NOVEL NOTCH AND BANDPASS FILTER STRUCTURES USING OTAS AND OPAMPS

Rasit Onur Topaloglu[1]
e-mail: rasitot@boun.edu.tr
esaki.ee.boun.edu.tr/~rasitot/

Hakan Kuntman[2]
e-mail: kuntman@ehb.itu.edu.tr
www.ehb.itu.edu.tr/~kuntman/

Oguzhan Cicekoglu[3]
e-mail: cicekogl@boun.edu.tr
www.elt.boun.edu.tr/cicekoglu.html

[1]Bogazici University, Department of Electrical and Electronics Engineering, Bebek-Istanbul, TURKEY
[2]Istanbul Technical University, Faculty of Electrical and Electronics Eng., 80626, Maslak, Ýstanbul, TURKEY
Fax: +90-212-285 35 65
[3] Bogazici University, Institute of Biomedical Engineering, Bebek-Istanbul, TURKEY
Fax: +90-212-287 24 65

Key words: OTA, operational amplifier, analog filters, active only notch and bandpass filters

ABSTRACT

This paper reports three current-mode filters without any external passive elements that realize notch and bandpass functions. Depending on the circuit, two or more OTAs, and two Op-amps are employed. Also a comparison between implementation with ICs and discrete components is given.

I. INTRODUCTION

Recently there is a tendency designing analog continuous time active filters that do not employ any external passive elements. In these filters the pole of the operational amplifier which gives the active element its integrating nature and the operational transconductance amplifier which can implement the necessary resistance dimension is utilized. One advantage of these active-only current mode filters over voltage mode filters is that they have higher bandwidths. Also, elimination of the passive elements results in reduction of chip area for integrated circuit implementations. On the other hand, having multiple functions in a single circuit, a property circuits 3 and 4 possess, is especially useful since the same topology can be used for two different filter functions. Very few publications exist in the literature on op-amp and OTA only current mode circuits [1-4]. For these type of filters the filter characteristics can be electronically tuned through the transconductance, g_m of the OTAs and/or the compensation capacitor of the opamps. For a second order filter the minimum number of integrators (op-amps) and the number of voltage-to-current relating elements (in classical filters they are resistors but here they are OTAs) is two in current-mode operation. To the best knowledge of the authors , this study is unique in that it offers new topologies in current mode active only notch filters, and that it designs the circuits with both MOSFETs and discrete components. This gives the advantage of implementing the filter either by ICs or discrete components available at hand. Both cases are dealt in this study with SPICE simulations. Simulation results show that filter characteristics are in good agreement with theory.

II. THE OTA AND OPAMP MODELS

Ideally, the OTA is assumed as an ideal voltage-controlled current source. g_m (transconductance gain), used to relate output current to the input voltage, is a function of the bias current, I_A. For the case of OTAs using MOS transistors in saturation the $g_m{}'$s are proportional to $\sqrt{I_A}$; for MOS transistors operating in weak inversion or bipolar transistors the g_m s are directly proportional to I_A. We have used doOTAs (double output OTAs) in our designs. They differ from OTAs in that they have two outputs with separately adjustable g_ms. Current flows in through (+), while it flows out through (-) output nodes of doOTAs in our designs. The OPAMP on the other hand can be modelled by a single pole model, which can be written as B/s for the operating range of frequencies, that is to say, between the first and second poles in the frequency domain. This model of the OPAMP is valid from a few kilohertz to a few hundred kilohertz. In this frequency range a bipolar monolithic OTA works as an ideal device.

III. THE PROPOSED NOTCH FILTERS

The proposed second order notch filters are shown in Fig1 up to Fig4. Fig3 and Fig4 have also the bandpass function built in them. Angular resonant frequency and quality factor denoted by w_O and Q, respectively, are independently adjustable by means of B_1 B_2, the gain-bandwidth products of both opamps, assuming the open-loop gain A(s) has the form of A(s)=B/s. Q factors can further be improved by adjusting the g_m's of OTAs. For the filter, B is used to adjust w_O by means of the compensation capacitors, whereas the OTA g_m transconductance values are used to set the Q value. No component matching constraints are imposed unless you want to have 0db gain for the notch response, which is further and advantage of these filters. The filter transfer functions T(s) =I_O / I_i, are given by the following equations:

