

**Citation for published version:**

Masood-ul Hasan, Yanqing Zhu, and Yichuang Sun,  
'Oscillation-Based DFT for Second-Order Bandpass OTA-C  
Filters', *Circuits, Systems, and Signal Processing*, September  
2017.

**DOI:**

<https://doi.org/10.1007/s00034-017-0648-9>

**Document Version:**

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# Oscillation-based DFT for Second-order Bandpass OTA-C Filters

Masood-ul Hasan<sup>1</sup> · Yanqing Zhu<sup>2,1</sup> · Yichuang Sun<sup>1</sup>

**Abstract** This paper describes a design for testability technique for second-order bandpass operational transconductance amplifier and capacitor filters using an oscillation-based test topology. The oscillation-based test structure is a vectorless output test strategy easily extendable to built-in self-test. The proposed methodology converts filter under test into a quadrature oscillator using very simple techniques and measures the output frequency. Using feedback loops with nonlinear block, the filter-to-oscillator conversion techniques easily convert the bandpass OTA-C filter into an oscillator. With a minimum number of extra components, the proposed scheme requires a negligible area overhead. The validity of the proposed method has been verified using comparison between faulty and fault-free simulation results of Tow-Thomas and KHN OTA-C filters. Simulation results in 0.25 $\mu$ m CMOS technology show that the proposed oscillation-based test strategy for OTA-C filters is suitable for catastrophic and parametric faults testing and also effective in detecting single and multiple faults with high fault coverage.

**Keywords** Analog circuit testing · Built-in self-test · Oscillation-based test (OBT) · OTA-C filter

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## 1 Introduction

Test and diagnosis techniques for digital systems have been developed and universally implemented during the last three decades. Advances in technology, increasing integration and mixed signal designs demand similar techniques for testing analogue circuitry. Design for testability (DFT) for analogue circuits is one of the most challenging jobs in mixed signal SoC design due to the sensitivity of the circuit parameters with respect to component variations and process technologies. A large proportion of test development time and total test time is spent on the analogue circuits because of the broad specifications and the strong dependency of the circuit parameters. To ensure the testability of a design is an even more formidable task since testability is not well defined within the context of analogue circuits. Testing of analogue circuits based on circuit functionality and its specification under typical operational conditions may result in poor fault coverage, long testing time and the requirement for dedicated test equipment. Furthermore the number of input and output pins of analogue IC relatively small compared to that of the digital circuits, the complexity due to continuous signal values in the time domain, and the inherent interaction between various circuit parameters make DFT for functional verification and diagnosis difficult. Therefore an efficient DFT procedure is required which uses a single signal as input or a self-generated input signal and has access to several internal nodes and the output must contain sufficient information about the circuit under test to enable fault detection and diagnosis. A number of test methods can be found in the literature and various correspondent DFT techniques have been proposed [1-3].

The oscillation-based test (OBT) structure uses vectorless output frequency comparison between fault-free and faulty circuits and consequently reduces test time, test cost, test complexity and area overhead. Furthermore, the testing of high frequency filter circuits become easier because no external test signal is required for this test method. OBT shows greatly improved detection and diagnostic capabilities associated with a number of catastrophic and parametric faults [4, 5].

In oscillation-based testing, the circuit under test is transformed into an oscillating circuit and the frequency of oscillation is measured. The frequency of the fault-free circuit is taken as a reference value. Discrepancy between the oscillation frequency and the reference value indicates possible faults. Fault detection can be performed as a built-in self-test (BIST) or in the frame of an external tester. In BIST, the original circuit is modified by inserting some test control logic which provides for oscillation during test mode. In the external tester, the oscillation is achieved by an external feedback loop network which is normally implemented as part of a dedicated tester. The OBT technique has been applied to various kinds of analogue and mixed-signal circuits [6-15].

As active analogue filters are widely used building blocks in RF, analogue and mixed-signal integrated circuits, significant research has been conducted on low-cost testing of analogue filters. To increase the fault coverage of OBT, [8, 16-18] combined measurement of both the oscillator's frequency and amplitude which

means more test costs. For the ladder filters that can not be partitioned, alternative way to applicate OBT have been addressed in [19-21]. [22] presented a new performance characterization of OBT by using parametric fault models and evaluated the ability of the addressed test strategy for testing two fifth order low pass filters. [23] proposed the extension of OBT into Complex Oscillation Based Test (COBT) with the exploitation of chaotic oscillation regimes and considered its application to band pass (BP) filters. Based on reconfiguration of original circuit to oscillator, [24] proposed a built-in self-test circuit for testing analogue and mixed-signal circuits and considered the proposed solution on testing the typical benchmark circuit of second-order active filter. To extend OBT, [25] used harmonics analysis to improve the fault coverage and the capacity of fault locating. In [26], diagnosis based on OBT was implemented by creating fault dictionary and running artificial neural networks as classifiers. These works were based on the presumption that the amplifying elements within the circuit under test (CUT) perform ideally. To resolve this problem, model related fault dictionaries were developed in [27], and full faults coverage may be achieved only if proper model of the operational amplifiers used and proper feed-back circuit synthesized. Using an on-chip Schmitt trigger as the frequency reference, [28] applied the oscillation-based built-in self-test (OBIST) strategy to low-pass and high-pass analogue filters designed in  $0.35\mu\text{m}$  and  $90\text{nm}$  CMOS technologies respectively. To increase the efficiency of the on-OBIST approach, [29] investigated the optimum value of the oscillation frequency, and focused on fault coverage for R short only.

