\omega\,v_3$

II-4 ROUGH ESTIMATE OF THE NUMERICAL VALUES

An evaluation of the rough estimate of the couplings represented by C_{30} and M_{30} is proposed for a signal to noise ratio of 10 and a signal to induction ratio of 100.

The others values are the following:

$f = 16{,}6$ MHz ; $\Delta f_i = 1$ kHz ; $\hat{I}_3 = 10$ mA peak;

$A_0 = 100$; $R_1 = 15\ k\Omega$; $R_f = 30\ k\Omega$; $C_T = 15$ pF; $C_0 + C_{in} + C_f\,(1+A_0) = 15$ pF.

where Δf represents the bandpass of a selective 2nd order filter. $(\pi/2)\,\Delta f_i$ is the equivalent noise bandwidth.

If the noise at the input is given by the equivalent resistance (R_1 // R_f), the noise current is given by:

$$\langle i_n^2 \rangle = 4kT\frac{1}{R_1//R_f}\frac{\pi}{2}\Delta f_i = 2.6\times10^{-21}A^2$$

So, for a given spectral sensitivity $S_\lambda = 0.5$ of the PD, we can determine the equivalent flux:

$\phi_n = i_n / S_\lambda = 288$ pW peak to peak.

For a signal to noise ratio of 10, the minimum signal to be detected is of 2,88 nW peak to peak. The corresponding minimum photoelectric current $i_{\phi min}$ is 1,44 nA peak to peak.

For a capacitive coupling we obtain:

$$C_{30}=1.4\ 10^{-18}F$$

and for an inductive coupling:

$$M_{30} = 4.4\ 10^{-15}H$$

These values are excessively low.

III CONCLUSION

Our modelling of the cross talk shows that a very low inductance or capacitive coupling is sufficient to disturb roughly the distance measurement.

This is why, the influence of this cross talk cannot be removed with shielding techniques.

To decrease the capacitive coupling, it is necessary to reduce the amplitude of the driver LD voltage v_3. To reduce the value of C_{30} we need to use an electrostatic shielding linked to the ground of the electronic circuit to be protected.

We can reduce the influence of the inductive coupling by decreasing the input resistance r of the transimpedance (that is to say an increase of the cut-off frequency f_2).

REFERENCES

[1] K. Seta and T. Oh'Ishi,Japanese J. of Appl. Physics,vol 26, pp L1690-L1692, 1987.

[2] M. Lescure, Th. Bosch and A. Dziadowiec, IEEE Trans. Instrum. Meas., vol. IM-40, n°6, pp 1046-1047, 1991.