ABSTRACT

Analog fully integrated continuous-time filters are reviewed. The important role of such filters in modern telecommunications is outlined, and several techniques for their implementation in mixed-signal chips are described. The article concludes with a brief review of techniques at the research stage, which promise to expand frequency range and reduce power dissipation.

INTRODUCTION

Unbeknownst to their users, today’s digital computers and telecommunications devices contain analog continuous-time filters as necessary and important components. Thus, in wireless systems such filters separate the desired channel from undesired ones and from interference; in high-speed digital links they perform equalization which corrects the phase distortion smearing the digital pulses; they do a similar job inside computer magnetic disk drives, and so on. Indeed, since we live in an analog physical world, analog filters are an indispensable part of a computer’s interface to it, whether this interface is an antenna, a magnetic head, or a transmission line.

The performance of analog continuous-time filters (called simply analog filters in the rest of this article) often represents the main bottleneck in improving the performance of “digital” systems. Figure 1 illustrates a case in point for a wireless communications receiver. An RF bandselect filter passes the appropriate band and applies it to the front-end, which includes a low-noise amplifier, mixer(s), and in some cases additional filters. The input to the analog-to-digital (A/D) converter (ADC) can be a baseband signal or an intermediate frequency signal. In either case, an analog filtering operation precedes the A/D conversion. This essential analog function is necessary to reject undesirable components, which may be orders of magnitude larger than the desired signal. Accomplishing the same job without analog filters would require ADCs and digital signal processors (DSPs) with a much larger number of bits to properly digitize and process large interfering signals, while keeping the quantization noise well below the desired signal. In most practical cases power dissipation constraints and difficulty of implementation prohibit the use of such ADCs and DSPs. Analog filters are needed, in both transmitters and receivers, following mixing operations, in order to reduce undesirable frequency components.

ISSUES IN ANALOG FILTER INTEGRATION

Demanding analog filters have been notoriously difficult to implement on chips. This often makes necessary the use of off-chip filters, and results in penalties in terms of subsystem board size, number of interfacing input/output chip pins, additional overhead circuits, and so on. Ulti-
mately, this increases system cost and power, and lowers design flexibility. Among the difficulties faced in filter integration are the following.

**Frequency response stability:** One must keep the frequency characteristics of the filter stable in the presence of fabrication tolerances and temperature variations, which seriously affect component values. Left on its own, an integrated filter’s frequency response can change by as much as 50 percent or more. This must be reduced to a few percent or, in some cases, to a fraction of 1 percent, by means of automatic tuning. The principle of automatic tuning is shown in Fig. 2a [1, 2]. A filter is placed in a feedback loop, and its frequency response is adjusted until it becomes locked to an off-chip stable reference such as a clock frequency. This is usually available in the system at little or no extra cost and may be derived from the system clock. Another choice of off-chip reference is a low-temperature-coefficient resistor, to the conductance of which the filter conductances are locked. The proper filter control voltage (or current) achieved by automatic tuning is then held while the filter is processing the signal. If the filtering function cannot be interrupted periodically, two identically constructed on-chip filters may be used in time-interleaved fashion between tuning and system operation. As an alternative approach, a “monitor” replica of a basic block of the filter is automatically tuned continuously, and the control signal used for that is also applied to the main filter, as shown in Fig. 2b [1, 2]. Good matching between the filter and the monitor block, an essential requirement for this method, is possible by placing both circuits in physical proximity on chip such that they undergo practically identical fabrication steps and are at practically the same temperature. This natural capability of integrated circuits to accurately match components gives fully integrated designs a fundamental advantage over discrete or partially integrated implementations.

**High-frequency performance:** Parasitic capacitances, as well as the intrinsic delays of transistors, all conspire to limit the maximum frequency of operation of analog filters. For demanding frequency responses, the band-edge frequencies are typically less than 1/100 of the integrated circuit (IC) transistor’s unity-gain frequency \( f_T \) (sometimes referred as the IC technology \( f_T \)).