OBT has also been applied to OTA-C filters. In [30-32], two different methods are proposed to convert lowpass second-order OTA-C filters into oscillators with negligible impact on filter performances. Adopting a second-order Gm-C structure as a case study, [33] presented an OBT scheme that makes use of a non-linear characteristic in the feedback loop and explored the test quality for short circuit and open circuit faults. In order to deal with untuned filters, [34] presented enhanced OBT scheme that performs a relative comparison between two oscillation frequencies and applied the scheme to Gm-C band-pass filter without tuning capability. [35] proposed a novel circuit structure based on a custom Schmitt trigger comparator and evaluated the test quality by adopting a second-order band pass filter as a case.

Oscillation based test has been seldom considered for bandpass filters in the literature, so in this work we describe a low-cost OBT scheme for bandpass OTA-C filters with two filter to oscillator conversion methods. Application of the oscillator-based DFT scheme to two integrator loop Tow-Thomas and KHN biquads are presented, because these structures are commonly used individually as filters and also as building blocks for high order filters.

The paper is organised in the following way. The oscillation conditions of quadrature OTA-C oscillator and two integrator loop networks to oscillator conversion schemes are discussed in Section 2. The bandpass filter into oscillator conversion method is discussed in Section 3 with examples of Tow-Thomas and KHN filters. The simulation results for Tow-Thomas and KHN filters and analysis are presented in Section 4. Finally, conclusion of the paper is given in Section 5.

## 2 Filter to Oscillator Conversions

### 2.1 The Quadrature OTA-C Oscillator

An ideal quadrature oscillator consists of two lossless integrators (inverting and non-inverting) cascaded in a loop, resulting in a characteristic equation with a pair of roots lying on the imaginary axis of the complex frequency plane. In practice, however, parasitics may cause the roots to be inside the left half of the complex frequency plane, hence preventing the oscillation from starting. Any practical oscillator must be designed to have its poles initially located inside the right half complex frequency plane in order to assure self-starting oscillation. Most of the existing theory for sinusoidal oscillator analysis [36] models the oscillator structure with a basic feedback loop. The feedback loop may be positive, negative or a combination of both. The quadrature oscillator model can ideally be described by a second-order characteristic equation:

$$(s^2 - bs + \omega_0^2)V_0(s) = 0 \quad (1)$$

The oscillation frequency  $\omega_0$  can be obtained by first substituting  $s = j\omega$  into (1) and considering real and imaginary parts separately. The oscillation conditions are obtained from the Barkhausen criterion [4]. It states that, at the frequency of oscillation  $\omega_0$ , the signal must transverse the loop with no attenuation and no phase shift. Fig.1 shows a general topology for a second-order quadrature oscillator to establish the oscillation condition by placing oscillation parameter  $b = 0$  and oscillator realization using OTA-C integrators.

The loop in Fig.1 determines the characteristics of the oscillator system. The pole polynomial of the oscillator can be defined as:

$$D(s) = \tau_1\tau_2s^2 + k_{12} \quad (2)$$

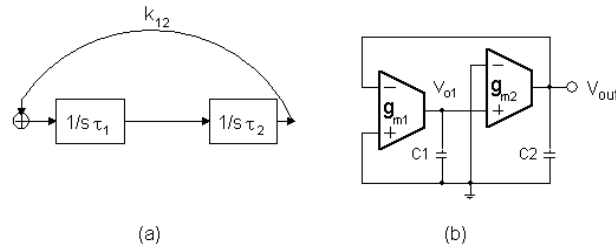


Fig. 1. Second-order quadrature oscillator (a) general topology and (b) OTA-C oscillator.

Oscillator frequency in terms of OTA transconductance gain and capacitors can be determined by comparing (1) and (2):

$$\omega_0 = \sqrt{\frac{k_{12}}{\tau_1 \tau_2}} = \sqrt{\frac{g_{m1} g_{m2}}{C_1 C_2}} \quad (3)$$

Consider the general form of the second order transfer function in (4), where  $\omega_z$  and  $\omega_p$  are the natural frequencies of the zero and the pole, and  $Q_z$  and  $Q_p$  are the quality factors.

$$\frac{Vo(s)}{Vi(s)} = K \frac{s^2 + \frac{\omega_z}{Q_z} s + \omega_z^2}{s^2 + \frac{\omega_p}{Q_p} s + \omega_p^2} \quad (4)$$