<u>Fig1: (notch function, circuit1)</u>

$$T_{BS}(s) = -g_{m2}(B_1B_2 + s^2)/\Delta \qquad (1a)$$

$$\Delta = g_{m1}s^2 + g_{m1}B_2 s + g_{m1}B_1B_2 \qquad (1b)$$

Take g_{m1}=g_{m2} for 0dB gain at passband. You can increase or decrease the pass-band gains by adjusting g_{m2}/g_{m1}. The angular resonant frequency and the quality factor are given by,

$$w_o = \sqrt{B_1B_2} \qquad Q = \sqrt{\frac{B_1}{B_2}} \qquad (1c)$$

The active sensitivities of the circuit are expressed as

$$S_{B_1}^{w_o} = S_{B_2}^{w_o} = S_{B_1}^{Q} = \frac{1}{2}$$

$$S_{B_2}^{Q} = -\frac{1}{2} \qquad (1d)$$

thus all sensitivities are no more than unity.

<u>Fig2: (notch and bandpass functions, circuit2)</u>

$$T_{BS}(s) = g_{m4}(B_1B_2 + s^2)/\Delta \qquad (2a)$$

$$T_{BP}(s) = g_{m3}B_2 s/\Delta \qquad (2b)$$

$$\Delta = g_{m2}s^2 + g_{m1}B_2 s + g_{m2}B_1B_2 \qquad (2c)$$

and the angular resonant frequency and the quality factor are given by,

$$w_o = \sqrt{B_1B_2} \qquad Q = \frac{g_{m2}}{g_{m1}}\sqrt{\frac{B_1}{B_2}} \qquad (2d)$$

Take g_{m2}=g_{m4} for 0dB gain at pass-bands of the notch response and g_{m3}=g_{m1} for 0dB gain at pass-band of the band-pass response. You can further adjust these pass-band gains by g_{m4}/g_{m2} and g_{m3}/g_{m1} for band-stop and band-pass responses respectively. The active sensitivities of the circuit are expressed as:

$$S_{B_1}^{w_o} = S_{B_2}^{w_o} = S_{B_1}^{Q} = -S_{B_2}^{Q} = \frac{1}{2}$$

$$-S_{g_{m2}}^{Q} = S_{g_{m1}}^{Q} = -1 \qquad (2e)$$

thus all sensitivities are no more than unity.

<u>Fig3: (notch and bandpass functions, circuit3)</u>

The angular resonant frequency, the quality factor, the sensitivities and the constraints are same as circuit3, with g_{m4} replaced by g_{m5} and g_{m3} replaced by g_{m4}. g_{m3} can be taken equal to g_{m2}.

IV. SIMULATION RESULTS, DISCUSSION AND DESIGN EXAMPLE

To confirm the theoretical validity of the filter proposed in Fig2, a design example was given and simulated with PSPICE simulation program. The OTAs are realised with the CMOS implementation shown in Fig8 (Figure8). The OPAMPs are realised with the CMOS implementation shown in Fig9. The circuit was supplied with symmetrical voltages of ±5V. The dimensions of the NMOS and PMOS transistors are given in Table1 and Table2. The model parameters used for SPICE simulations are illustrated in Table3. Although we have used doOTAs in our circuits, not all outputs are used. So, we have used single output OTAs with parallel connected inputs to simulate doOTAs for convenience. But if IC implementation is needed, doOTAs should be used to save chip space since less number of transistors are needed for implementing doOTAs if both outputs are used as compared to that of implementing with OTAs.

The filter is designed to realize a filter response with a natural frequency of f_o=102kHz. To achieve this, the biasing voltages of the CMOS OTAs are chosen as 0V and the compensation capacitors of the OPAMPs are taken as 230pF. A GBW of 102kHz is obtained for both OPAMPs with these capacitances. If we have built the filter for a higher frequency, we would need smaller capacitors and we could benefit this property in IC implementations. The biasing voltage of 0V achieves a transconductance gain of 230uA/V. With these biasing voltages and compensation capacitance values the pole quality factor of the filter is obtained as Q=1. The notch frequency is *102k*Hz. The dependence of the OPAMP open-loop voltage gain on the biasing capacitor is obtained with SPICE simulation program and illustrated in Fig4. As it can be seen from Fig4 the gain-bandwidth product are determined as 2.19MHz, 484kHz, 162kHz and 102kHz, for compensation capacitor values of 10pF, 50pF, 150pF and 230pF respectively with a 79dB gain at low frequencies. Fig5 shows the transconductance gain of OTAs for different bias voltages. It is observed that the lesser the g_m is set, the lesser the bandwidth becomes. Fig6 shows the simulated frequency responses of the proposed filter, for various temperatures. The simulation results agree quite well with the theoretical analysis. Fig6 also shows that our FET design works good for higher temperatures than lower ones.

