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A New CMOS Second-Order Temperature-Compensation Branch-Current Reference

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Abstract – In this work a new compact scheme of a branch-current reference is proposed, composed of two cross-connected classical current mirrors: a modified Wilson mirror and a Widlar mirror. This reference-source type can be connected to a charge with grounded or connected to voltage-supply end only by a supplementary mirror branch. The first and second-order thermal-compensation conditions are deduced. The design and the simulation results are included. The second-order thermal-compensation reference-source performances are: a reference-current maximum variation of 0.9% across a temperature range of 0 - 100°C and a „supply regulation" of 4550ppm/V.

Keywords: CMOS analog integrated circuits, current references, temperature compensation.

I. INTRODUCTION

The compact and simple reference-current sources, composed of two cross-connected current mirrors were brought back in discussion by work [1]. There it was considered first time the controlled temperature dependence of the two source-branch current ratio. Then this concept permitted to design a source composed of a diode-completed Widlar mirror and a "reverse"-Widlar mirror with cross-connection (fig.1), for the purpose of obtaining a spectacular second-order thermal-compensation of a branch current, which led to very good performances. But, in work [1] it is no showed or discussed the reference-current extraction for use in a charge with grounded or connected to voltage-supply end. Of course, this current can be sent to the charge by a supplementary current mirror but, as it was established by our simulations, the new-mirror current ratio is temperature-affected too. So, the output current stability should have an important degree of compromising as against the obtained performance on the main source branch. Moreover, a simple additional branch cannot always assure a satisfactory value of the „supply-regulation" parameter for the output-towards-charge circuit.

The works [2] and [3] have introduced for these compacted sources the notion of reference "**total**" current, representing the sum of two-branch currents.

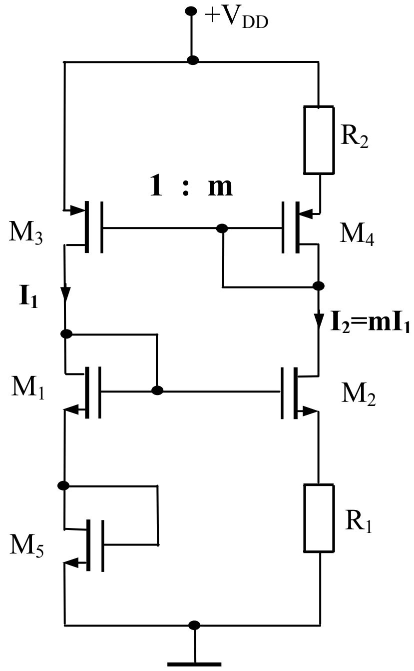

Fig.1. Second-order temperature-compensation branch-current reference (Fiori-Crovetti)

It is a new concept, which permits a very simple, serial interconnection of the reference-current source and the charge having grounded or connected to voltage-supply end. This kind of source assures two good performances relating to the maximum current variation in a given temperature range and to the „supply regulation" parameter. Some total-reference-current-source parameters became worse because of a resistance-value growth, which lead to the increase of circuit minimum supply voltage and occupied-on-chip area. Maybe future researches will guide to the attenuation of this total-reference-current-source disadvantages.

In the present work another cross-connected classical-mirror combination is proposed (not used to date): a modified Wilson mirror (with resistance replacing the diode) and a normal Widlar mirror. The first (having three components) assures implicitly a

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first-order temperature compensation for a certain diode-replacing-resistance value and is simpler than a diode-completed-Widlar mirror (four components, to assure the first-order thermal compensation, [1]). Moreover, the proposed here reference current source can assure a better supply regulation.

Section II establishes the first-order thermal-compensation condition for the total-reference current. Section III includes the deduction of second-order thermal compensation condition. Section IV shows the calculus of first and second-order temperature coefficients of ratio m. The practical source scheme is the section-V object. In section VI the results of simulations of the proposed circuit with output branch are exposed. Finally, section VII summarizes the work conclusions.

II. FIRST-ORDER TEMPERATURE COMPENSATION OF THE REFERENCE CURRENT

Paper [1] presents a reference-current source, composed of a lower modified-Widlar mirror and an upper reverse-Widlar mirror, cross-connected, where second-order thermal compensation is achieved for the left-branch I_1 current (fig.1). The authors established that it is necessary to impose as well the temperature dependence of the two-branch current ratio *m* ($I_2=mI_1$). They affirm that without transistor M_5 ("diode" connected) it is not possible the first-order thermal compensation. Thus, the lower current mirror is a modified-Widlar one.