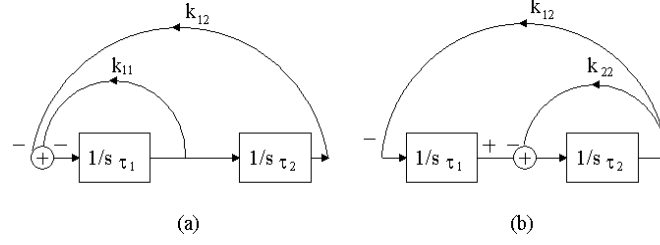
The poles of the transfer function can be expressed in terms of quantities  $\omega_p$  and  $Q_p$ , given by

$$p_{1,2} = \sigma \pm j\omega = -\frac{\omega_p}{2Q_p} \pm j \frac{\omega_p}{2Q_p} \sqrt{4Q_p^2 - 1} \quad (5)$$

To generate oscillation, the poles of the system are initially located close to the  $j\omega$ -axis but in the right-half complex plane. The loop gain is set slightly larger than unity to obtain a sinusoidal signal with an exponentially increasing amplitude so that the circuit is unstable and self-starting. As the oscillation builds up, the inherent nonlinear saturation characteristic of the OTA pulls the poles toward the  $j\omega$ -axis until they finally reach it at some stable value of the amplitude on a limit cycle.

## 2.2 Two-integrator Loop Network to Oscillator Conversions

Two-integrator loop systems have two feedback loops consisting of two ideal integrators and two amplifiers. The summed-feedback and distributed-feedback configurations of two-integrator loop networks are shown in Fig.2.



**Fig. 2. Two-integrator loop (a) summed-feedback and (b) distributed-feedback network configurations.**

The denominator of the transfer function will determine the location and characteristics of poles of the system. The pole polynomial  $D(s)$ , pole angular frequency  $\omega_0$  and quality factor  $Q$  of the summed-feedback configuration can be derived, respectively as:

$$D(s) = \tau_1 \tau_2 s^2 + k_{11} \tau_2 s + k_{12} \quad (6)$$

$$\omega_o = \sqrt{\frac{k_{12}}{\tau_1 \tau_2}} \quad (7)$$

$$Q = \frac{1}{k_{11}} \sqrt{k_{12} \frac{\tau_1}{\tau_2}} \quad (8)$$

Similarly the pole polynomial and parameters of the distributed-feedback system in Fig.2 (b) can be derived as:

$$D(s) = \tau_1 \tau_2 s^2 + k_{22} \tau_2 s + k_{12} \quad (9)$$

$$\omega_o = \sqrt{\frac{k_{12}}{\tau_1 \tau_2}} \quad (10)$$

$$Q = \frac{1}{k_{22}} \sqrt{k_{12} \frac{\tau_1}{\tau_2}} \quad (11)$$

Equations (8) and (11) indicate that the quality factor  $Q$  can be adjusted by  $k_{11}$  or  $k_{22}$  independently without affecting the angular frequency of the poles. Two-integrator networks can oscillate at frequency  $\omega_0$  with constant amplitude if they

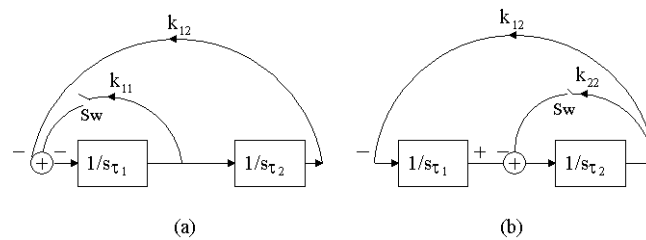
have sufficiently high quality factor. To keep the frequency of the oscillation at  $\omega_0$ , the quality factor should be increased only by changing the values of the components which do not affect  $\omega_0$ . It can be seen from the (8) and (11) that the quality factor  $Q$  tends to infinity without producing any change in the pole angular frequency, if  $k_{11}$  or  $k_{22}$  goes to zero respectively.

We present two techniques to convert a two-integrator loop network into an oscillator by minor modification in the original circuit. The first method uses a switch to break up or open the feedback path  $k_{11}$  or  $k_{22}$  that is, opening the Q-loops, as shown in Fig.3.

The second method is based on the cancellation of  $Q$  factor damping coefficient  $k_{11}$  or  $k_{22}$  by introducing an extra positive feedback path in summed-feedback and distributed-feedback configuration respectively. The two-integrator network into oscillator conversion method using an extra feedback path is shown in Fig.4.

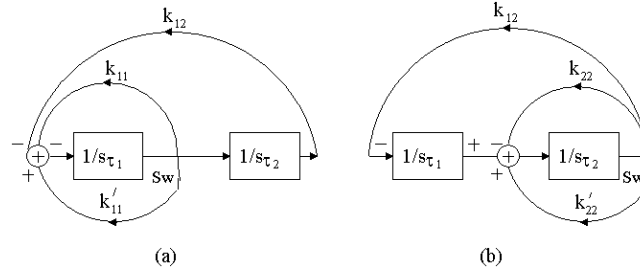
The proposed two-integrator loop network into oscillator conversion methods have their own advantages and disadvantages. The open-loop conversion method requires a minimum number of additional components for modification that is, one switch, and hence acquires small chip area, less circuit complexity and low power consumption. The switch only modification, however, breaks the feedback using a switch in small signal path; therefore the switch resistance and parasitics capacitance will affect the performance of the original system. The second method based on cancellation of damping feedback coefficients requires more components, resulting in relatively large chip area and high power consumption than the switch only method. However the circuit modifications can be directly incorporated at the integrator output node without producing significant effects on the performance and circuitry of the network.