The large signal behaviour of the circuit is tested by applying a 100kHz sinusoidal current (a signal at the pass-band of the band-pass function) to the input and observing the dependence of the total harmonic distortion on the input signal level. The results obtained are summarized in Fig7. It can be observed from Fig7 that the total harmonic distortion remains at acceptable levels below 15uA input voltage where THD is 3.81% and increases rapidly for input voltage levels larger than 17.5uA. These values may be thought of to be acceptable for specific IC implementations. However for better responses, the OTA parameters should be improved, which is a task that is out of the scope of this paper. For an idea, better current mirrors may be employed. Various SPICE transient simulations are run at these values of input levels to test the circuit. In these simulations, different values of resistive loads are tested and the circuit is found to be working good. The notch frequency, thus the tuning frequency of the notch filter can be adjusted by varying the compensation capacitor C. Since Q is dependent on g_m parameters (for circuit2 and circuit3), the tuning of the filter can be done easily by gate voltage of the transistor M6 in the OTA circuits. This property is important since integrated filters must be tuned. Current or voltage-controlled parameters make the filter suitable for on chip tuning techniques. One more advantage of this circuit comes from its nature, that is to say being a current mode type. This gives the filter the ability to work good up to 3Mhz.

Other circuit configurations are not tested with MOSFETs but with ideal models. LM13700 (National Semiconductors) and ua748 macro models are also used to test the BP and notch filter in Fig2. Fig10 and Fig11 shows the notch and bandpass responses respectively designed for 102kHz. Fig10 also shows a comparison between various bias currents. It is observed that the frequency response gets better over the FET counterpart for higher bias currents. This is caused by the BJTs used in discrete components. They can carry more current. A bandwidth of 28MHz is observed for a bias current of 10mA(used for all OTAs in the circuit). Yet for small currents (uA range), FET design gave a better responses (higher bandwidth). Designing with discrete components, you can also change the g_m's of OTAs, limited in between 235uA/V and 75uA/V for our MOSFET design, for a wider range which allows you to increase Q factor in our filters for your convenience. However, those drawbacks against the FET design will be foreshadowed by building better OTAs, and the small chip size used will be an asset for particular applications.

V. CONCLUSION

This paper reports three active-only type notch filter structures two of which simultaneously realise the band-pass responses. The filter structures are easily cascaded since they have high output impedances. The number of active components (two OTAs and two OPAMPs used in circuit1) is minimal for a second order denominator in active-only current mode type of filters.

Acknowledgments: This work is in part supported by Boðaziçi University research found with the project code 01X101.

REFERENCES

1. TSUKTANI T., ISHIDA M., TSUIKI S., FUKUI Y., Current mode biquad without external passive elements, ELECTRONICS LETTERS, vol. 23, no. 3, pp. 197-198, 1996.
2. ABUELMA'ATTI, M. T., ALZAHER, H. A., "Universal three inputs and one output current-mode filter without external passive elements", Electronics Letters, vol.33, no.4, pp.281-283, 1997.
3. TSUKUTANI, T., HIGASHIMURA, M., KINUGASA, Y., SUMI, Y., FUKUI, Y., "A general class of current-mode active-only filters", Proceedings of the International Technical Conference on Circuits/Systems, Computers and Communications, ITC-CSCC'99, pp.297-300, Niigata-Japan, July 13-15, 1999.
4. TSUKTANI T., HIGASHIMURA M., SUMI Y., FUKUI Y., Electronically tunable current-mode active-only biquadratic filter, INTERNATIONAL JOURNAL OF ELECTRONICS, vol. 87, no. 3, pp. 307-314, 2000.