**Quality of passive components:** Inductors, the workhorse of several types of passive discrete-component filters for over a century, can be made on chips; however, their quality factor (the ratio of their inductive impedance to their loss resistance) is limited. Losses in these inductors are due to their series resistance, as well as the resistance of the substrate, to which they are coupled through electric and magnetic fields. The low quality factor means that highly selective filters, such as band-pass filters with a narrow passband or low-pass filters with a narrow transition band, cannot presently be made in integrated form using on-chip inductors in classical passive filter configurations. Integrated resistors can be made using a variety of techniques, but to allow for flexibility in terms of resistance range, physical size, and linearity, special fabrication steps are needed. High-quality capacitors can be made using well-developed techniques employed to fabricate the insulator (oxide) of metal oxide semiconductor field effect transistors (MOSFETs). Nevertheless, to allow for high-density capacitors, special fabrication steps are needed.

**Signal-handling capability:** For a wide dynamic range, large signal swings are essential. These are becoming increasingly difficult to achieve as power supply voltages are reduced. In addition, for a given power supply voltage, large signal capability implies the use of components that, singly or collectively, exhibit linear behavior.

**Noise:** Resistors and transistors produce
noise, which sets the lower limit of signal amplitudes that can be processed.

**Power dissipation:** Active elements dissipate power. Many of the required filter responses in telecommunications are possible today with integrated filters, but the required high power dissipation may prevent the use of such filters, notably in portable applications where battery life is one of the prime concerns. The causes for excessive power dissipation in today’s integrated filters can be understood from fundamental considerations; see “Power Dissipation and Chip Area” below.

### APPROACHES TO INTEGRATED ANALOG FILTERS

There have been intensive research efforts for the past 20 years to make possible the full integration of analog filters, with significant results. The main techniques used for such filters are illustrated in Fig. 3, using the basic building block — an integrator — in each case.

#### G\textsubscript{m}-C Filters

Figure 3a shows a transconductance-capacitance, or G\textsubscript{m}-C, integrator [2]. Here the input is fed into a transconductor (a voltage-to-current converter), of transconductance (conversion factor) G\textsubscript{m}; the output current of the transconductor is integrated by a capacitor. The unity-gain frequency of this integrator is G\textsubscript{m}/C. To make a linear transconductor possible, a combination of several transistors is used, often along with resistors, so that the effect of nonlinearities of the devices is drastically reduced. Each of the many devices within each transconductor contributes noise, and care must be taken to keep this noise sufficiently low. This type of integrated filter is the most widely used at present. Tuning is made possible by using an array of capacitors or transconductors and connecting an appropriate number of them in parallel through switches, or making the transconductor voltage- or current-controlled, in which case continuous tuning becomes possible. Invariably, the use of continuous tuning brings some loss of maximum signal handling capability. For modest linearity applications, MOSFETs operating in the accumulation region may replace the capacitors, thus allowing implementation in fabrication processes intended for purely digital very large scale integration (VLSI) [3].

#### Active RC Filters

Figure 3b shows the classical active RC integrator. A resistor of conductance G is used for voltage-to-current conversion, and the capacitor is placed in the feedback loop of an operational amplifier (op amp). The integrator unity-gain frequency is G/C. This quantity can be tuned by using an array of capacitors or resistors, along with switches. The noise of the op amp can be made low by design, so the resistor noise is dominant. This single resistor produces less noise than the multitransistor transconductor in Fig. 3a, assuming G\textsubscript{m} = G; thus, for the same capacitance C, the RC integrator has lower noise than the G\textsubscript{m}-C integrator by, typically, a factor of 2 to 3. If highly linear resistors are available on-chip, this type of filter can offer excellent linearity, thus resulting in wide dynamic range. The art of active RC filter design has been developed over half a century, and is well understood [2]. However, many active RC structures not based on integrators, although popular in discrete-component filters, cannot be used on chips. The reason is that the significant parasitic capacitances present at every node in integrated circuits deteriorate the performance, unless such parasitics happen to be connected.
across ports with a well-controlled voltage; it can be shown that such is the case in the integrator of Fig. 3b.

