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Analog integrated filters or continuous-time filters for LSI and VLSI

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Résumé. — On donne un résumé sur les filtres (en temps continu) pour la (V)LSI. Les méthodes les plus importantes, leur implantation sur silicium (bipolaire et MOST) ainsi que les applications sont présentées. En conclusion, on présente les limitations et difficultés.

Abstract. — A survey is given on analog (continuous-time) filters for (V)LSI. The most important design methods, the filter elements on silicon (bipolar as well as MOS) and applications are briefly reviewed. Limitations and challenges to designers conclude the survey.

1. Introduction.

Information processing is being done more and more by digital means. The transmission of information (analog as well as digital) is and will remain an analog issue. Between the analog outside world and the digital processors there are interfaces. They are of a mixed type (Fig. 1). Buffer circuits, analog-to-digital (A/D) and digital-to-analog (D/A) converters, coder/decoder combinations (codecs), modulator/demodulator combinations (modems), line interfaces, receivers and transmitters are examples of interfaces.

Transition to VLSI involves shrinking of the digital processors as well as of the analog pre- and post-processors (interfaces). The trend is to combine all

parts on one chip. Digital and analog on one chip can be done in bipolar processes as well as in MOS processes, but bi(C)MOS processes are really optimum for the combination.

Many analog processors are chips with integrated amplifiers, modulators, oscillators, etc., often with external selectivity (external resistors, capacitors, coils, transformers, crystal, crystal filter, SAW filter, etc.). The versatility of the analog processors is due to the fact that different external components (component values) can be chosen.

Chips with integrated selectivity often have fixed filters. With the integration of the external components the versatility is lost. This can be overcome by using controllable and/or programmable integrated circuits. VLSI requires (digitally) programmable analog processors (BUS-controlled analog processors). Variability and programmability are important features for integrated filters.

Digital filters are found in the digital processors. Analog circuits (including filters) are found in the interfaces to the analog outside world. Sampled-signal filters (CCD filters, switched capacitor filters) are found in between. Where the analog/sampled-signal/digital divisions are located depends on

- costs (chip area),
- power consumption and
- performance,

(Fig. 2).

Some examples can be given. Anti-aliasing filters are analog. Noise-shaping filters can be analog, sampled-

Fig. 1. — Mixed analog/digital interfaces between an analog outside world and a digital processor. Processors and interfaces are put on one chip. We go to (V)LSI.

Fig. 2. — Analog filters, sampled-signal filters and digital filters. All have their advantages and disadvantages. The choice is made after considering cost (chip area), power consumption and performance.

signal in type or digital. Receiver selectivity will mainly be analog. Switched-capacitor filters will be somewhat more accurate than continuous-time filters. Image processing will be digital. Analog is preferred for high frequencies and low power. Interference across neighbouring circuits is low for analog filters. For low frequencies digital processing (time sharing) often gives the best solution.

(V)LSI also means using simple circuits and methods. Design time should be as short as possible. The probability of a first correct design should be as high as possible. Even a complex filter takes up only a small part of the chip. Implementation of (V)LSI generally means using lower supply voltages (i.e. changing from 18 V and 12 V to 5 V, 3 V and even lower) and to lower supply currents per transistor.

This survey deals with analog integrated filters. We start with the most useful design methods for integrated filters, make a survey of filter elements on silicon, consider applications and conclude with limitations and challenges facing further design.

2. Methods for filter design.

Conventional analog filters are made from resistors, inductors, capacitors and transformers (or mutual inductances): R-L-C-T filters. Filters are designed as lossless ladder networks between resistive terminations. The coils and mutual inductances cannot be integrated.

In integrated circuits we have transistors instead. In network theory (filter theory) transistors are not considered to be elements (as R, L, C and ideal transformer). Network theory « idealizes » the transistor. Two idealizations are :

- the « VCCS » (Voltage-Controlled Current Source) or « transconductor » :
a two-port with zero input current, the output

current of which is proportional to the input voltage, and

- the « nullor » (also idealized operational amplifier) :

a two-port element with zero input voltage and zero input current [1].

On the other hand, electronics uses combinations of transistors to make better approximations to the idealizations and to design different elements.

Simple R-C-nullor filter types, which are in common use, are

- Sallen and Key filters (Fig. 3).

They are derived from a passive low-pass filter by inclusion of an amplifier with finite voltage gain (for design tables, see [2]).

- Cascade of second-order sections (Fig. 4).

Fig. 3. — Sallen and Key type filters. Addition of an amplifier (with finite voltage gain) to a passive R-C low-pass filter shifts the poles of the transfer function away from the negative real frequency axis. Simple (all-pole) filters can be made. A compensation is shown for the first-order roll-off with frequency of the amplifier gain.