FIGURES

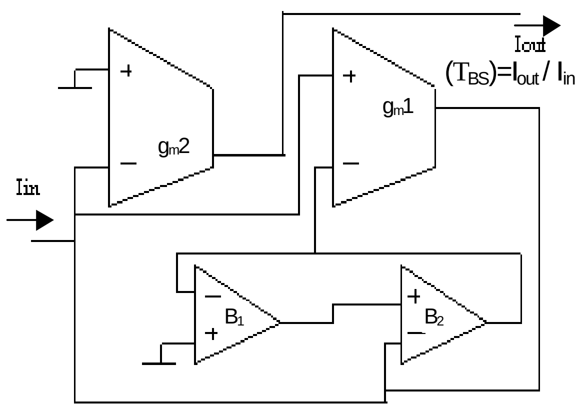

Figure1. circuit1, notch filter

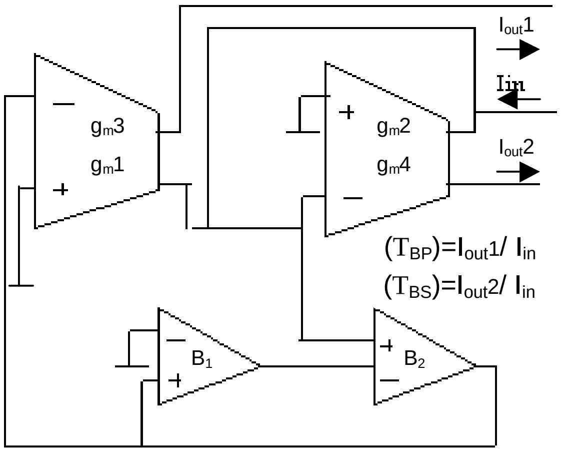

Figure2. circuit2, notch and bandpass filter

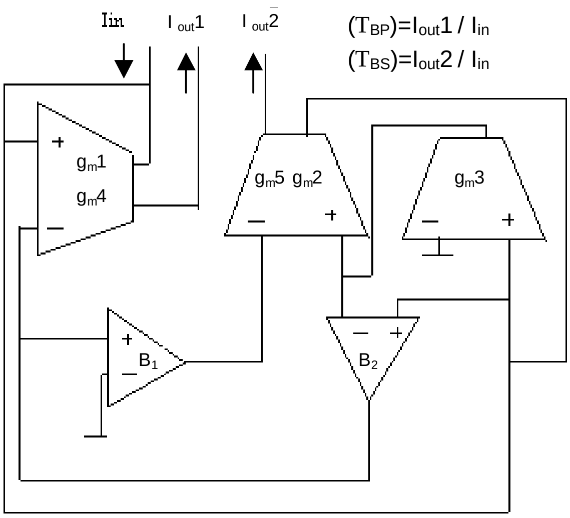

Figure3. circuit3, notch and bandpass filter

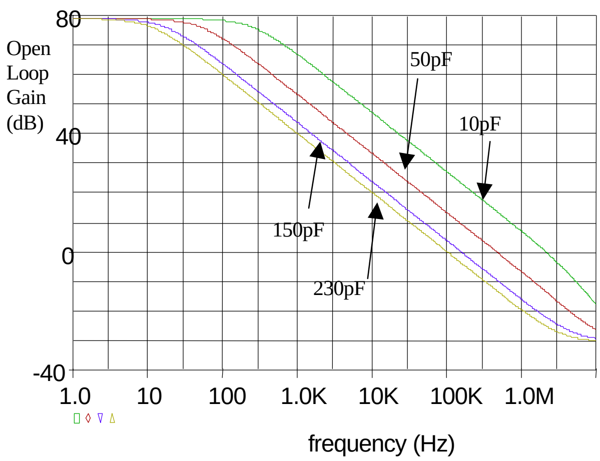

Figure4. dependence of gain bandwidth product on the compensation capacitor