In our previously work [30-32], two different methods are proposed to convert lowpass second-order OTA-C filters into oscillators with negligible impact on filter performances, which use switches only and switches and one OTA, respectively. In this paper, the implementation based on proposed methods are presented for the Tow-Thomas and KHN bandpass second-order OTA-C filters because they are commonly used as low-order filters or basic building blocks for high-order OTA-C filters.



**Fig. 3. Two integrator loop network into an oscillator using switch: (a) summed-feedback (b) distributed-feedback configuration.**





**Fig. 4. Two integrator loop network into oscillator conversion using an extra feedback path: (a) summed-feedback (b) distributed-feedback configuration.**

It is noted that this section has mainly introduced the principles of filter to oscillator conversion. For the converted oscillators to work properly in order to achieve meaningful measurements for testing, nonlinear mechanism must be available for the oscillator to start properly and to be stable. So the whole oscillator is in fact a highly nonlinear system.

Practically, to ensure that oscillation can start, the oscillator must be designed to have its poles initially located inside the right-half complex plane with clearance. The loss of gain in the amplifiers due to saturation will then enable the system to settle on a limit cycle. For the oscillation to be self-sustained, the limit cycle must be stable. The nonlinear saturation effect can be modelled using the well-known describing function.

In OTA-C oscillators, the nonlinear saturation of the OTAs plays the key role in oscillation settling and stabilization. Recently, an improved describing function for OTA-based circuits including OTA-C filters and oscillators has been proposed [41], which may be used in our oscillation-based DFT systems of OTA-C filters.

### 3 Bandpass Filter to Bandpass Oscillator Conversions

Bandpass OTA-C filters can be easily converted into a bandpass oscillator by adding a positive feedback path from its output to the input, as shown in Fig.4. The addition at positive feedback path consists of a feedback coefficient  $k$  from the bandpass output to the input of the filter. As a result, the existing noise of the system is amplified in the loop and produces an oscillating signal. The oscillation frequency is determined by the bandwidth and the gain of the system. If the bandpass filter is tuned to a single central frequency, the oscillating signal will be pure sinusoid and the oscillation frequency will be equal to the centre frequency. Linear feedback is not particularly efficient in producing self-sustained oscillations and a nonlinear circuit is required in the feedback path to guarantee self-starting and self-maintained oscillation. Adding a nonlinear block is a more sophisticated scheme providing better controllability by using the nonlinear saturation charac-

teristic of the OTA. The effect of this is equivalent, in a first order approximation, to decreasing the transconductance gain of the saturated OTA and thus allows us to control the value of oscillator parameter. The oscillation-based DFT structures for OTA-C filters use only output frequency comparison between fault-free and faulty circuits, therefore complicated amplitude stabilization schemes are not necessary.

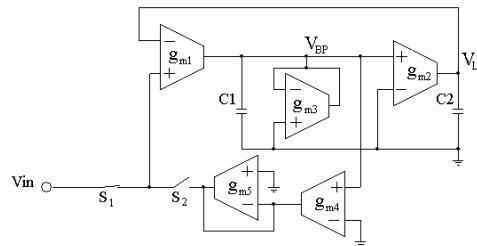
The implementation of the proposed method based on positive feedback for second-order bandpass OTA-C filters is discussed in the following section. The filter to oscillator conversion method is presented for the Tow-Thomas and KHN second-order bandpass OTA-C filters.

### 3.1 Tow-Thomas Filter to Bandpass Oscillator

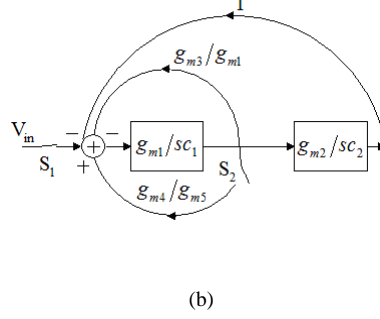
Tow-Thomas (TT) bandpass OTA-C filter can be converted into a bandpass oscillator by using positive feedback from the bandpass output to the filter input. According to Fig.4, the implementation of bandpass oscillation-based DFT in TT OTA-C filter requires two OTAs and MOS switches as illustrated in Fig. 5. The modified filter structure allows us to independently tune the filter centre frequency, gain and quality factor without affecting the amplitude and oscillation conditions. The amplitude of oscillation can be adjusted with the help of transconductance ratio  $k$  of the positive feedback loop.

