Design of Butterworth-Type Transimpedance and Bootstrap-Transimpedance Preamplifiers for Fiber-Optic Receivers

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Abstract —**This paper introduces a set of design tools for maximally flat transimpedance and bootstrap-transimpedance preamplifiers driven by capacitive current sources (photodetectors). A four quadrant graphic approach is shown to provide a good insight into a variety of solutions, all of them representing Butterworth-type circuits having the required bandwidth and transimpedance.**

The bootstrap-transimpedance is a novel configuration suggested here which by combining bootstrapping of the detector capacitance with resistive negative feedback is proved to be superior to other preamplifiers implemented in the same technology.

The design approach presented in this paper is recommended for use in the early stages of the preamplifier development, the resulting charts serving as guidelines to the design at the transistors level.

I. Introduction

A PRINCIPAL design problem of fiber-optic communication systems for transmission of analog or digital signals is the optical to electrical conversion at the receiver input. Previous authors described both high input impedance and transimpedance front-ends mainly from a noise performance point of view [1]–[6]. The transimpedance preamplifier's primary advantage is simplicity, therefore, in local data communication where the bandwidth and sensitivity requirements are relatively modest [10], such front-ends are very attractive to the designer. In such applications the transimpedance circuit can be designed to provide the required bandwidth and gain all in one stage so that equalizing circuits which follow integrating front-ends are not necessary in this case. In fact, in many applications the preamplifier's output is coupled directly to a comparator's input which completes the receiver section.

This work is concerned with a systematic approach to the design of transimpedance circuits driven by capacitive current sources (photodetectors). In particular, it is concerned with how one properly chooses the voltage gain of the basic amplifier, its poles location, its input impedance and feedback resistor so that the resulting transimpedance front-end is maximally flat (Butterworth design) and has the required bandwidth and transimpedance. The transimpedance value which in sensitivity-optimized receivers might be of secondary importance, is of primary importance here as no additional gain stages (or filter-equalizers) are available between the preamplifier output and the comparator input.

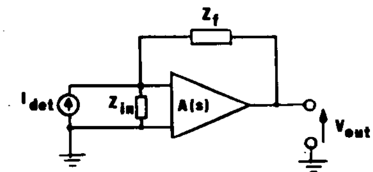

Fig. 1. Circuit model of transimpedance preamplifier.

For a given gain–bandwidth product of the active devices, the maximum achievable output voltage/input current ratio at the required bandwidth is limited in all schemes by the total input capacitance (detector and amplifier). To reduce that effect, a novel configuration that combines bootstrapping of the detector capacitance with resistive negative feedback is suggested in this paper. This bootstrap-transimpedance circuit is shown to have superior performance when compared to the high input impedance (with or without bootstrapping) and the transimpedance (feedback configuration) preamplifiers.

The design tools developed here are applicable to both transimpedance and bootstrap-transimpedance preamplifiers and are recommended for use in the early stages of the design process, mainly as guidelines to the design at the transistors level.

II. Transfer Function

Previous works [6], [7], [9], presented designs and implementations of transimpedance preamplifiers for use in optical communication. However, a detailed analysis that takes into consideration the voltage amplifier's frequency response in computing the transimpedance transfer function was not given. In [7], an important contribution to the study of such preamplifiers, the receiver model was based on a frequency independent voltage gain. The transfer function of the circuit model in Fig. 1 is given by

$$\frac{V_{\text{out}}}{I_{\text{det}}}(s)=\frac{-A(s)}{\dfrac{A(s)-1}{Z_f(s)}-\dfrac{1}{Z_{\text{in}}(s)}}. \quad (1)$$

The input capacitance C'_{in} includes amplifier capacitance C_a and detector capacitance C_{det} so that when the

Manuscript received January 29, 1981; revised August 25, 1981 and January 8, 1982.
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series resistance of the photodetector is small enough

$$C'_{\text{in}} = C_a + C_{\text{det}} \tag{2}$$

and

$$Z_{\text{in}}(s) = \frac{R_{\text{in}}}{1 + sC'_{\text{in}}R_{\text{in}}}. \tag{3}$$