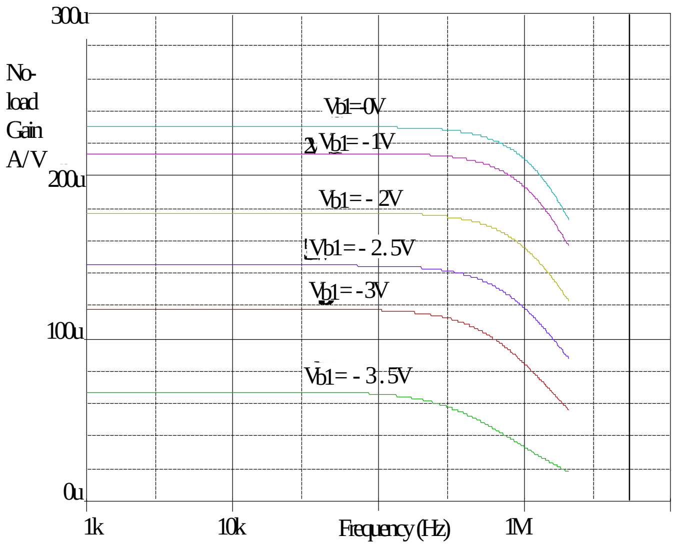

Figure5. frequency response of CMOS OTA for various bias voltages

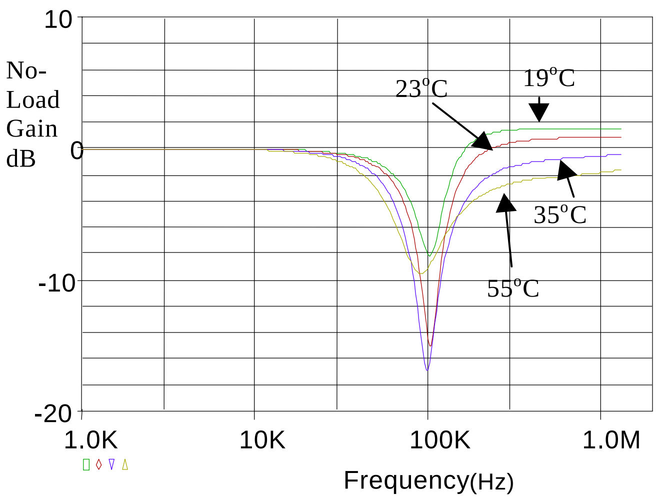

Figure6. circuit2, temperature dependence of the frequency response of notch function

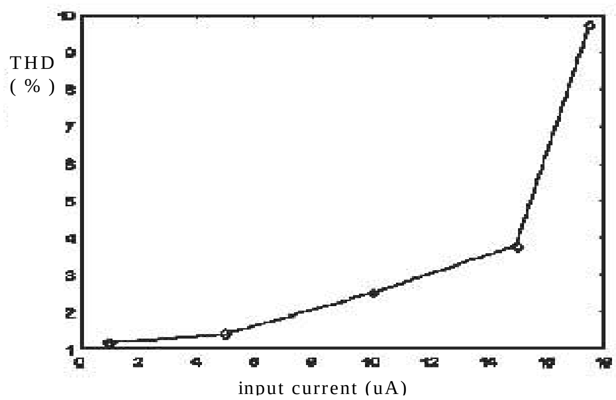

Figure7. Dependence of Total Harmonic Distortion on input current at pass-band. (input:100kHz sinusoidal)