In the normal filter mode, the switch  $S_1$  is closed and the switch  $S_2$  is opened. The transfer function of the bandpass second order filter can be derived as:

$$\frac{V_{BP}}{V_{in}} = \frac{\frac{g_{m1}}{C_1} s}{s^2 + \frac{g_{m3}}{C_1} s + \frac{g_{m1}g_{m2}}{C_1C_2}} \quad (12)$$



(a)



**Fig. 5. Tow-Thomas filter to bandpass oscillator conversion (a) using two OTAs and MOS switches and (b) topology (Fig. 4(a)).**

The centre frequency  $\omega_0$  and the quality factor  $Q$  are given by:

$$\omega_0 = \sqrt{\frac{g_{m1}g_{m2}}{C_1C_2}} \quad (13)$$

$$Q = \frac{1}{g_{m3}} \sqrt{\frac{g_{m1}g_{m2}C_1}{C_2}} \quad (14)$$

By opening the switch  $S_1$  and closing the switch  $S_2$ , the filter network will be converted into an oscillator. The characteristic equation of the resulting oscillator can be described as:

$$\left[ s^2 + \frac{g_{m3}g_{m5} - g_{m1}g_{m4}}{g_{m5}C_1} s + \frac{g_{m1}g_{m2}}{C_1C_2} \right] V_0(s) = 0 \quad (15)$$

Where oscillation parameter  $b$  is given by:

$$b = \frac{1}{g_{m5}C_1} (g_{m3}g_{m5} - g_{m1}g_{m4}) \quad (16)$$

The above equations show the different expressions for cut-off frequency, quality factor and oscillator parameter  $b$ . It means we can adjust the oscillation condition and amplitude of oscillation by changing any of the transconductances that control  $b$  only without affecting other parameters. The gain,  $k = g_{m4}/g_{m5}$ , of the feedback path should be equal or greater than 1 for guaranteed self-starting and self-maintained oscillation.

### 3.2 KHN Filter to Bandpass Oscillator

Fig. 6 shows the implementation of bandpass oscillation -based DFT in KHN OTA-C filter using the method shown in Fig.4. The modified filter structure can perform the same functions as original filter. The transconductance ratio of the positive feedback path can generate self-starting and sustain oscillation.

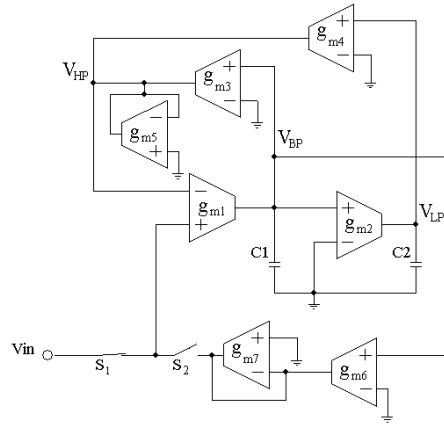
In the filter mode, the switch  $S_1$  is closed and the switch  $S_2$  is opened, and the modified circuit behaves as a normal OTA-C filter. The transfer function of the bandpass second order filter can be written as:

$$\frac{V_{BP}}{V_{in}} = \frac{\frac{g_{m1}}{C_1} s}{s^2 + \frac{g_{m1}g_{m3}}{g_{m5}C_1} s + \frac{g_{m1}g_{m2}g_{m4}}{C_1C_2g_{m5}}} \quad (17)$$

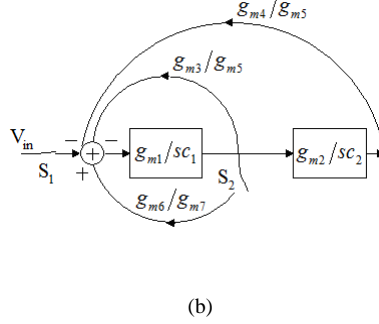
The centre frequency  $\omega_0$  and the quality factor  $Q$  are given in (18) and (19) respectively.

$$\omega_0 = \sqrt{\frac{g_{m4}g_{m1}g_{m2}}{g_{m5}C_1C_2}} \quad (18)$$

$$Q = \frac{1}{g_{m3}} \sqrt{\frac{g_{m4}g_{m5}g_{m2}C_1}{g_{m1}C_2}} \quad (19)$$



(a)



**Fig. 6. KHN OTA-C filter to bandpass oscillator conversion (a) using two OTAs and MOS switches and (b) topology (Fig. 4(a)).**

By opening the switch  $S_1$  and closing the switch  $S_2$ , the filter network is converted into a bandpass oscillator and oscillates with the centre frequency of the filter. The second order characteristic equation of the oscillator is given by:

$$\left[ s^2 + \frac{g_{m1}(g_{m3}g_{m7} - g_{m5}g_{m6})}{g_{m7}g_{m5}C_1} s + \frac{g_{m4}g_{m1}g_{m2}}{g_{m5}C_1C_2} \right] \cdot V_0(s) = 0 \quad (20)$$

Where oscillation parameter  $b$  is given by:

$$b = \frac{g_{m1}}{g_{m7}g_{m5}C_1} (g_{m3}g_{m7} - g_{m5}g_{m6}) \quad (21)$$

with the transconductance ratio  $k = g_{m6}/g_{m7}$ .

Equations (18), (19) and (21) show the different expressions for frequency  $\omega_o$ , quality factor  $Q$  and oscillator parameter  $b$ . Ideally we can tune the oscillator frequency via  $g_{m2}$  and  $g_{m4}$  and adjust quality factor via  $g_{m3}$  for KHN bandpass filter without affecting the amplitude of the oscillation of the converted oscillator. However, when parametric faults or parasitics are considered,  $g_{m6}$  and  $g_{m7}$  must also be tuned to maintain constant amplitude of the oscillation.

