100MHz, 6th Order, Leap-Frog gm–C High Q Bandpass Filter and On-Chip Tuning Scheme

James Moritz and Yichuang Sun
School of Electronic, Communication and Electrical Engineering, University of Hertfordshire
Hatfield, Herts, AL10 9AB, United Kingdom

Abstract - A 100MHz centre frequency, 10MHz bandwidth 6th order Butterworth bandpass filter using a leap-frog gm-C structure is described. Fully differential Nauta OTAs are used in the design, which include compensation of parasitic output resistance and feed-forward effects to obtain high Q. A mixed-signal, on-chip tuning system based on Dishal's method is described. The filter and tuning system have been simulated using Mosis 0.18mm BSIM 3v3 models and PSpice.

I. INTRODUCTION

High Q bandpass filters are an important technology in traditional RF transceiver design. However, the trend towards highly integrated transceivers has seen architectures developed which do not require high Q filters, in part due to the difficulty of implementing such filters on-chip. Nonetheless, the availability of practicable high frequency, narrow-band filters would increase the options in transceiver architecture available to chip designers. For example, an undersampling A-D converter can directly digitize high frequency signals while minimizing demands placed on subsequent digital processing, but requires an input anti-aliasing filter having a high quality bandpass response, with a bandwidth that is a small fraction of the center frequency. This paper describes the design of a filter that might be used in this application, having 100MHz centre frequency and a 6^{th} order Butterworth response with a bandwidth of 10MHz.

A fully differential OTA-C leap-frog structure is described, based on a Nauta OTA design using feed-forward compensation in order to achieve high resonator Q at high frequency. An important problem with this type of filter is provision of on-chip tuning; while a leap-frog structure offers low sensitivity, coupling between filter sections makes tuning individual sections of the filter difficult. The popular master-slave techniques are limited in accuracy, particularly at high frequencies due to the large effects of parasitics. For this design, we describe a mixed-signal tuning scheme based on Dishal's method to tune individual filter sections.

II. FILTER DESIGN

The prototype Butterworth bandpass LC ladder filter with f_0=100MHz, BW = 10MHz is shown in Figure 1(a). Figure 1(b) shows the OTA-C structure used to simulate the ladder prototype. This circuit is fully differential and uses only grounded capacitors. The circuit can be divided into capacitively-loaded gyrator "resonators"(gm3, 4, 7, 8, 11, 12), and inter-resonator "coupling" OTAs (gm5, gm6, gm9, gm10). The remaining OTAs form the terminating resistances (gm2, gm14), and gain-setting input and output buffers for the filter (gm1, gm13).

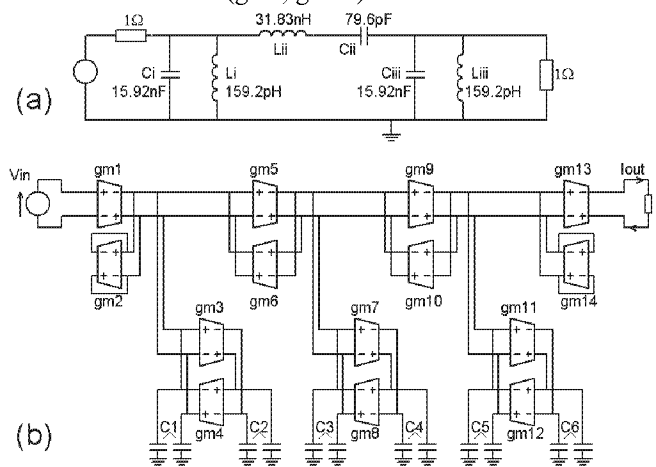

Figure 1. (a) Ladder filter prototype, (b) OTA-C simulation of ladder prototype

One might use impedance scaling to obtain equal-value transconductances and achieve accurate matching between OTAs. However, this results in a wide range of capacitor values, leading to poor matching between resonators. Instead, equal capacitor and transconductance values are used in all resonators. This ensures good matching of the resonant frequency of each section and minimizes the variation in signal amplitudes in different parts of the circuit, helping to maximize dynamic range [1]. This then requires the coupling and terminating OTAs have a large spread in value. However, while errors in transconductance in this case affect bandwidth and passband ripple in the overall filter response, these effects are relatively minor compared to the gross distortions caused by errors in resonator frequency. The scaling process gave rise to the following component values:

C_1-C_6 = 3.18pF
g_{m3}, g_{m4}, g_{m7}, g_{m8}, g_{m11}, g_{m12} = 1mS
g_{m1}, g_{m2}, g_{m6}, g_{m10}, g_{m13}, g_{m14} = 100μS
g_{m5}, g_{m9} = 50μS

III. NAUTA OTA DESIGN

The differential transconductance circuit described by Nauta, Fig. 2, [2] was selected for this filter. This circuit has advantages for high frequency designs using sub-micron CMOS technologies; there are no internal nodes to give rise to parasitic poles, the relatively high output conductance of the MOSFETs can be compensated by choosing transistor sizes appropriately, and large input and output voltage swings are possible at low supply voltage. If the MOSFETs used have a square-law characteristic, non-linearities are completely cancelled; with real, non-ideal devices this is not achieved, but linearity will still be improved compared to simple OTAs. The addition of four more inverters provides control of common-mode gain and output resistance, as shown in the complete Nauta OTA circuit in Fig. 2.

Considering the inverting output, the output of g1 is effectively loaded by a resistance $1/(g_5+g_6)$ for common mode signals, while it is loaded with a resistance $1/(g_5-g_6)$ for differential mode signals. This results in a low output resistance for common-mode signals, and a higher resistance for differential mode signals that can be made positive or negative by adjusting g5 or g6. The same function is performed for the non-inverting output by g3 and g4. The transconductor output resistance, $1/g_{o1}$, is absorbed into the common-mode control circuit by making the transconductance of g6 to (g_5+g_{o1}).

The 1mS transconductors utilize the circuit of Fig. 2. The circuit was designed and simulated using BSIM 3v3 spice models for a TSMC 0.18µm CMOS process available from MOSIS [3]. A semi-empirical approach using these models was used to obtain transistor dimensions for the OTAs. Plotting the gain of the complete Nauta circuit over a range of supply voltages with different sized g3 and g6 revealed the size giving optimum compensation for finite output resistance. All devices in the 1mS OTAs have a channel length of 1µm; the widths are given in Table 1. The coupling OTAs use just g1 and g2 of figure 2, since they share output nodes with the much larger resonator OTAs, and the common-mode gain control sections (g3 – g6 of figure 2) of the resonator OTAs can be slightly enlarged to absorb the output resistance and common-mode gain of

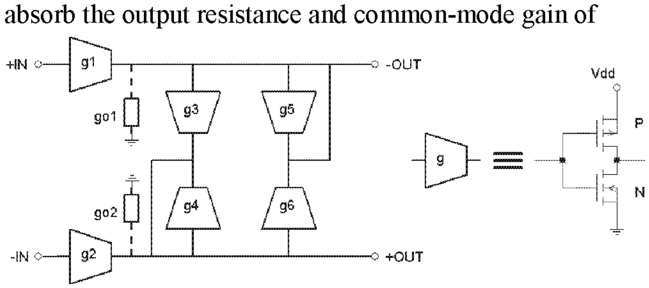

Figure 2. Nauta transconductor circuit

TABLE I. MOSFET WIDTHS

	Inverter (see Fig. 1)		
Width (µm)	g1, g2	g3, g4	g5, g6
W_P	32.9	17.0	16.5
W_N	8.3	4.4	4.1