Z_f is a resistor but because of parasitic capacitance associated with it one should write

$$Z_f(s) = \frac{R_f}{1 + sC_fR_f}. \tag{4}$$

A. Two-Pole Butterworth Design

Assuming that the voltage amplifier is described by

$$A(s) = \frac{-A_0}{1 + st_A} \tag{5}$$

(1) may be rewritten as

$$\frac{V_{\text{out}}}{I_{\text{det}}}(s) = \frac{-R_f\left[\dfrac{A_0R_{\text{in}}}{(A_0+1)R_{\text{in}} + R_f}\right]}{1 + s\left[\dfrac{t_A(R_{\text{in}} + R_f) + R_{\text{in}}R_f(C_{\text{in}} + A_0C_f)}{(A_0+1)R_{\text{in}} + R_f}\right] + s^2\left[\dfrac{t_AR_{\text{in}}R_fC_{\text{in}}}{(A_0+1)R_{\text{in}} + R_f}\right]} \tag{6}$$

where

$$C_{\text{in}} = C'_{\text{in}} + C_f \tag{7}$$

or

$$\frac{V_{\text{out}}}{I_{\text{det}}}(s) = \frac{R_{t0}}{1 + s\left(\dfrac{1}{QW_0}\right) + s^2\left(\dfrac{1}{W_0^2}\right)} \tag{8}$$

where R_{t0}, Q, and W_0 are defined as follows:

$$R_{t0} = -R_f\left[\frac{A_0R_{\text{in}}}{(A_0+1)R_{\text{in}} + R_f}\right] \tag{9}$$

$$W_0 = \sqrt{\frac{(A_0+1)R_{\text{in}} + R_f}{t_AR_{\text{in}}R_fC_{\text{in}}}} \tag{10}$$

$$Q = \sqrt{\frac{t_AR_{\text{in}}R_fC_{\text{in}}}{(A_0+1)R_{\text{in}} + R_f}} \cdot \left[\frac{(A_0+1)R_{\text{in}} + R_f}{t_A(R_{\text{in}} + R_f) + R_{\text{in}}R_f(C_{\text{in}} + A_0C_f)}\right]. \tag{11}$$

The poles of $V_{\text{out}}/I_{\text{det}}(s)$ are

$$s = -W_0\left[\frac{1}{2Q} \pm \frac{1}{2}\sqrt{\frac{1}{Q^2} - 4}\right]. \tag{12}$$

The values of s are real for $Q < 0.5$ and at $Q = 0.5$ the poles coincide at the value $-W_0/2Q$. For $Q > 0.5$ the two values of s become complex conjugates with the real part remaining at $-W_0/2Q$ and having the magnitude of $|s| = W_0$.

In applications where a monotonic step response is required $Q = 0.5$ is recommended, however, with $Q = 0.707$ a "maximally flat" frequency response is obtained at the expense of only 4.3-percent overshoot in the step response (two-pole Butterworth design). The -3-dB frequency for the coinciding real poles case is given by

$$W_h = W_0\sqrt{\sqrt{2} - 1} = 0.64W_0 \tag{13}$$

while in the two-pole Butterworth design the -3-dB frequency is

$$W_h = W_0. \tag{14}$$

B. Three-Pole Butterworth Design

A two-pole voltage amplifier model leads to a third-order denominator in the transimpedance transfer function. Such an analysis closely describes practical situations. For a voltage amplifier modeled by

$$A(s) = \frac{-A_0}{(1 + st_{1A})(1 + st_{2A})} \tag{15}$$

the preamplifier transfer function is

$$\frac{V_{\text{out}}}{I_{\text{det}}}(s) = \frac{N}{D(s)} \tag{16}$$

where

$$N = -A_0R_{\text{in}}R_f \tag{17}$$

$$\begin{aligned} D = {} & \left[(A_0+1)R_{\text{in}} + R_f\right] \\ & + s\left[t_{1A} + t_{2A})(R_{\text{in}} + R_f) + R_{\text{in}}R_f(C_{\text{in}} + A_0C_f)\right] \\ & + s^2\left[(t_{1A} + t_{2A})R_{\text{in}}R_fC_{\text{in}} + t_{1A}t_{2A}(R_{\text{in}} + R_f)\right] \\ & + s^3\left[t_{1A}t_{2A}R_{\text{in}}R_fC_{\text{in}}\right]. \end{aligned} \tag{18}$$

The preamplifier can be designed so that it will have a real pole and two complex-conjugate poles. For one particular three-pole configuration the peak in the frequency response disappears and the response becomes maximally flat (three-pole Butterworth), this being achieved when the poles are equally spaced around the circumference of a circle of radius W_h(-3-dB frequency) [8]. In this case the two complex poles should be located at 60 degrees relative to the negative real axis of the complex plane. Although $Q = 1$ for such poles location, the overshoot in the step response is only 8.15 percent.