## 4 Simulation and Analysis

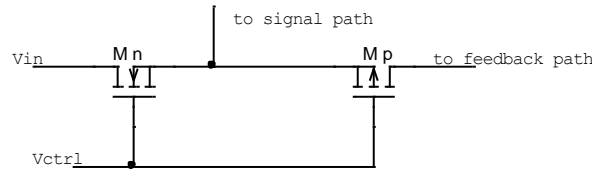
The OBT procedure divides the filter under test into two modes of operation:

1. Filter mode, in which the switch  $S_1$  is closed and the switch  $S_2$  is opened, the system is connected to its normal input.
2. Test mode, in which the switch  $S_1$  is opened and the switch  $S_2$  is closed, the normal input is disconnected and the filter becomes an oscillation.

The system is first tested in filter mode and the cut-off or centre frequency of the filter under test is measured. Then in test mode the filter is converted into an oscillator and the frequency of oscillation is evaluated. The oscillator frequency depends strongly on the transconductances of the OTAs and the filter capacitor values. Deviations in the oscillation frequency with respect to the fault free oscillation frequency of the oscillator indicate faulty behaviour of the filter.

These OBT procedures require certain modifications in the original filter circuits. All these modifications can be carried out by insertion of MOS transistor switches into the original filter circuits. The modified filter circuits shown in previous sections require two types of switches: switches in signal path and switches in feedback path to establish oscillation conditions. The switches in the signal paths are realized using MOS transistors with minimum values of “on” resistance, whereas all other switches are designed for minimum size [31].

Fig. 7 shows the implementation schematic of switches  $S_1$  and  $S_2$  in Fig. 5(a) and Fig. 6(a), which are realised by NMOS and PMOS transistors respectively. These transistor switches are controlled by the DC voltage  $V_{ctrl}$ . When  $V_{ctrl}$  is high,  $M_n$  is short-circuited and  $M_p$  is open-circuited. The system is connected to its normal input and its additional feedback loop is disconnected so it becomes a normal OTA-C filter and works in the filter mode. When  $V_{ctrl}$  is low,  $M_n$  is open-circuited and  $M_p$  is short-circuited. The normal input is disconnected and the additional feedback loop is connected, thus the system becomes an OTA-C oscillator and works in the test mode.



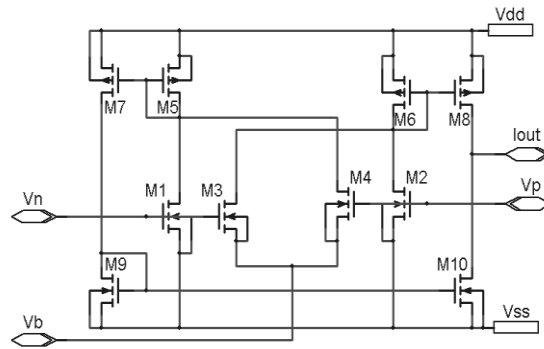
**Fig. 7. Schematic diagram of switches used in simulation.**

The proposed OBT methods were applied to Tow-Thomas and KHN OTA-C filters as examples herein. Case studies were designed for bandpass filters using 0.25 $\mu$ m CMOS technology models available through MOSIS. A simple cross-coupled CMOS OTA shown in Fig. 8 was employed to realize the OTA-C filters. To verify the proposed OBT methods, simulations were performed in PSPICE simulator using BSIM 3 model.

The bandpass Tow-Thomas and KHN OTA-C filters were designed to have a centre frequency of 30MHz. An equal transconductance design was adopted with  $g_m = 740\mu$ S and circuit capacitances  $C_1 = 5.8$ pF and  $C_2 = 3.5$ pF. Simulation results showed that the centre frequency of the Tow-Thomas and KHN OTA-C filter were 29.6MHz, close to the design value. Simulation results demonstrated that the oscillation frequencies for both converted bandpass oscillators are different from their filter centre frequencies. The fault-free frequencies of the oscillators convert-

ed from the Tow-Thomas and KHN OTA-C filters were approximately 23.4MHz with peak to peak amplitude of 1.15V and 21MHz with peak to peak amplitude of 1.4V respectively. This is due to the OTA parasitics affecting the performance of the oscillators. The main OTA parasitics which can affect the oscillation frequency are their finite input/output impedances, frequency-dependent transconductance and other second-order effects. Results confirmed that the self-starting and sustained oscillation was achieved by the circuit modifications described in the previous sections. Simulation results also confirmed that both quadrature and bandpass converted oscillators from each type of filter produced fault-free oscillation frequency close to the frequency of the corresponding filter.

To verify the fault coverage, we have to define and model faults for the filter under test. A fault can be either catastrophic or parametric. Catastrophic faults such as open and short circuit faults cause the total failure of the circuit. They are easy to detect but difficult to locate and correct. Parametric faults are caused by deviation in circuit parameters and manufacturing process. They are more difficult to detect since the circuit can behave in an acceptable manner.