the small transconductors. The 50µS and 100µS OTAs have W_P and W_N of 3µm and 0.75µm, and 6µm and 1.5µm respectively. Channel lengths are 2µm.Although output resistance compensation in the Nauta OTA circuit allows the DC gain of integrators to become infinite, the high frequency Q possible in a 2-integrator loop resonator using two such compensated OTAs is rather low. The reason for this is the presence of a non-reciprocal feed-forward capacitance, C_{gd}, between gate and drain input and output of the MOSFET inverters making up the transconductors [4]. The reactive current flowing between input and output nodes creates a parasitic zero in the integrator frequency response, resulting in a phase shift of less than 90^0 at high frequency. A feed-forward compensation scheme described by Hughes [5] has been used. Fig. 3(a) shows that a feed-forward current i_{Cff} flows at the output due to the non-reciprocal capacitance; a compensating current i_{Cc} is introduced by the unity-gain buffer and the compensating capacitor C_c which cancels i_{Cff}. The overall resulting parasitic capacitance of $2C_{ff}$ can be absorbed into the integrator capacitors. C_c is made up of the gate-channel capacitance of a PMOS device to achieve good matching to C_{ff}, which is mostly due to the larger PMOS transistors in the OTA. The addition of feed-forward compensation to the filter resonators results in increased Q across the tuning range as shown in Fig. 4.

Although the "coupling" OTAs have much smaller feed-forward capacitance due to their smaller area, they result in excess coupling between adjacent resonators, which is also frequency dependent. This increases passband ripple, and results in an asymmetrical passband about the center frequency.

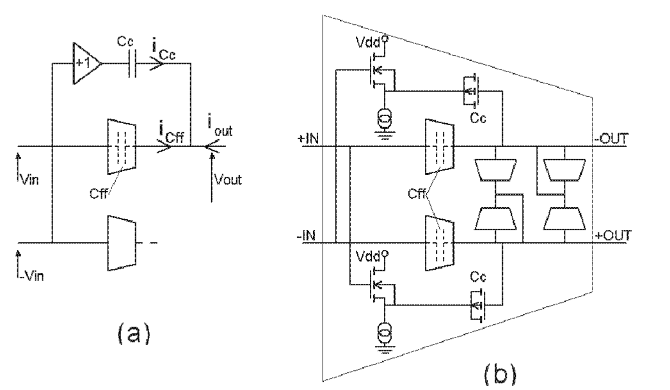

Figure 3. (a) Feed-forward compensation, (b) Compensated Nauta OTA circuit

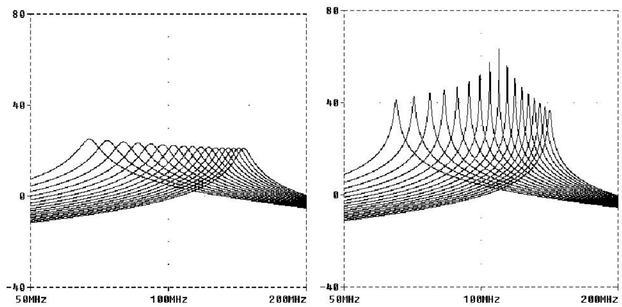

Figure 4. Resonator response (a) without feedforward compensation, and (b) with compensation (amplitude scale in decibels)

A similar feed-forward compensation scheme to the resonator OTAs is utilized to eliminate this effect.

The OTAs are tuned by varying the supply voltage V_{DD}; the nominal transconductance is obtained at V_{DD} of approximately 2.3V. Varying the supply voltage over a range of 1.7 – 3.3 volts provides a tuning range of approximately 1 octave, 70 – 140MHz. Power consumption of the complete filter is approximately 18mW at the nominal center frequency.

IV. Frequency Tuning

Frequency tuning of leap-frog filters is made difficult by the coupling between filter sections; tuning one resonator will move all the filter poles in a complicated way. An analogous problem exists with LC ladder bandpass filters. One technique for tuning ladder filters was described by Dishal [6], and we have previously shown [7] that Dishal's technique can be applied to off-line tuning of leap-frog filters. In this technique, a signal is injected to the filter input at the center frequency, and the amplitude of signals in all filter resonators except the first are forced to zero by shorting them with a MOS switch. The first resonator is tuned to obtain maximum signal amplitude at that resonator node. The short circuit is removed from the second resonator, which is then tuned for minimum amplitude at the first resonator node. The process continues with successive resonators being brought into operation, and tuned to obtain alternately a maximum or minimum in amplitude at the first resonator. A mixed-signal tuning scheme using this technique has been devised in which the tuning voltage is stored as a digital value. This approach is relatively immune to drift problems associated with analogue storage, and requires less power and chip area.