With the goal of second-order thermal compensation, the reverse-Widlar mirror (M_3, M_4, R_2) achieves a ratio m with positive first-order temperature coefficient. The scheme in fig.1 is named in succession the Fiori-Crovetti source. Unfortunately, the thermal-compensated current I_1 cannot be extracted from the source left branch and used in a charge without affecting the scheme and thus, the wanted thermal compensation. In [1] are deduced the first and second-order compensation conditions for the current I_1.

The proposed new reference-current source is shown in Fig.2. It is composed of lower modified-Wilson mirror (with M_1, M_2, R_1) and an upper Widlar mirror (with M_3, M_4, R_3) which are cross-connected. Without the resistance R_3, the scheme is known classical one [4]. Here, with the help of R_3 it will be achieved the second-order thermal compensation of the reference current I_1.

To obtain the first-order thermal-compensation of the proposed-source current I_1 in fig.2, one establishes here the condition that the resistance R_1 must fulfill. The current I_1 equation can be obtained starting from that written on source lower loop:

$$V_{GS\,1} = I_2 R_1 = mI_1 R_1 \qquad sau \qquad V_{Tn} + \sqrt{\frac{I_1}{\beta_n \alpha_1}} = mI_1 R_1$$

wherefrom one obtains the equation:

$$mR_1 I_1 - \sqrt{\frac{I_1}{\beta_n \alpha_1}} - V_{Tn} = 0 \qquad (1)$$

Here $\beta_n = \mu_n C_{ox}/2$ is the gain factor of NMOS transistors (the same for all transistors, non-considering dimensions); α_1 represents the transistor dimension-ratio W_1/L_1 for M_1; V_{Tn} is the NMOS - transistor threshold voltage (the same for all transistors, non-considering dimensions) and m is the branch-current ratio ($I_2=mI_1$).

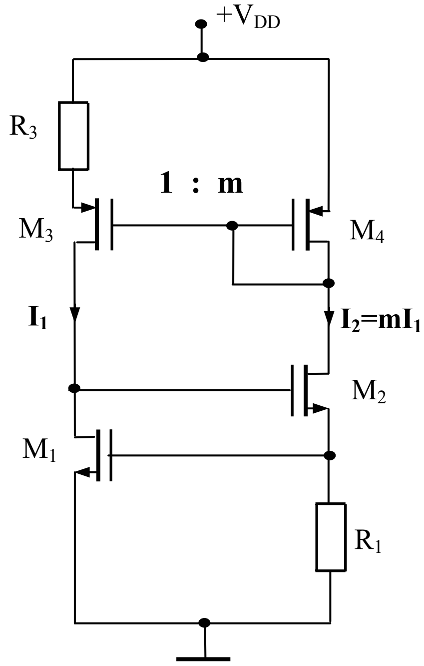

Fig.2. Proposed second-order temperature-compensation branch-current reference

After replacing here the above β_n -factor expression one obtains:

$$mR_1 I_1 - \sqrt{\frac{2I_1}{\mu_n C_{ox} \alpha_1}} - V_{Tn} = 0 \qquad (2)$$

To establish the first-order thermal-compensation condition for the current I_1 in relation (2), noted as f(T), one uses the "total" derivative of the function

$$f(T) = f(I_1, \mu_n, R_1, V_{Tn}, m) = 0 \qquad (3)$$

in which, the five variables are, everyone at its turn, a temperature function:

$$\frac{\delta f}{\delta I_1}\frac{dI_1}{dT} + \frac{\delta f}{\delta \mu_n}\frac{d\mu_n}{dT} + \frac{\delta f}{\delta R_1}\frac{dR_1}{dT} + \frac{\delta f}{\delta V_{Tn}}\frac{dV_{Tn}}{dT} + \frac{\delta f}{\delta m}\frac{dm}{dT} = 0 \qquad (4)$$