**Fig. 8. Schematic diagram of OTA cell used in simulation.**

We have considered two simple types of catastrophic fault; the stuck-short fault (SSF) and stuck-open fault (SOF). All catastrophic faults are considered to be in transistors, capacitors and interconnections. SSF and SOF are modelled for PSPICE simulation using a low impedance of  $10\ \Omega$  and a high impedance of  $100\text{M}\Omega$  respectively. Parametric faults are inherently more difficult to detect since the filter under test can still function as a filter. The parametric faults to be considered of  $\pm 10\%$  to  $\pm 50\%$  variations in the values of component specifications.

To quantify the fault coverage and the efficiency of the bandpass OBT method for Tow-Thomas and KHN OTA-C filters, the catastrophic and parametric faults were injected in the bandpass filter circuits. A comprehensive list of parametric and catastrophic faults has been also injected into OTAs ( $g_{m1}$ ) at transistor level. The oscillation frequency was measured in test mode and frequency deviation calculated with respect to the corresponding fault-free frequency. The simulation re-

sults of the catastrophic and parametric faults are given in Tables 1 and 2 respectively.

The bandpass oscillators have an important advantage that their oscillation conditions and amplitude of oscillation can be adjusted independently by tuning the transconductance gain,  $k$  of the feedback loop. This means that the amplitude of the oscillation can be stabilized at a constant level under different faulty conditions without affecting the oscillation frequency. The results in Tables 1 and 2 show that all injected parametric faults produced stabilized oscillation with maximum 6% deviation from the fault free oscillation amplitude.

The effectiveness of the proposed OBT methods for OTA-C filters has been demonstrated through extensive simulations using PSPICE simulator and BSIM 3 model. The second-order two-integrator loop, Tow-Thomas and KHN OTA-C filters were used for case studies and analysis of the fault coverage of the OBT methods. Seventy-nine catastrophic and parametric faults were injected in the each testable OTA-C filter, in which fifty faults were parametric and the rest of twenty-nine faults were catastrophic. The fault coverage and the number of undetectable faults are given in Table 3.

**Table 1. Catastrophic fault detections using OBT based on bandpass oscillator**

Faults Comp	TT Filter (Test mode)			KHN Filter (Test mode)			
	$f_{osc}$ (MHz)	Vout(V)	$\Delta f_o/f_o$ (%)	$f_{osc}$ (MHz)	Vout(V)	$\Delta f_o/f_o$ (%)	
C <sub>1</sub>	45.5	1.17	94.44	43.3	1.97	106.67	
C <sub>2</sub>	106	0.35	353	81.4	0.45	288	
C <sub>1</sub> & C <sub>2</sub>	213	1.05	809	95.2	1.97	353	
M3s or d	N.O	-	-	24.4	0.92	16.19	
M4s or d	25.3	1.16	8.12	14.4	1.98	-31.43	
SOF	M3g	N.O	-	25.6	0.38	21.91	
	M4g	25.5	1.16	8.97	14.9	1.98	-29.04
	M <sub>1,6</sub>	N.O	1.18	-	N.O	-	1.11
	M <sub>2,5</sub>	N.O	-1.5	-	N.O	-	-1.48
	M7-10	N.O	1.26	-	N.O	-	1.49
SSF	All*	N.O	S	-	N.O	-	S

SOF=Stuck Open Fault, SSF=Stuck Short Fault, g=Gate, d= Drain, s=Source, N.O=no oscillation,\*=SSF between any two terminals, S=saturated

**Table 2. Parametric fault detections using OBT based on bandpass oscillator**

Comp	Dev	TT Filter (Test mode)			KHN Filter (Test mode)		
		$f_{osc}$ (MHz)	Vout(V)	$\Delta f_o/f_o$ (%)	$f_{osc}$ (MHz)	Vout(V)	$\Delta f_o/f_o$ (%)
C <sub>1</sub>	-50%	29.4	1.20	25.64	27.6	1.52	31.43
	-40%	27.8	1.20	18.80	25.7	1.51	22.38
	-30%	26.5	1.19	13.25	24	1.51	14.29
	-20%	25.3	1.18	8.12	22.6	1.50	7.62
	-10%	24.3	1.17	3.85	21.8	1.45	3.81
	10%	22.5	1.14	-3.85	20	1.39	-4.76
	20%	21.8	1.13	-6.84	19.2	1.38	-8.57
	30%	21.1	1.12	-9.83	18.5	1.36	-11.90