**MOSFET-C Filters**

If the integrator of Fig. 3b is converted to balanced form, and the resistors are replaced by MOSFETs, we obtain the MOSFET-C integrator of Fig. 3c [4]. The MOSFETs are operated in the so-called triode region, where they act as nonlinear resistors. It can be shown that the effect of the even-order nonlinearities of the two MOSFETs, which are the dominant ones by far, cancel out at the output. Thus, unlike the internal circuitry of the transconductor in Fig. 3a, no multiple-transistor nonlinearity cancellation schemes are needed, and the noise performance is similar to that of RC integrators (see above). The unity-gain frequency is again \( G/C \), where \( G \) is the equivalent conductance of the MOSFETs; this conductance depends on the gate voltage \( V_C \), so the latter can be used to make the structure continuously tunable. The linearity is not as high as that of RC integrators, but is adequate in many cases. Here, the advantage is that no resistors are needed. For both this technique and the active RC technique, it can be shown that the op amp may be replaced by an operational transconductance amplifier (OTA), the latter being an op amp without a low-impedance output stage, with a sufficiently large transconductance [5, 6].

Unlike the transconductors in a \( G_m-C \) filter, this element does not need to be highly linear. If the technology allows it, on-chip charge pumps can be used to increase the gate voltage \( V_C \) of the MOSFETs, thus increasing the allowed signal swing [7]. This is easy to accomplish, since the gates do not draw any DC currents. A recent chip which achieves very low intermodulation distortion using this approach is described in [8].

**\( G_m-C \)-Op Amp Filters**

Yet a fourth type of integrator uses a transconductor in lieu of the resistor in an active RC integrator, as shown in Fig. 3d [9]. This integrator (as well as the active RC and MOSFET-C integrators) is less sensitive to parasitic capacitances than the \( G_m-C \) integrator. A comparison of the various techniques, and a discussion of higher-order effects and their compensation, can be found elsewhere [6].

It should be noted that the MOSFET-C topology is fundamentally a balanced one, whereas the \( G_m-C \), active RC, and \( G_m-C \)-op amp techniques can be implemented in either single-ended or balanced form. Often, a fully balanced structure is preferred for these, too, since such a topology is insensitive to parasitic interference coupled into the filter through the chip’s substrate, and through power supply and ground lines.

For complete filters, integrators are used to make larger structures. A popular building block is a biquadratic filter (or biquad), using two integrators to make possible a second-order transfer function. Examples of \( G_m-C \) [2] and MOSFET-C [5] biquads are shown in Fig. 4. The MOSFET-C biquad can implement low-pass, high-pass, band-pass, all-pass, and notch transfer functions by appropriate choice of element values. To achieve this in a \( G_m-C \) filter, additional transconductors are used [2]. High-order transfer functions can be implemented using biquads in cascade. Alternatively, high-order topologies such as leap-frog can be used, which have low sensitivities [2].

Today, all integrated filter techniques presented achieve cutoff frequencies that exceed 100 MHz [3, 10, 11], and are found in large-volume products. Nevertheless, it has not been possible to integrate all analog filters needed in telecommunications applications using such techniques due to rather fundamental reasons, discussed in the following section.