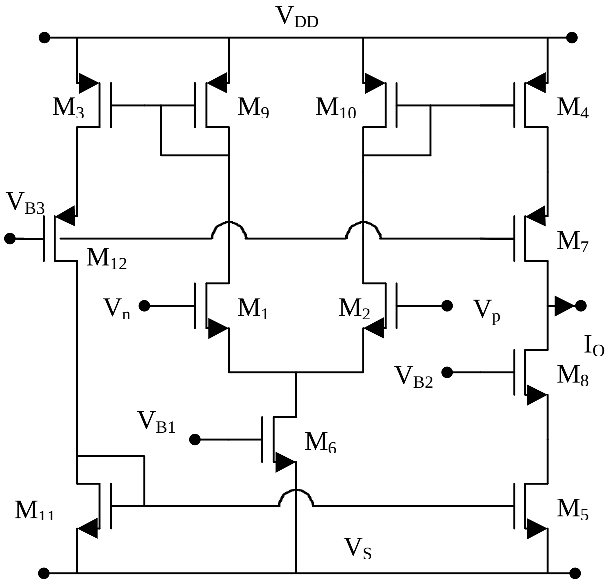

Figure8. CMOS OTA circuit

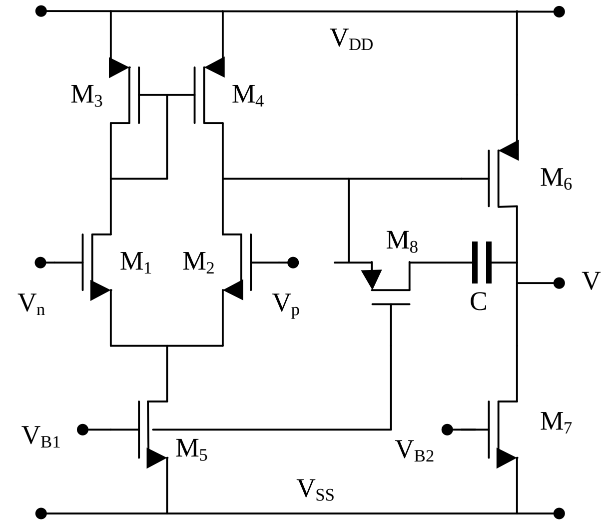

Figure9 CMOS OPAMP

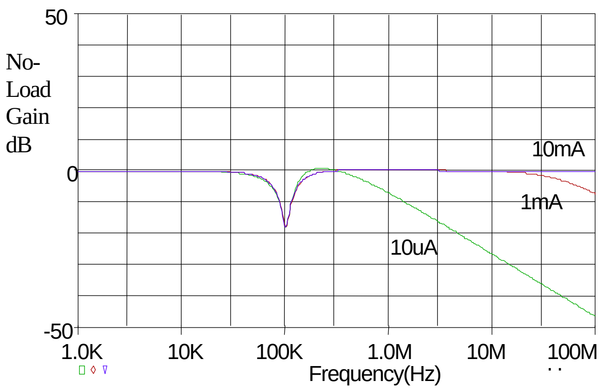

Figure10. circuit 3, notch response for various OTA bias currents

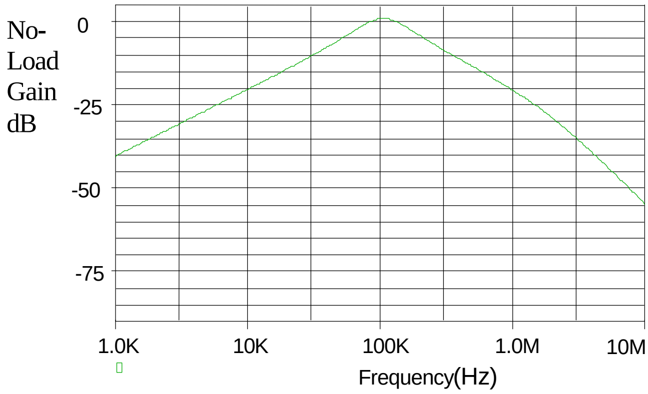

Figure11. circuit 3, proposed bandpass response

TABLES

Table1. Dimensions of transistors used in CMOS OTA

Transistor	L(μm)	W(μm)	Transistor	L(μm)	W(μm)
M1	38	5	M7	2	221
M2	38	5	M8	2	102
M3	10	20	M9	10	20
M4	10	400	M10	10	20
M5	10	237	M11	13	10
M6	21	6	M12	8	10

Table2. Dimensions of transistors used in CMOS OPAMP

Transistor	L(μm)	W(μm)	Transistor	L(μm)	W(μm)
M1	10	180	M5	32	12
M2	10	180	M6	10	373
M3	10	280	M7	10	650
M4	10	280	M8	10	31

Table3. Model parameters of NMOS and PMOS transistors used for SPICE simulations

```
MODEL nmos NMOS LEVEL=2 LD=0.2045U
+TOX=394.00000E-10 NSUB=2.174E+16 VTO=0.8819
+KP=5.081000E-5 GAMMA=0.9693
+PHI=0.6 UO=579.8 UEXP=0.1531 UCRIT=81740
+DELTA=7.67 VMAX=66140 XJ=0.200000U
+LAMBDA=2.2660E-2 NFS=3.91E+11
+NEFF=1 TPG=1.000000 RSH=21.830000
+CGSO=2.6885E-10 CGDO=2.6885E-10 CGBO=3.8386E-10
+CJ=3.9770E-4 MJ=0.4410 CJSW=4.2372E-10
+MJSW=0.338141 PB=0.800000
.MODEL pmos PMOS LEVEL=2 LD=0.2637U
+TOX=394.00008E-10 NSUB=6.803E+15 VTO=-0.7613
+KP=1.8019E-5 GAMMA=0.5422
+PHI=0.6 UO=205.6 UEXP=0.3569 UCRIT=98800
+DELTA=3.331 VMAX=999900 XJ=0.200000U
+LAMBDA=4.612000E-2 NFS=3.23E+11
+NEFF=1 TPG=-1.000000 RSH=70.780000
+CGSO=3.4667E-10 CGDO=3.4667E-10 CGBO=3.6132E-10
+CJ=2.0787E-4 MJ=0.4926 CJSW=1.764000E-10
+MJSW=0.049688 PB=0.800000
```