Introducing, such as in [1], the (relative) temperature coefficients for five variables, defined in the form

$$k_v = \frac{dv}{vdT} \qquad or \qquad \frac{dv}{dT} = vk_v \qquad (5)$$

(with a variable noted here as "v") and representing the relative variation of that variable with temperature, the following equation is written:

$$\frac{\delta f}{\delta I_1} I_1 k_{I1} + \frac{\delta f}{\delta \mu_n}\mu_n k_{\mu n} + \frac{\delta f}{\delta R_1} R_1 k_{R1} + \frac{\delta f}{\delta V_{Tn}} V_{Tn} k_{VTn} + \frac{\delta f}{\delta m} m k_m = 0 \qquad (6)$$

After the partial derivative calculus in relation (2) and the radical replacement by the expression extracted from the same equation, that is

$$\sqrt{\frac{2I_1}{\mu_n C_{ox}\alpha_1}} = mR_1 I_1 - V_{Tn} \tag{7}$$

the following expression can be obtained:

$$\frac{1}{2}(mR_1I_1 + V_{Tn})k_{I1} + \frac{1}{2}(mR_1I_1 - V_{Tn})k_{\mu n} + \\ + mR_1I_1k_{R1} - V_{Tn}k_{VTn} + mR_1I_1k_m = 0 \tag{8}$$

If imposes here the condition of current first-order thermal compensation, k_{I1}=0, one obtains the necessary value for the resistance R_1:

$$R_1 = \frac{V_{Tn}}{mI_1} \cdot \frac{k_{\mu n} + 2k_{VTn}}{k_{\mu n} + 2k_{R1} + 2k_m} \tag{9}$$

which is simpler that the obtained in [1] one. A great value of m assures a lower value of R_1. In this relation they exist really two unknown variables: R_1 and k_m. If the obtained in simulation R_1 value (for that the variation slope of I_1 current against temperature is minimum) is introduced here one obtains, for the adopted process, a temperature coefficient k_m of negative value. Consequently, to achieve the second-order temperature compensation of current I_1, it must be used an upper mirror of normal Widlar type (M_3, M_4, R_3) unlike the used in fig.1 one, of reverse Widlar type [1].

An important observation is referred to the voltage drop on R_1 which is mI_1R_1 (9). This is done by the product of V_{Tn} and the fraction which is nearly constant (9). So, knowing that $k_{\mu n}$ is negative and has a great absolute value, to reduce the voltage drop on R_1, it is necessary to choose an integrated resistor with very small or negative temperature coefficient k_{R1}. Because the value of resistance R_1 is relatively great it is recommended to use a resistor type having great resistance per square, limiting the chip area.

It must be mentioned that the cross-connexion of a simple current mirror and a modified-Wilson one [4] can establish itself in a zero current state, which imposes the circuit completion by a starting one. But our upper mirror is of Widlar type and the simulations do not signalized the necessity of this completion. Also, it must be mentioned that the try at first-order thermal compensation of current I_2 (instead of I_1), to profit of Wilson-mirror great output resistance, do not has success.

III. SECOND-ORDER TEMPERATURE-COMPENSATION OF THE REFERENCE CURRENT

The establishment of second-order compensation condition for the reference current I_1 in fig.2, pursues the procedure from [1]. So, starting from the first-order temperature coefficient of the reference current in (8), put in the form:

$$k_{I1} = \frac{V_{Tn}(k_{\mu n} + 2k_{VTn}) - mR_1I_1(k_{\mu n} + 2k_{R1} + 2k_m)}{V_{Tn} + mR_1I_1} = \frac{N(T)}{D(T)} \tag{10}$$

the second-order current temperature coefficient will be [1]:

$$k_{I1I1} = \frac{\frac{dN(T)}{dT}D(T) - \frac{dD(T)}{dT}N(T)}{D^2(T)} \tag{11}$$

In relation (11), the numerator N(T)=0 because it is just the first-order compensation condition for the total current [2]. Consequently, after simplification by D(T), from relation (11) one obtains:

$$k_{I1I1} = \frac{1}{D(T)} \cdot \frac{dN(T)}{dT} \tag{12}$$

If now is imposed the second-order temperature-compensation condition, that is k_{I1I1}=0, having D(T) as finite quantity, the following condition results:

$$\frac{dN(T)}{dT} = 0 \tag{13}$$

The calculus of this condition is developed here and it will consider that for the adopted process, in conformity with the transistor and resistor-model-parameters table, the parameters k_{VTn} and k_{R1} do not depend on temperature, thus, for the variables V_{Tn} and R_1 do not exists a second-order temperature coefficients. Also, one will consider that the current I_1 is constant against the temperature, this fact just representing the current-first-order-temperature-compensation condition.