**Power Dissipation and Chip Area**

Currently, the most significant problem of the integrated filter types just discussed is their power dissipation. To understand the issues in this regard [12, 13], consider the integrator in Fig. 3a. Each time a current is needed to charge the capacitor, it must be taken from the power
supply through the transconductor. Each time the capacitor needs to be discharged, the current is dumped to ground, again through the transconductor. Each such cycle of charge drain and dumping represents energy loss. Since for an operating frequency \( f \) there will be \( f \) such cycles per second, the power dissipation is proportional to \( f \), other things being equal. In addition, the power dissipation is proportional to the desired signal power, since increasing the latter necessitates larger voltages and currents. The output signal power, \( S \), equals the desired signal-to-noise power ratio \( S/N \) times the noise output power. The latter can be shown to be proportional to the absolute temperature \( T \). For a filter consisting of such integrators (e.g., the one shown in Fig. 4a), the noise can also be shown to be proportional to the filter selectivity factor \( Q \) [12].

Similar considerations apply to filters using the other approaches in Fig. 3. From such observations one expects the power dissipation of a biquad filter to be proportional to \( f, S/N, T, \) and \( Q \). Indeed, it can be rigorously derived that [12, 13]:

\[
P = a \times m \times f \times (S/N) \times T \times Q,
\]

where \( a \) is a proportionality factor that depends on the filter topology and on how large the voltage swings are relative to the power supply voltage, and \( m \) is the excess noise factor of the (trans)conductance elements; this factor is defined as the ratio of the noise obtained from such an element, to the minimum noise possible for the given (trans)conductance value, as dictated from physics (thermal or shot noise). For active RC and MOSFET-C structures, the excess noise factor is close to 1; for \( G_{m-C} \) and \( G_{m-C-op} \) amp structures, it can be significantly higher, with values of 2 to 3 being common. Thus, a large value for the noise factor \( m \) translates directly to a disproportionate disadvantage in terms of power dissipation, other things being equal. Notice also that each 3 dB improvement in \( S/N \) means proportionally doubling the required power dissipation and chip area. The basic reason for this is that noise power is inversely proportion-al to admittance level; to improve \( S/N \) by 3 dB, one would have to double all (trans)conductances and capacitances, thus doubling the required currents through them and therefore doubling \( P \), and practically doubling the chip area. Comparative studies of \( G_{m-C} \) and MOSFET-C filters give the advantage to the latter (see the references in [10]).

Integrated intermediate frequency and baseband filters for wireless receivers using the above techniques have been demonstrated [8, 14, 15]; their power dissipation is currently well above fundamental limits, due to several nonide-alities. Efforts are underway to improve this.

**ON THE RESEARCH FRONT**

While the techniques described above have reached a rather mature state and are used in mass-volume products, other filtering approaches are currently in the research stage and may have a way to go before they become mature enough for commercial use. A brief review of such techniques follows.

**ON-CHIP LC FILTERS WITH ACTIVE ENHANCEMENT**

As already mentioned, on-chip inductors have large energy losses due to parasitic resistances in the inductor itself and the chip substrate; for this reason, they do not allow the direct implementa-tion of highly selective LC filters. One way to circumvent this problem is to have an active circuit supply enough energy to cancel these losses. The process through which this is done must be carefully controlled by automatic means, and such means must also be provided for frequency tuning. A 1.9 GHz integrated filter using automatic tuning has recently been demonstrated [16]. The performance of such filters is not presently sufficient for use in receivers, but is appropriate for use in some transmitters where it can eliminate an off-chip filter following an upconversion mixing operation. There is a significant IC technology effort to improve on-chip inductor \( Q \)s. As inductors improve, the filter performance achievable with active enhancement will also improve.

**EXTERNALLY LINEAR FILTERS**

A filter intended to process a signal linearly does not have to be linear internally, as long as its input-output behavior is linear. The MOSFET-C filters described above are such an example. Allowing for internal nonlinearities gives more degrees of freedom [17]. The behavior from internal points to the output is not linear, and thus internally generated noise does not follow linear circuit laws. The noise can be made to decrease as the total input signal is decreased, thus extending the input range over which \( S/N \) is sufficiently high. This can be a blessing in some applications (e.g., spectral shaping) but, in its present form, is not appropriate for channel selectivity in many receivers, since a large out-of-band blocker signal can increase the noise level and bury the desired signal in noise [17]. Research is underway to control this effect. In theory, there are many ways externally linear (but internally nonlinear) behavior can be achieved. Companding (compressing and expanding), either instantaneous or syllabic, can be applied to the output. A most important aspect is that noise power dissipation is greatly reduced [18]. Of all the theoretically possible externally linear filters, the class that has received the most attention is that of log domain filters, which are now briefly discussed.