Turning to the problem of relating the desired pole locations to the circuit parameters, the following conditions have to be satisfied [8]:

$$\frac{(A_0+1)R_{\text{in}}+R_f}{t_{1A}t_{2A}R_{\text{in}}R_fC_{\text{in}}}=8\pi^3f_h^3 \tag{19}$$

$$\frac{(t_{1A}+t_{2A})(R_{\text{in}}+R_f)+R_{\text{in}}R_f(C_{\text{in}}+A_0C_f)}{t_{1A}t_{2A}R_{\text{in}}R_fC_{\text{in}}}=8\pi^2f_h^2 \tag{20}$$

$$\frac{(t_{1A}+t_{2A})R_{\text{in}}R_fC_{\text{in}}+t_{1A}t_{2A}(R_{\text{in}}+R_f)}{t_{1A}t_{2A}R_{\text{in}}R_fC_{\text{in}}}=4\pi f_h \tag{21}$$

where f_h is the transimpedance preamplifier bandwidth

$$f_h=\frac{W_h}{2\pi}. \tag{22}$$

The transimpedance value at frequencies below f_h is given by

$$R_{t0}=-R_f\left[\frac{A_0R_{\text{in}}}{(A_0+1)R_{\text{in}}+R_f}\right]$$

as in the two-pole design.

III. The Design Problem

The practical case for a fiber-optic receiver designer is the following.

Given the required transimpedance (R_{t0}), bandwidth (f_h), and detector capacitance (C_{det}), choose the voltage gain (A_0), poles locations of the amplifier (t_A or t_{1A} and t_{2A}), input resistance (R_{in}), total input capacitance (C_{in}), and feedback impedance (R_f and C_f) so that the resulting circuit meets the requirements.

While the bandwidth requirement is directly related to the data-rate, the required transimpedance value (R_{t0}) is dictated by the following factors:

a) minimum expected optical input power;
b) detector conversion factor (A/W);
c) available comparator (assuming that the preamplifier is followed by a comparator);
d) dynamic range;
e) noise;
f) noise figure of the following amplifier—if one is required.

For the model in Fig. 2 the spectral density of the total output noise referred to the input as an equivalent noise current source, is given by

$$\frac{\overline{dI_{\text{eq}}^2}}{df}=\frac{4KT}{R_f}+\frac{\overline{dI_{na}^2}}{df}+\frac{\overline{dE_{na}^2}}{df}\left|\frac{1}{R_f}+jwC_f+jwC_{\text{det}}\right|^2 \tag{23}$$

where I_{na} and E_{na} represent the equivalent of the amplifier noise referred to its input.

Equation (23) only considers thermal and shot noise due to the preamplifier, assuming that the photodetector noise contribution can be neglected. Considerations regarding noise performance optimization should be made part of the design at the transistor level and will not be discussed here.

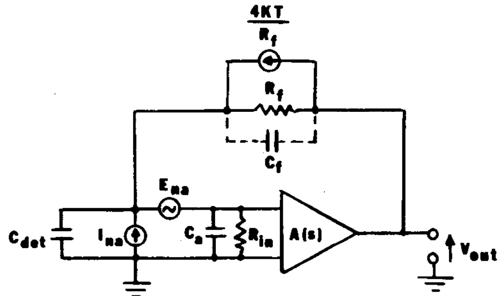

Fig. 2. The noise sources in the transimpedance preamplifier.

As mentioned in the previous section, the "maximally flat" response is reasonable for many applications, therefore, the design problem may be redefined as follows.

A. Two-Pole Butterworth

Given: $f_h, R_{t0}, C_{\text{det}}, C_f$
Find: $A_0, R_{\text{in}}, C_{\text{in}}, R_f, t_A$
So that:

$$\sqrt{\frac{t_AR_{\text{in}}R_fC_{\text{in}}}{(A_0+1)R_{\text{in}}+R_f}}\cdot\left[\frac{(A_0+1)R_{\text{in}}+R_f}{t_A(R_{\text{in}}+R_f)+R_{\text{in}}R_f(C_{\text{in}}+A_0C_f)}\right]=\frac{1}{\sqrt{2}}. \tag{24}$$

Recall that

$$f_h=\frac{1}{2\pi}\sqrt{\frac{(A_0+1)R_{\text{in}}+R_f}{t_AR_{\text{in}}R_fC_{\text{in}}}}$$

$$R_{t0}=-R_f\left[\frac{A_0R_{\text{in}}}{(A_0+1)R_{\text{in}}+R_f}\right]$$

$$C_{\text{in}}=C_a+C_{\text{det}}+C_f. \tag{25}$$

B. Three-Pole Butterworth

Given: $f_h, R_{t0}, C_{\text{det}}, C_f$
Find: $A_0, R_{\text{in}}, C_{\text{in}}, R_f, t_{1A}, t_{2A}$
So that: (19)–(21) are satisfied.

IV. Computer-Aided Design

Due to the existing degrees of freedom in the design problems formulated above, computer-aided design is most desirable. A four-quadrant graphic approach is suggested for presenting the results so that the designer has a good insight into a variety of solutions before a detailed design at the components level is undertaken.