A tuning scheme for a single resonator is shown in Fig 5, with idealized waveforms in Fig. 6. A tuning voltage ramp incremented in small steps is generated by a D-A converter. The AC signal at the resonator is rectified and applied to an integrator. After a tuning step, (i), the integrator output is reset. The reset interval (i) – (ii) is long enough for the transient AC waveform at the resonator to settle to its final value. The AC signal amplitude at the resonator is rectified and integrated for a fixed period (ii) – (iii), after which the integrator voltage is proportional to the resonator signal amplitude. This integrated value is compared with a previous maximum value, which is stored on a capacitor. If the current integrated value is larger, as at point (iv), the current value is transferred to the capacitor via the sample switch as the new maximum value, and the current tuning data is latched into the register. The cycle then repeats at (v), where a further update of the maximum value occurs at (vi). If the current integrated value is smaller than the maximum, as at point (vii), the system proceeds to the next tuning step without further action. After the complete tuning range has been swept, the value in the register is that which resulted in the maximum signal amplitude at the resonator. The tuning voltage for an amplitude minimum can be obtained by inverting the comparator output.

The scheme described allows the maximum amplitude to occur in any particular point of the sequence, so does not require monotonicity from the tuning D-A converter. Since the comparator is only required to detect differences in relative amplitude between different tuning voltage steps, it is not necessary for the envelope detector or integrator to be highly linear, or to have low DC offset voltage. However, it

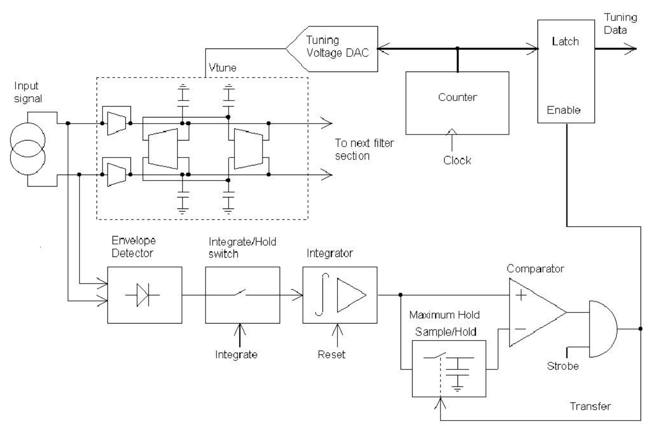

Figure 5. Mixed-signal tuning scheme for a single resonator

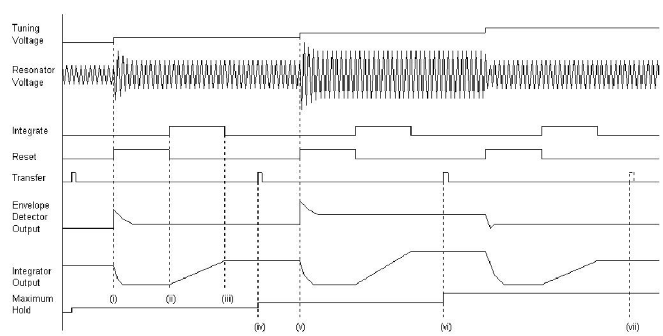

Figure 6. Waveforms for single resonator tuning

is necessary for the comparator and sample/hold to have an offset voltage that is smaller than the difference in integrator output amplitudes between adjacent tuning steps close to the amplitude response maximum.