After the evaluation of derivative in (13) and introduction of first and second-order temperature coefficients of variables μ_n, V_{Tn} and m, it result in:

$$V_{Tn}(k_{\mu n} + 2k_{VTn})k_{VTn} + V_{Tn}k_{\mu n\mu n} - mI_1R_1 \cdot \\ \cdot[(k_{\mu n} + 2k_{R1} + 2k_m)(k_{R1} + k_m) + k_{\mu n\mu n} + 2k_{mm}] = 0 \tag{14}$$

Here one used notations with repeated index for second-order coefficients, defined as first-order-temperature-coefficient derivatives against temperature [1], [3].

Now, in relation (14) will be substituted the factor

$$\frac{mI_1R_1}{V_{Tn}} = \frac{k_{\mu n} + 2k_{VTn}}{k_{\mu n} + 2k_{R1} + 2k_m} \tag{15}$$

and one obtains, after some simplifications, the second-order thermal-compensation condition:

$$k_{\mu n\mu n} - (k_{\mu n} + 2k_{VTn})\left(k_{VTn} + k_{R1} + k_m + \frac{k_{\mu n\mu n} + 2k_{mm}}{k_{\mu n} + 2k_{R1} + 2k_m}\right) = 0 \tag{16}$$

In relation (16) the first term is a negative quantity and the parenthesis product gives a positive one. Thus, the second-order temperature-compensation condition will fulfil for a particular pair of values of m and σ. These can be calculated with approximation by repeated trials. With their help can be calculated approximately the resistances R_1 and R_3 values that will be used at start in simulation.

IV. FIRST AND SECOND-ORDER TEMPERATURE COEFFICIENTS OF RATIO m

The first-order temperature-coefficient calculus for the branch-current ratio m in the proposed source (fig.2) follows the procedure from [3] but applied to the normal-Widlar mirror.

Thus, on the M_3 and M_4 input loop (fig.2) it can write the equation:

$$I_1 R_3 = V_{GS4} - V_{GS3} \quad (17)$$

wherefrom, after substitution of voltages against currents and reciprocal reduction of threshold voltages, one obtains:

$$\sqrt{\frac{mI_1}{\beta_p \alpha_4}} - \sqrt{\frac{I_1}{\beta_p \alpha_3}} = I_1 R_3 \quad (18)$$

where β_p is the PMOS-transistor gain factor and α is the W/L dimensional factor of transistors M_3 respectively M_4. The relation (18) is written now:

$$\sqrt{\frac{m}{\alpha_4}} - \sqrt{\frac{1}{\alpha_3}} = R_3 \sqrt{I_1 \beta_p} \quad (19)$$

and then, using the notation $\sigma = \alpha_3 / \alpha_4$, it is put in the form

$$\sqrt{\sigma m} - 1 = R_3 \sqrt{I_1 \beta_p \alpha_3} \quad (20)$$

With this relation, after establishing m and σ variables, one can calculate the resistance R_3 value. Here, it is necessary to remark the fact that, to obtain a positive value for resistance R_3 , it must fulfil the condition:

$$\sqrt{\sigma m} > 1 \quad sau \quad \sigma m > 1 \quad (21)$$

To calculate the first-order temperature coefficient of m ratio, noted k_m, the relation (20) is put in an adequate form to easy calculation of total derivative against temperature:

$$f_1(T) = f_1(I_1, R_3, \mu_p, m) = R_3 \sqrt{I_1 \frac{\mu_p C_{ox}}{2} \alpha_3} - \sqrt{\sigma m} + 1 = 0 \quad (22)$$

Here, the gain factor has been replaced by the known relation

$$\beta_p = \frac{\mu_p C_{ox}}{2} \quad (23)$$

The $f_1(T)$ function total derivative against temperature is written:

$$\frac{\delta f_1(T)}{\delta T} = \frac{\delta f_1}{\delta I_1}\frac{dI_1}{dT} + \frac{\delta f_1}{\delta R_3}\frac{dR_3}{dT} + \frac{\delta f_1}{\delta \mu_p}\frac{d\mu_p}{dT} + \frac{\delta f_1}{\delta m}\frac{dm}{dT} = 0 \quad (24)$$