A. Two-Pole Butterworth

Substituting t_A in (24) by (25) and solving for A_0, the following is obtained:

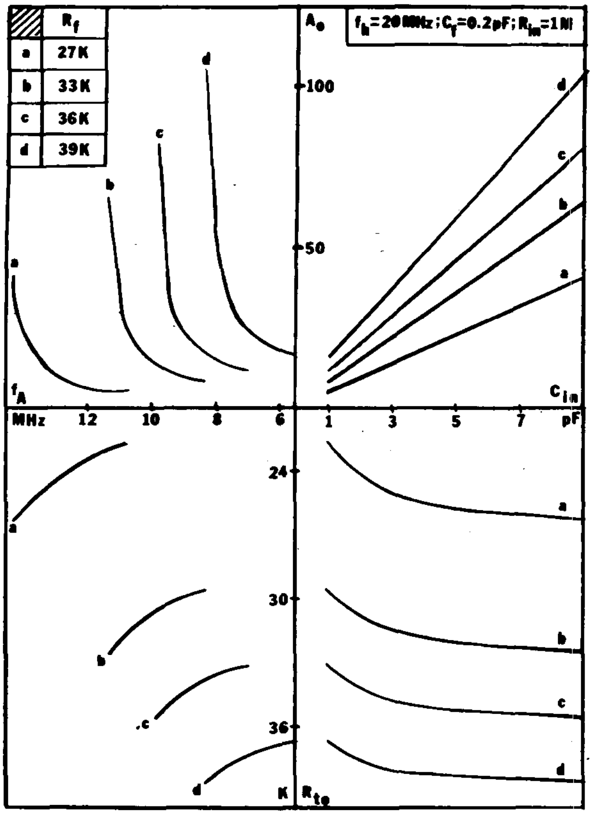

Fig. 3. Design chart for two-pole Butterworth preamplifier $f_h = 20$ MHz, $C_f = 0.2$ pF, $R_{\text{in}} = 1$ MΩ.

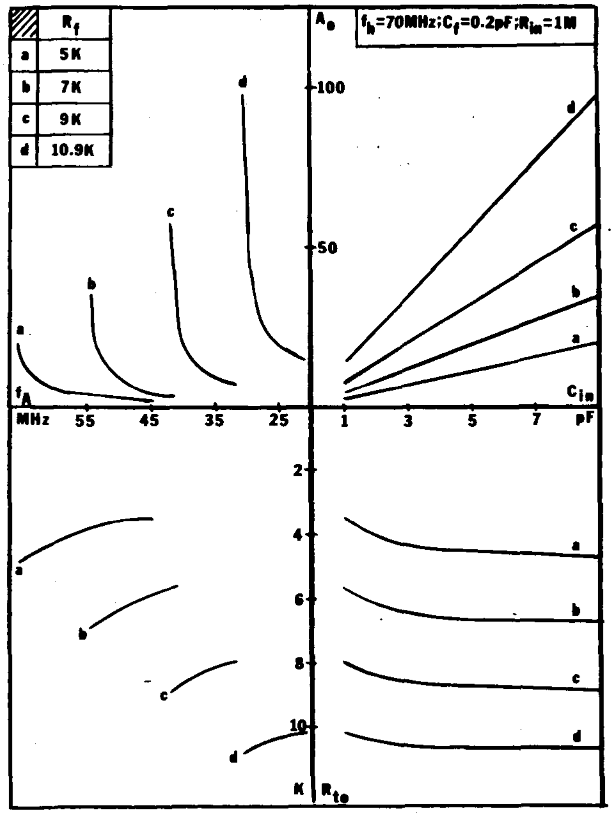

Fig. 4. Design chart for two-pole Butterworth preamplifier $f_h = 70$ MHz, $C_f = 0.2$ pF, $R_{\text{in}} = 1$ MΩ.

$$A_0 = \frac{R_{\text{in}} R_f (C_{\text{in}} - C_f) + \dfrac{(R_{\text{in}} + R_f)}{W_h^2 R_{\text{in}} C_{\text{in}}} - \dfrac{\sqrt{2} R_f}{W_h}}{\dfrac{\sqrt{2} R_{\text{in}}}{W_h} - \dfrac{(R_{\text{in}} + R_f)}{W_h^2 C_{\text{in}} R_f} - R_{\text{in}} R_f C_f} - 1. \tag{26}$$

Using (26) a computer can be programmed to calculate and plot A_0 versus C_{in}, this being repeated for several values of R_f in the vicinity of the required R_{t0}. For each calculated value of A_0 the program will use (9) to plot the resulting R_{t0} versus C_{in}. Finally, the voltage amplifier bandwidth (f_A) will be calculated and plotted using the following:

$$f_A = \frac{2\pi f_h^2 R_{\text{in}} R_f C_{\text{in}}}{(A_0 + 1) R_{\text{in}} + R_f}. \tag{27}$$

The latter is obtained using $f_A = 1/2\pi t_A$ and (10) and (14).