	40%	20.5	1.11	-12.39	18	1.34	-14.29
	50%	20	1.10	-14.53	17.5	1.33	-16.67
	100%	17.7	1.03	-24.36	15.7	1.33	-25.24
	-50%	39.7	1.23	69.66	35	1.45	66.67
	-40%	34.8	1.21	48.72	30.5	1.45	45.24
	-30%	31.0	1.19	32.48	27.6	1.46	31.43
	-20%	27.9	1.18	19.23	24.8	1.44	18.10
	-10%	25.4	1.17	8.55	22.8	1.37	8.57
$C_1 \& C_2$	10%	21.6	1.14	-7.69	19.4	1.37	-7.62
	20%	20.1	1.14	-14.10	18.4	1.37	-12.38
	30%	18.8	1.13	-19.66	17	1.37	-19.05
	40%	17.6	1.12	-24.79	15.9	1.37	-24.29
	50%	16.6	1.12	-29.06	14.9	1.37	-29.05
	100%	12.9	1.10	-44.87	12.1	1.37	-42.38
	-50%	12.9	1.10	-44.87	11.8	1.25	-43.81
	-40%	15.1	1.11	-35.47	13.8	1.31	-34.29
	-30%	17.3	1.12	-26.07	15.6	1.37	-25.71
	-20%	19.4	1.14	-17.09	17.3	1.40	-17.62
$g_m$	-10%	21.4	1.15	-8.55	19	1.42	-9.52
	10%	25.2	1.16	7.69	23	1.39	9.52
	20%	27	1.17	15.38	25.2	1.40	20
	30%	28.7	1.17	22.65	26.2	1.41	24.76
	40%	30.4	1.18	29.91	27.3	1.41	30
	50%	32	1.18	36.75	28.4	1.42	35.24
	100%	39.3	1.18	67.95	33.5	1.50	59.52

**Table 3. Fault coverage and the number of undetectable faults**

Test Method	TIL Filter			TT Filter			KHN Filter		
	CF	PF	FC (%)	CF	PF	FC (%)	CF	PF	FC (%)
Only MOS Switches [30,31]	2	8	87	-	13	83.5	1	6	91
One OTA+MOS Switches [32]	1	3	95	-	2	97.5	1	3	95
Bandpass OBT in this work	-	-	-	-	1	98.5	-	1	98.5

CF = Catastrophic Faults, PF = Parametric Faults and FC = Fault Coverage

Fault simulation results in Table 3 show that the oscillation-based test provides high fault coverage around 98% (since 78 out of all 79 faults injected to the TT filter were detected and only one parametric fault was undetected, and the same for the KHN filter) and capable of simultaneously detecting single and multiple faults.

OBT method requires switch insertion in the small signal path which may degrade filter performance and lower fault coverage due to the less controllability over the oscillation parameters. The positive feedback method using extra OTAs and MOS switches has an advantage of high fault coverage with better controllability over oscillation parameters. Furthermore the modification required for the positive feedback method can be carried out at the filter primary output node without altering the original filter circuits and the performance of the filters. However the extra OTA requires chip area and power consumption, therefore the posi-

tive feedback method has relatively large chip area and power consumption. The bandpass OBT method using two extra OTAs and MOS switches requires more chip area and power consumption than the two lowpass OBT methods. But the bandpass OBT method has the highest fault coverage and controllability of oscillation conditions and amplitude of oscillation due to the inclusion of all filter components in the test circuit and the addition of the positive feedback loop independent of the original filter.

The main ideas behind the proposed OBT strategy consist in converting the filter into an oscillator and then test oscillation frequency. However, because of the technology parameters dispersion caused by the process variations and environmental effects, the oscillation frequency may deviate from the one nominally predicted [34]. When applying OBT to a Gm-C circuit, the tuning scheme becomes necessary. In [38-40], we proposed a simplified tuning circuit and algorithm for bandpass filters. Due to the advantage of minimal additional circuit, the proposed method is suitable for on-chip tuning in OBT.

Finally, the OBT technique may be readily applied to higher order OTA-C filters and other analogue circuits. All these characteristics make the proposed OBT methods very attractive for final production testing as well as wafer-probe testing.

## 5 Conclusions

We have proposed a vectorless, dynamic DFT method for OTA-C filters, based on converting the filter under test into an oscillator using minor modifications in the original filter circuits. The OBT technique requires only measurement of the frequency deviation to detect faults, hence requires very small test time and has good noise immunity. Compared with other OBT methods, the proposed scheme converts the filter under test into a quadrature oscillator with a positive feedback loop only using extra OTAs and MOS switches. The proposed filter-to-oscillator conversion techniques provide better controllability over oscillation conditions, and modifications are carried out at the filter primary output node without altering the original circuits and performance of the filters. The design is easily implemented with little area overhead and has negligible impact on the filter performance. Furthermore, its advantages include a very simple test procedure, the elimination of time consuming specification testing and the requirement of only minor circuit modifications. The case studies of bandpass Tow-Thomas and KHN OTA-C filters show that the proposed oscillation-based test strategy is suitable for catastrophic and parametric faults testing and effective in detecting single and multiple faults with high fault coverage. Although the proposed OBT based topologies for second-order bandpass filters were realised using OTA-C filters, the testing method and topologies can also be used for other types of active filters such as active-RC filters. In the future, our further research will investigate combined measurement of both the oscillator's frequency and other parameters, such as oscillation amplitude and power supply current. Moreover, the detailed study of sizing the

switches and the extra OTAs could be conducted to improve the proposed OBT methods. Test at very high frequencies, more practical simulations and experiment of chip fabrication of the whole DFT system including measurement and control circuits will also be conducted in the follow up work.

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