After partial derivative calculation and replacement of simple derivative with corresponding first-order temperature coefficients, defined as in [1] and [2] in general form (for a variable v):

$$k_v = \frac{dv}{dT} v \quad (25)$$

it results :

$$R_3 \sqrt{\frac{I_1 \mu_p C_{ox}}{2} \alpha_3} \cdot k_{I1} + 2R_3 \sqrt{\frac{I_1 \mu_p C_{ox}}{2} \alpha_3} \cdot k_{R3} + + R_3 \sqrt{\frac{I_1 \mu_p C_{ox}}{2} \alpha_3} \cdot k_{\mu p} - \sqrt{\sigma m} \cdot k_m = 0 \quad (26)$$

Here, the current thermal-compensation condition, k_{I1}=0, is imposed, then the radical appearing in the left part of the relation (20) written as:

$$R_3 \sqrt{I_1 \cdot \frac{\mu_p C_{ox}}{2} \alpha_3} = \sqrt{\sigma m} - 1 \quad (27)$$

is introduced. This results in:

$$\left(\sqrt{\sigma m} - 1\right) k_{\mu p} + \left(\sqrt{\sigma m} - 1\right) 2 k_{R3} - \sqrt{\sigma m} \cdot k_m = 0 \quad (28)$$

and from this :

$$k_m = \frac{\sqrt{\sigma m} - 1}{\sqrt{\sigma m}} \cdot \left(k_{\mu p} + 2 k_{R3}\right) \quad (29)$$

With the obligatory (21) condition the above expression fraction is positive while the parenthesis factor (including the temperature coefficients) is negative for the 0.35µm adopted process. Thus, k_m<0, as such it is necessary for the second-order thermal-compensation of the branch-current source in fig.2.

To do the same compensation for the source of [1] it was necessary a temperature coefficient k_m>0, so that the authors used an upper reverse-Widlar mirror (fig.1).

Based on relation (29) further one will determine the second-order temperature coefficient of current ratio m, defined simply [1] as:

$$k_{mm} = \frac{dk_m}{dT} \quad (30)$$

Thus:

$$k_{mm} = \frac{dk_m}{dT} = \frac{\partial k_m}{\delta m}\frac{dm}{dT} + \frac{\delta k_m}{\delta k_{\mu p}}\frac{dk_{\mu p}}{dT} + \frac{\delta k_m}{\delta k_{R3}}\frac{dk_{R3}}{dT} = = \frac{\partial k_m}{\delta m} \cdot m k_m + \frac{\delta k_m}{\delta k_{\mu p}} \cdot k_{\mu p \mu p} + \frac{\delta k_m}{\delta k_{R3}} \cdot k_{R3R3} \quad (31)$$

where: k_{mm}, $k_{\mu p \mu p}$ and k_{R3R3} represent the second-order temperature-coefficients of m, μ_p and R_3. The coefficient $k_{\mu p \mu p}$ has been established in paper [2] while the coefficient k_{R3R3} is comprised in the table of integrated-resistor model parameters. For the process used in the present work and the N^+ diffusion-sheet-resistance type, k_{R3R3}=0.

Calculating the partial derivative and replacing k_m with relation (29), after some term reduction, the next expression is obtained:

$$k_{mm} = \frac{\left(\sqrt{\sigma m} - 1\right)}{2\sigma m}\left(k_{\mu p} + 2k_{R3}\right)^2 + \frac{\sqrt{\sigma m} - 1}{\sqrt{\sigma m}}\left(k_{\mu p \mu p} + 2k_{R3R3}\right) \quad (32)$$

After the adoption of ratio m [2] and, considering the condition (21) of dimension ratio σ, k_{mm} can be calculated. But, the calculus precision will not be very good because of using typical-parameter values given in model tables for certain transistor-channel dimensions. It is known, in the 0.35µm process, transistor parameters depend moreover on dimensions in the proximity of smaller as few µm values.

V. PRACTICAL SOURCE SCHEME

Considering the current I_1 (fig.1 and fig.2) cannot be used in a charge, the circuit in fig.3 has been introduced. Here, it has been added a supplementary output branch, with transistor M_5 (with similar dimension as M_4) and the resistance R_4 (close of R_3 value) optimized by simulation for a better thermal compensation of current I_o. Thus, the output has been achieved by a normal-Widlar mirror.