Typical results of such computing are given in Fig. 3. The curves in the four quadrants represent a useful design tool and can be obtained in a few seconds if the work is done by computer. Typical design charts for higher speed circuits are given in Fig. 4. In both charts (Figs. 3 and 4) $R_{\text{in}} = 1$ MΩ was used as a typical value, high input resistance is required in most cases to reduce the input current noise. A parasitic capacitance value of 0.2pF (C_f) was assumed to be associated with the feedback resistor (R_f), this could be higher in some cases [6]. Techniques for minimizing this parasitic capacitance effect were suggested in [6] and [7].

B. Three-Pole Butterworth

We solve (19) for A_0 and substitute A_0 in (20). Equations (20) and (21) can now be solved for t_{1A} and t_{2A}. This is done using the following substitutions:

$$t_{1A} + t_{2A} = x \tag{28}$$

$$t_{1A} t_{2A} = y. \tag{29}$$

Solving (20) and (21) for x and y leads to the following expressions:

$$y = \frac{R_f^2 C_f + R_{\text{in}} R_f C_f - R_{\text{in}} R_f C_{\text{in}}}{(R_{\text{in}} + R_f) \dfrac{4\pi f_h R_{\text{in}} R_f C_{\text{in}} - R_{\text{in}} - R_f}{R_{\text{in}} R_f C_{\text{in}}} + 8\pi^3 f_h^3 R_{\text{in}} R_f^2 C_{\text{in}} C_f - 8\pi^2 f_h^2 R_{\text{in}} R_f C_{\text{in}}} \tag{30}$$

$$x = y \frac{4\pi f_h R_{\text{in}} R_f C_{\text{in}} - R_{\text{in}} - R_f}{R_{\text{in}} R_f C_{\text{in}}} \tag{31}$$

$$A_0 = \frac{8\pi^3 f_h^3 y R_{\text{in}} R_f C_{\text{in}} - R_f - R_{\text{in}}}{R_{\text{in}}}. \tag{32}$$

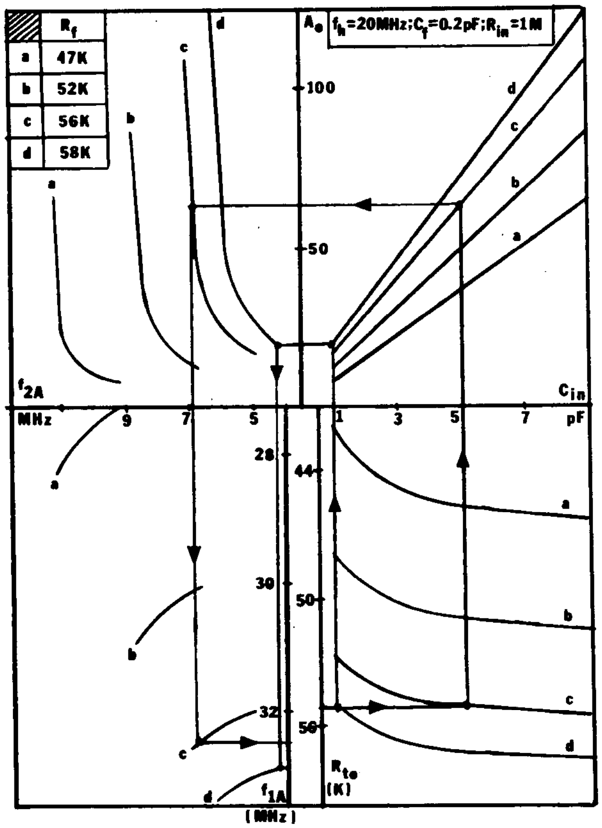

Fig. 5. Design chart for three-pole Butterworth preamplifier $f_h = 20$ MHz, $C_f = 0.2$ pF, $R_{\text{in}} = 1$ MΩ.