**LOG DOMAIN FILTERS**

In log domain filters [19], an input current signal is converted into a voltage by a logarithmic con-verter (typically, a bipolar transistor), and is then processed internally before it is converted into an output current signal by an exponential converter (another bipolar transistor). Such filters are internally nonlinear, and use transistors in lieu of linearized transconductors. The internal nonlinearities can be shown to cancel out, and the input-output behavior is linear, as long as good device-to-device matching is achieved. Inadequate understanding of the log domain technique has led some engineers to doubt the principle on which it is based; however, a careful study shows that the principle is rigorous [19], and there are no fundamental problems with it. Nevertheless,
there are practical issues which need to be solved to improve the attainable performance. These are the subjects of active research.

**Dynamically Biased Filters**

In a filter with class A circuits, bias currents must be chosen to be at least equal to the maximum anticipated peak signal amplitude, to avoid turning off the active devices. Large currents produce large power dissipation and large noise. When only a small input signal is present, the above power dissipation is wasted, and the large noise unnecessarily diminishes the signal-to-noise ratio. Better performance can be obtained if the bias currents are allowed to vary dynamically depending on total input signal amplitude; then, if the total input signal is small, the power dissipation and noise are reduced. This translates to extended battery life and increased input signal range. However, as was noted in the case of general externally linear filters above, these advantages can be negated if a large out-of-band blocker signal is present. Research is underway to address this problem. A chip that achieves 112 dB input range with a peak S/N of 61 dB was recently demonstrated [20].

**Conclusions**

Analog continuous-time filters are important components of modern communication systems. Integrating such filters onto mixed-signal chips obviates the need for off-chip filters, and provides important advantages in terms of cost, physical size, pin count, and design flexibility. The Gm-C, active RC, MOSFET-C and Gm-C-op amp techniques for integrating continuous-time filters are reviewed here; these have made it into large-volume commercial products such as wireless transceivers and hard disk drives. Issues involving power dissipation and chip area in such filters are discussed. Other techniques, currently in the research stage, are briefly discussed as well. The number of references has been limited due to editorial rules; in the sources quoted, the reader will find many more references to important work in this field.

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**References**


**Biography**

YANNIS P. TSIVIDIS [S’71, M’74, SM’75, F’86] (tsividis@ee.columbia.edu) received a B.S. degree from the University of Minnesota, Minneapolis, and M.S. and Ph.D. degrees from the University of California, Berkeley, in 1972, 1973, and 1976, respectively. In 1975, in order to prove the feasibility of MOS technology for analog and mixed-signal integrated circuits, he designed and built a fully integrated MOS operational amplifier and demonstrated its use in a PCM codec. Since that time, his work has focused on device and circuit issues in the merging of analog and digital circuits on the same chip. He is the Charles Batcherlor Professor of Electrical Engineering at Columbia University, New York. He has worked for Motorola Semiconductor and AT&T Bell Laboratories, and has taught at the University of California, Berkeley, MIT, and the National Technical University of Athens, Greece. He is co-recipient of the 1987 IEEE Circuits and Systems Society Darlington Best Paper Award. He is the recipient of the 1984 IEEE Baker Best Paper Award, the 1986 European Solid-State Circuits Conference Best Paper Award, and the 1989 IEEE Circuits and Systems Society Guillemin-Cauer Best Paper Award. He has received the Great Teacher Award at Columbia University, and a Golden Jubilee Medal from the IEEE CAS Society.