Using the current formula for the Widlar-mirror transistor M_3 given in [4] and adapted to scheme in fig.3:

$$\sqrt{I_1} = \frac{1}{2R_3}\left(-\sqrt{\frac{1}{\beta_p \alpha_3}} + \sqrt{\frac{1}{\beta_p \alpha_3} + 4R_3 V_{ov4}}\right) \quad (33)$$

where

$$V_{ov4} = \sqrt{\frac{I_2}{\beta_p \alpha_4}} = \sqrt{\frac{mI_1}{\beta_p \alpha_4}} \quad (34)$$

represents the overload voltage (part of V_{GS4}) for M_4, and, using a similar relation for the current I_o in transistor M_5 (which has a dimension ratio $\alpha_5 = \alpha_4$), the current ratio is obtained:

$$\sqrt{\frac{I_o}{I_1}} = \frac{-\sqrt{\frac{1}{\beta_p\alpha_5}} + \sqrt{\frac{1}{\beta_p\alpha_5} + 4R_4V_{ov4}}}{-\sqrt{\frac{1}{\beta_p\alpha_3}} + \sqrt{\frac{1}{\beta_p\alpha_3} + 4R_3V_{ov4}}} \cdot \frac{R_3}{R_4} = \frac{\sqrt{1+4R_4V_{ov4}\beta_p\alpha_4} - 1}{\sqrt{1+4R_3V_{ov4}\beta_p\alpha_3} - 1} \cdot \sqrt{\sigma}\,\frac{R_3}{R_4} \quad (35)$$

or

$$\frac{I_o}{I_1} = \left(\frac{\sqrt{1+4R_4V_{ov4}\beta_p\alpha_4} - 1}{\sqrt{1+4R_3V_{ov4}\beta_p\alpha_3} - 1}\right)^2 \cdot \sigma\left(\frac{R_3}{R_4}\right)^2 \quad (36)$$

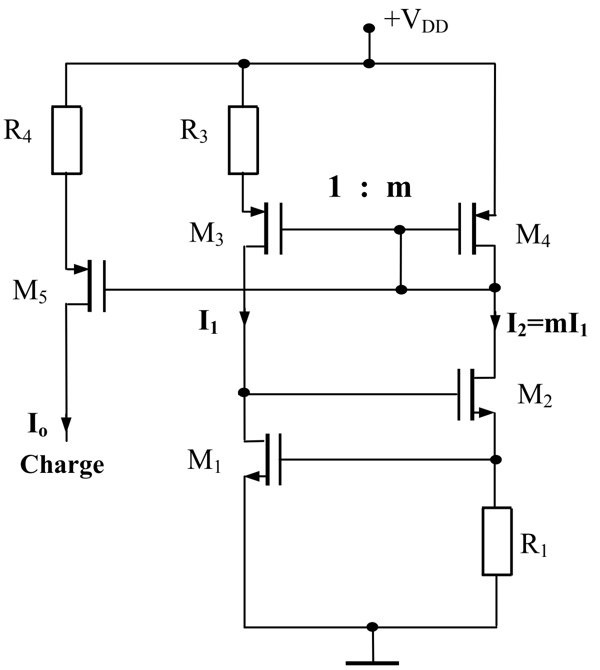

Fig.3. Final reference-current source with output towards the charge with grounded end

VI. SIMULATION RESULTS

The simulation of scheme in fig.2 in view of finding optimal values for transistor dimensions and resistances with the goal of achieving the second-order-temperature compensation of current I_1, has been done starting from the situation obtained after first-order thermal compensation. There have been already established the transistor M_1, M_2, M_4 dimensions and the resistance R_1 value. Then, the W_3 value for transistor M_3 has been found which assures the necessary ratio σ (fulfilling the condition (21)).

Simulations have been achieved, for different m and σ values and for a branch current comprised in the range 8...15µA. The performance results have been similar: the maximum current I_1 variation across the temperature range of 0...100°C is situated around 2.25%, minimum supply voltage is V_{DDmin}=4.7V, the „supply regulation" parameter is SR=12000ppm/V. This results correspond to values: I_1=8.3µA, m=1, σ=6.4, L_1=20µm, $L_{2,3,4}$=5µm, $W_{1,2}$=2µm, W_3=25µm, W_4=3,6µm, R_1=270kΩ and R_3=45kΩ.