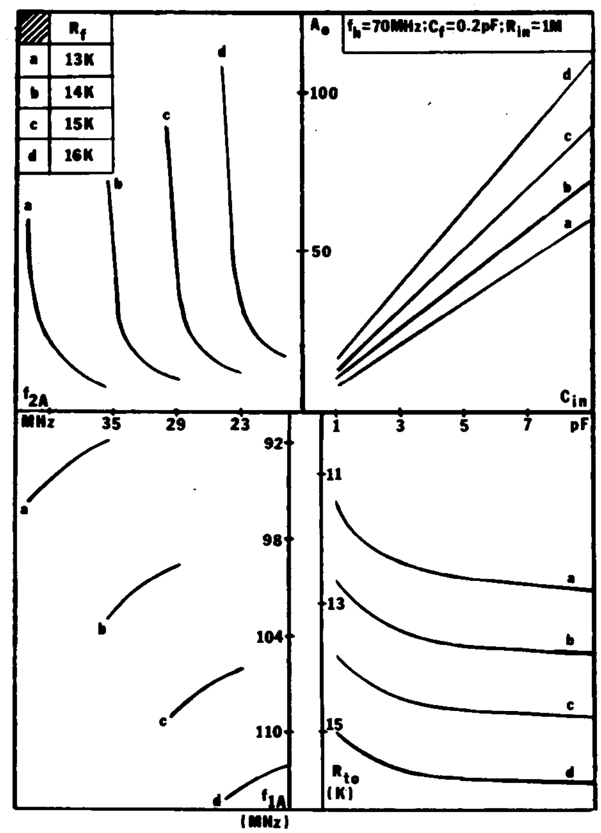

Fig. 6. Design chart for three-pole Butterworth preamplifier $f_h = 70$ MHz, $C_f = 0.2$ pF, $R_{\text{in}} = 1$ MΩ.

Using (30) and (32) a computer can be programmed to calculate and plot A_0 versus C_{in}, this being repeated for several values of R_f in the vicinity of the required R_{t0}. For each calculated value of A_0 the program will use (9) to plot the resulting R_{t0} versus C_{in}. The voltage amplifier poles (or the related corner frequencies) will be calculated and plotted using the following:

$$t_{2A} = \frac{x + \sqrt{x^2 - 4y}}{2} \tag{33}$$

$$t_{1A} = \frac{y}{t_{2A}} \tag{34}$$

$$f_{1A} = \frac{1}{2\pi t_{1A}} \tag{35}$$

$$f_{2A} = \frac{1}{2\pi t_{2A}}. \tag{36}$$

Typical results are shown in Figs. 5 and 6.

V. An Improved Configuration: The Bootstrap-Transimpedance

A comparison of two solutions that have the same transimpedance and bandwidth but different input capacitance will emphasize the importance of keeping the total input capacitance as low as possible, the penalty for higher input capacitance being the requirement for a higher gain–bandwidth product voltage amplifier. Both solutions shown in Fig. 5 have a transimpedance value of 55 K (R_{t0}) at 20-MHz bandwidth (f_h). However, the solution that has a 5-pF input capacitance requires an amplifier with a larger gain bandwidth product than the solution that has 1-pF input capacitance.

Total input capacitance can be reduced by implementing the model suggested in Fig. 7. In this case the ac voltage drop on the detector capacitance C_{det} is zero, thus "no current will flow through C_{det}". Solving for the circuit's transfer function, the same results as in the previous sections are obtained except the following:

$$C_{\text{in}} = C_a + C_f. \tag{37}$$

Recall, that for the model in Fig. 1, the total input capacitance was given by $C_{\text{in}} = C_a + C_{\text{det}} + C_f$.

The tools developed in the previous section are, therefore, applicable to this new configuration—the bootstrap transimpedance. The spectral density of the preamplifier noise referred to its input is given by (23) as for the transimpedance circuit, but in fact for a given gain–bandwidth product voltage amplifier, lower noise level can be achieved using a higher value feedback resistor R_f, as it will be shown later.

Turning to a comparison between the three preamplifier configurations in common use (Fig. 8(a), (b), (c)) and the new bootstrap-transimpedance suggested here (Figure 8(d)), its advantages will be clarified.

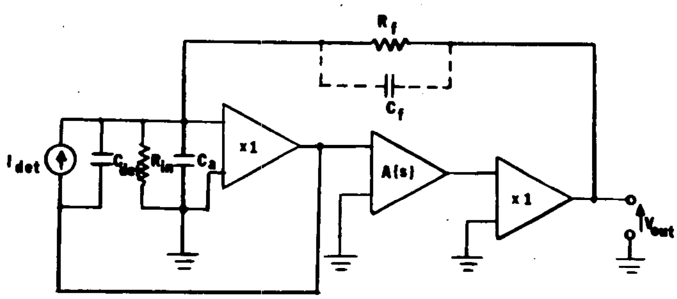

Fig. 7. Circuit model of improved transimpedance preamplifier—the bootstrap transimpedance.