This tuning scheme can readily be extended to higher-order leap-frog filters with several resonators, using Dishal's method as shown in Fig. 7. Each resonator, A, B, C, has a separate digital-to-analogue converter to generate its tuning voltage. Resonators B and C also include MOSFET switches to reduce the resonator voltages to zero. Because of the inherent matching of on-chip components, and since the resonant frequencies are the same, the tuning voltages for each resonator will be similar. This allows the use of a partial master-slave tuning scheme which reduces the total time taken by tuning. To achieve this, the tuning process is divided into "coarse" and "fine" tuning stages. The tuning voltages VtuneA, B, C are developed by summing the output of a single Coarse Tune digital-to-analogue converter with attenuated outputs from separate During the coarse tuning stage, resonator A is initially tuned for maximum response using coarse voltage steps at VtuneA generated by the Coarse Tune digital-to-analogue converter. The same coarse tuning voltage is applied to VtuneB and VtuneC, which will result in these two resonators being approximately correctly tuned due to matching.

Fine tuning of the individual resonators then proceeds using the Dishal technique. Resonator A is tuned for maximum amplitude using the Fine Tune A digital-to-analogue converter. Switch B is then opened, and Resonator B is tuned for minimum amplitude using Fine Tune B DAC. Finally, Switch C is opened and Resonator C is tuned for maximum amplitude. The frequency response of the filter after completion of the coarse- and fine-tuning steps is shown in Fig. 8.

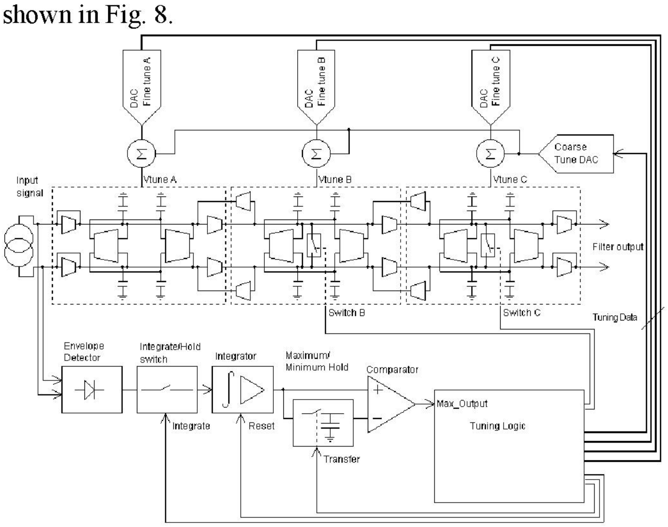

Figure 7. Tuning scheme for 6th order LF bandpass filter

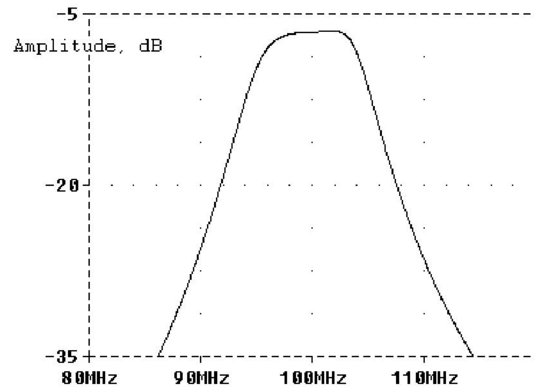

Figure 8. Filter response after tuning

V. CONCLUSIONS

The filter circuit and tuning system above are intended to show the feasibility of achieving relatively high Q, high order bandpass filter responses at high frequencies. An OTA-C filter has been developed that is capable of implementing the intended 6th order Butterworth response with 100MHz centre frequency and 10MHz bandwidth. The tuning scheme described is able to individually tune the filter resonators to achieve the required passband shape.

In the current work, several parts of the tuning system have been simulated as idealised behavioural models in PSpice. Further work is required to realise these as transistor-level designs for practical implementation. The current work does not include Q-tuning; this is required in practice because achieving high resonator Q depends on matching the transconductances of the MOSFETs in the common-mode section of the Nauta OTA circuit to the output resistances of the transconductor part. These quantities are not generally matched. It would be possible to use a master-slave Q tuning scheme with the leap-frog filter structure.

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