Comparing the above obtained performances with the reported in [1] and [3] ones one may see the first are not completely satisfactory. It is especially matter of maximum current I_1 variation.

The simulation of the practical scheme in fig.3, including the seeking of the R_4 resistance optimal value which achieves the output-current I_o second-order thermal compensation, confirmed the relation (36) correctness and leads to superior performances in comparison with the initial one. Thus, having R_4=35.8kΩ (imposed by the thermal compensation condition for the output branch), I_1=8.3µA, a charge with a voltage drop of 1V, for the output current I_o one has obtained the following performances: maximum current variation in the temperature range of 0...100°C situated around 0.9% (fig.4) at a current value of 4.3 µA (imposed by the relation (36)), minimum supply voltage is V_{DDmin}=4.7V, the „supply regulation" parameter is SR=4550ppm/V. These results correspond to situation when M_5 is identical to M_4.

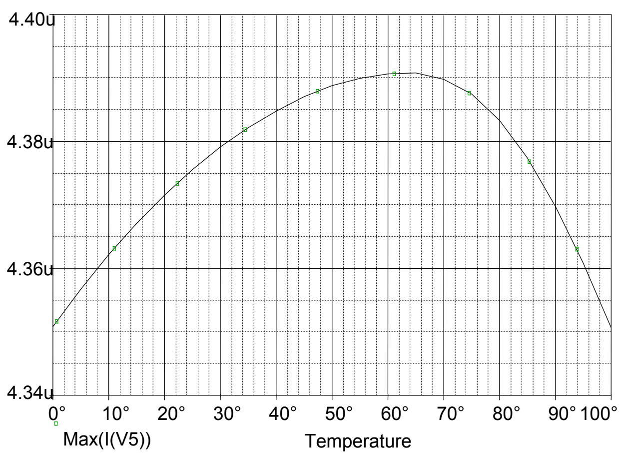

Fig.4. Reference-current variation against the temperature

Another very important parameter for the output current is the "load regulation" and the measured value of this is LR=1550ppm/V, which represents a very good one.

The obtained performance in maximal current variation is 2 times worse than the reported in [3] one and 2.5 times worse than the reported in [1] one, for the same temperature range. In exchange, such as we expected in the Wilson-mirror case, the obtained „supply regulation" parameter in the fig.3 scheme is about the value reported in [1] and 1.7 times better than the reported in [3] one. It can be improved by increasing the transistor cannel length.

The minimum necessary supply voltage for our circuit is relatively great, 4.5V, such as expected [2], [3]. For that reason, the models 5V of transistors have been used in simulation.

VII. CONCLUSIONS

This work analyzed the current first and second-order thermal compensation in a source composed by two usual cross-connected current mirrors, completed by an extra output branch, with the object to simply attach a grounded-end charge. To obtain a second-order thermal-compensation of the reference current it was used a modified source in comparison with known ones: from [1], where is compensated the one branch current, and from [3], where is compensated the total current. With regard to scheme in paper [1], where the reference current cannot be used in the charge without affecting the scheme and the thermal compensation, here a practical solution to convey the reference current in the charge is proposed (fig.3).

In the present work are deduced, by similar methods as in [1], [2] and [3], the first and second thermal-compensation conditions for a branch reference current in the scheme. The first and second-order ratio-m temperature-coefficient formulas for the upper Widlar-type mirror are deduced too.

Having the advantages: circuit simplicity, simplest charge connection to the current reference and good performances, the current thermal compensation brings a disadvantage too: the increase of a resistance of the modified-Wilson current source. This causes two undesirable consequences: the increase of occupied on chip area and the increase of minimum supply voltage (V_{DDmin}). These problems can be solved if a resistor R_1 with very small or negative temperature coefficient is used.

The performance in reference-current maximum variation of 0.9%, in the range of 0-100^{o}C, is 2…2.5 times worse as the realized in [1] and [3] one.

The achieved supply regulation is better than the reported in [3] one and around the reported in [1] one. The charge regulation parameter is discussed only here and has a very good value of 1550ppm/V while in [3] it is the same with the supply regulation and has 3 times worse value.

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