TABLE I
FOUR PREAMPLIFIER DESIGNS HAVING THE SAME TRANSIMPEDANCE (11 K ± 10 PERCENT) AND BANDWIDTH (70 MHz)

Configuration	Poles Location	Transimpedance		R_L	R_f	A_o	f_A	f_{1A}	f_{2A}	A_of_A or A_of_{2A}
Fig. 8a	Two coinciding real poles	$R_{to}=A_oR_L$	11K	291	/	37.8	109 MHz	/	/	4.12 GHz
Fig. 8b	Two coinciding real poles	$R_{to}=A_oR_L$	11K	1456	/	7.55	109 MHz	/	/	823 MHz
Fig. 8c	Two-pole Butterworth	R_{to}	11.84 K	/	12K	78.87	23.1 MHz	/	/	1.82 GHz
	Three-pole Butterworth	R_{to}	11.58 K	/	12K	28	/	85.17 MHz	52.16 MHz	1.46 GHz
Fig. 8d	Two-pole Butterworth	R_{to}	11.53 K	/	12K	24.54	14.45 MHz	/	/	354 MHz
	Three-pole Butterworth	R_{to}	10.28 K	/	12K	6	/	81.5 MHz	45.2 MHz	271 MHz

A. Voltage Amplifier: Gain–Bandwidth Product Requirements

For a given transimpedance and bandwidth, a bootstrap-transimpedance implementation requires lower gain bandwidth from its voltage amplifier than the other three approaches. A typical situation is described in the following example.

A preamplifier having $R_{t0}=11$ K ± 10 percent and $f_h=$ 70 MHz has been designed using four different approaches (Fig. 8). Detector capacitance is 4 pF, amplifier capacitance is 0.8 pF, resistor capacitance is 0.2 pF, input resistance is 10 MΩ. The results given in Table I emphasize the superiority of the suggested bootstrap-transimpedance, practically meaning simple and low cost circuitry. In the three-pole Butterworth design the required gain bandwidth is only 271 MHz while in all other designs this figure is considerably higher.

B. Noise Considerations

For the circuits in Fig. 8 (a) and (b) the spectral density of the equivalent noise referred to input is given by

$$\frac{\overline{dI_{\text{eqv}}^2}}{df}=\frac{4KT}{R_L}+\frac{\overline{dI_{na}^2}}{df}+\frac{\overline{dE_{na}^2}}{df}\left|\frac{1}{R_L}+jwC_{\text{det}}\right|^2. \tag{38}$$

Assuming that in all four designs of Fig. 8, I_{na} and E_{na} are the same, the difference in the noise level will depend mainly on the R_L or R_f value. In the Bootstrap-Transimpedance circuit a 20 K feedback resistor, for example, with a 328-MHz gain–bandwidth amplifier ($A_0=262$, $f_{2A}=1.25$ MHz, $f_{1A}=130.7$ MHz) will have a 70-MHz bandwidth as required, higher transimpedance value and less noise than the previous designs. (Note that at 328-MHz gain bandwidth the amplifier is still much simpler than in the other designs.)

C. Signal Level

Having the ability to increase the transimpedance value (by increasing R_f) without running into gain–bandwidth problems makes the signal processing task easier (for the defined minimum input signal, output signal is larger).

D. Optical Signal Coupling

In fiber-optic receivers the area of the detecting device is of primary importance. Coupling between fiber and detector is easily performed with "large" detectors. Unfortunately, large area detectors have large capacitance and, therefore, their use with preamplifiers that have detector-capacitance limited performance is not recommended.

In [11] a very "small-geometry" detector was used to

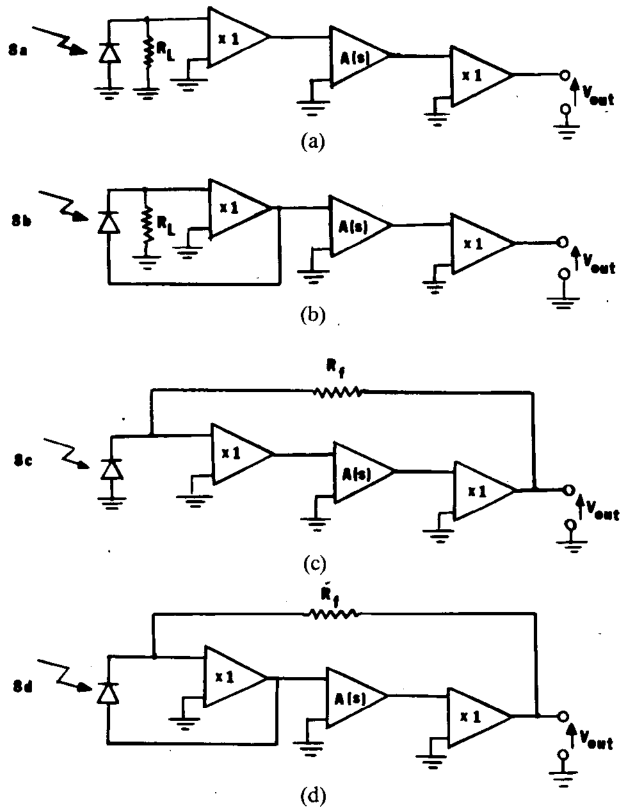

Fig. 8. Four preamplifier configurations based on the same building blocks.

reduce the total input capacitance and, therefore, a lens was necessary for coupling light into it.

The bootstrap-transimpedance performance does not depend on the detecting device capacitance, thus making coupling easier.

E. Practical Bootstrapping Considerations

The implementation of the bootstrap-transimpedance configuration depends on the application. The used bootstrapping technique, in particular, will depend on the detector type and the available voltage follower. For low-voltage p-i-n diodes (less than 5 V reverse bias) the voltage follower can be designed so that the p-i-n detector is dc coupled between its input and output. In applications with photovoltaic detectors (low speed receivers) zero dc voltage between input and output is required. This can be achieved by an operational amplifier wired as a voltage follower.

Avalanche-photo-diodes (APD) will usually be ac coupled to the input and output of the voltage follower because of the high dc voltage required for their biasing. This imposes certain restrictions on the transmitted data format. Manchester encoding or other schemes with short-term zero average are suitable in this case.

VI. Sensitivity to Parameters Fluctuations

The described design is based on a "maximally flat" solution, therefore, the logarithmic sensitivity of Q should be the designer's main concern. The following is an analysis of the two pole Butterworth design from this point of view.

Q is given by (11) and in a practical case, fluctuations of A_0 and t_A will have the dominant role. Differentiating (11) the following is obtained:

$$S^Q_{A_0} = \frac{\frac{dQ}{Q}}{\frac{dA_0}{A_0}} = \frac{1}{2}\left[\frac{A_0R_{\text{in}}}{(A_0+1)R_{\text{in}}+R_f}\right] - \frac{A_0C_f}{A_0C_f+C_{\text{in}}+t_A\dfrac{R_{\text{in}}+R_f}{R_{\text{in}}R_f}} \tag{39}$$

$$S^Q_{t_A} = \frac{\frac{dQ}{Q}}{\frac{dt_A}{t_A}} = \frac{1}{2}\left[\frac{R_{\text{in}}R_f(C_{\text{in}}+A_0C_f)-(R_{\text{in}}+R_f)t_A}{R_{\text{in}}R_f(C_{\text{in}}+A_0C_f)+(R_{\text{in}}+R_f)t_A}\right]. \tag{40}$$

In most practical cases $A_0R_{\text{in}} \gg R_f$ and, therefore, the absolute value of $S^Q_{A_0}$ is smaller than 0.5. S^Q_{ta} is always smaller than 0.5 so that for a worst-case design the following can be used:

$$\frac{dQ}{Q} \leqslant \sqrt{\left(S^Q_{A_0}\frac{dA_0}{A_0}\right)^2+\left(S^Q_{t_A}\frac{dt_A}{t_A}\right)^2} = \frac{1}{2}\sqrt{\left(\frac{dA_0}{A_0}\right)^2+\left(\frac{dt_A}{t_A}\right)^2}. \tag{41}$$

For many applications a Q ranging from 1/2 to 5/6 (zero to 10-percent overshoot) is reasonable, therefore, $\pm 1/6$ (approximately) fluctuations can be allowed around the nominal $Q=0.707$. In this case, using (41) the allowed total fluctuations of A_0 and t_A are given by (42) which is equivalent to ± 33 percent for both A_0 and t_A:

$$\left(\frac{dA_0}{A_0}\right)^2+\left(\frac{dt_A}{t_A}\right)^2=\frac{2}{9}. \tag{42}$$

The logarithmic sensitivity of R_{t_0} is given by

$$S^{R_{t_0}}_{A_0} = 1-\frac{A_0R_{\text{in}}}{(A_0+1)R_{\text{in}}+R_f}. \tag{43}$$

So that for ± 33-percent variations in A_0 one should expect less than ± 33-percent variations in R_{t_0}. In fact, in most practical cases the expression in (43) is approximately zero.

VII. Conclusions

It has been shown that transimpedance and bootstrap-transimpedance preamplifiers for fiber-optic receivers can systematically be designed using the suggested charts. Using a computer, such charts are easy to obtain and they provide a good insight into a variety of solutions all of them representing maximally flat circuits having the required bandwidth and transimpedance. The charts are recommended as guidelines to the detailed design at the components level.

The bootstrap-transimpedance, a novel practical configuration which minimizes total input capacitance, has been shown to reduce the voltage amplifiers required gain–bandwidth product for a given transimpedance and bandwidth, or to reduce noise and provide higher transimpedance when using a given voltage amplifier. This configuration also makes coupling between the photon source and detector easier as larger detectors can be used without compromising transfer function.

Although fiber-optic local data communication was the background of this work, the developed tools are applicable to optical and nuclear detector-preamplifiers in general.

Acknowledgment

The author wishes to thank Digital Equipment Corporation for permission to publish this paper and Mrs. B. Cross for typing the manuscript.

The comments made by those who reviewed this paper are gratefully acknowledged.

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