Fig. 4. — Cascade of second-order sections. Factorization of the transfer function in (complex conjugate) pole pairs (and zero pairs) opens the way for implementation as a cascade of second-order filter sections. Cauer-elliptic types of filter can be made.

The transfer function is written as a product of second-order transfer functions and is implemented as a cascade of second-order filters (for design tables, see [3]).

The high sensitivity of the filter characteristics (e.g. attenuation) of the above filters to tolerances on the element values has led designers to return to the classical filters. Their properties are appreciated as never before [4-6], see figure 5. Nowadays, the classical lossless ladders between resistive terminations are simulated in many ways (not only in analog filters, but also in switched-capacitor filters and even in digital filters).

Fig. 5. — Lossless ladders between resistive terminations show important properties: zero first-order sensitivity at attenuation zeroes, high attenuation peaks in the stop band, high far-off attenuation (addition of attenuations of all ladder sections). The above properties lower the requirements on the tolerances of the filter elements. They should be preserved in RC-active design methods.

Fig. 6. — Direct simulation of inductances (high-pass filter). Starting from an L-C prototype filter, all inductances are replaced by gyrator-capacitor combinations. One transconductor is needed to make a resistance, two transconductors yield a gyrator (see lower part of the figure). We arrive at a transconductor-capacitor filter.

We mention some of the methods :

- inductance simulation (Figs. 6-8),
- super-capacitor types of method (Figs. 9-10), and
- signal-flow graph methods (Figs. 11-13).

Inductance simulation is done with gyrator-capacitor combinations. The gyrator itself is not an element in integrated circuits. It can be put together from transistors and resistors and be made in various ways.

One way is to implement the two voltage-to-current relations of the gyrator by transconductors. Two transconductors are used for one gyrator. One transconductor is used for a resistor. We arrive at transconductor-capacitor filters. In figure 6 an example is shown of a fifth-order high-pass filter, in figure 7 we show a similar low-pass filter and in a figure 8 some all-pass sections

Fig. 7. — Direct simulation of inductances (low-pass filter). Starting from an L-C prototype filter, all inductances are replaced by gyrator-capacitor combinations. Transconfigurations yield an equivalent transconductor-capacitor circuit with all transconductors input-earthed (the latter configuration is easier to check for unwanted latch-up).

Fig. 8. — Gyrator and transconductor methods (all-pass). The first-order and second-order gyrator all-pass sections above have no L-C counterpart. They show a delay in one direction and zero delay in the opposite direction (nonreciprocal). The transconductor-capacitor version has input-earthed transconductors. The sections can be inserted in any filter in front of the load resistor (constant-resistance all-pass sections).

(see [17]). The all-pass sections are nonreciprocal and have no LC counterpart. Their constant-resistance character permits them to be inserted in any filter in front of the load resistor.

Alternatively, the gyrator (or better, Positive Immittance Inverter (PII)) can be made from nullors and resistors.

We have the following possibilities :

- 1 nullor, 6 resistors (see [8])
(lossless by compensation),
- 2 nullors, 4 resistors (12 possibilities)
(lossless, independent of element values),
- 3 nullors, 2 resistors (20 possibilities)
(lossless, independent of element values),

(see also [9] and [10]).

The super-capacitor types of method are all based on resonators with two nullors and four resistors (and two capacitors).

Fig. 9. — Filter with (R — operational amplifier) gyrators (high-pass). The high-pass filter has earthed inductors. An earthed inductance can be simulated by an electronic circuit with 2 operational amplifiers and 4 resistors (a gyrator) and a capacitor. The inductance (value and losses) can be made (to a first order) independent of the high-frequency roll-off of the amplifiers.

In the high-pass filter in figure 9 all inductors (earthed) are simulated by (R-operational amplifier) gyrators and capacitors. The losses of the gyrator can be made largely independent of the first-order frequency roll-off of the operational amplifiers (gain-bandwidth product) [11]. Low-pass filters need an extra transformation (Bruton transformation) to get all electronic components connected with one side to earth (Fig. 10). The transformation makes the capacitors « super-capacitors ». Matching the operational amplifiers can provide compensation for the effects of the first-order frequency roll-off of the two amplifiers [11]. For (second-order) all-pass sections we refer to [12], for band-pass filters to [13] and [14].

Signal-flow graph (SFG) methods generally start from an L-C prototype filter (or from a transfer function) from which a signal-flow graph is derived. In the graph we have elementary operations : additions, subtractions, integrations, etc. It is implemented, just

Fig. 10. — Filter with super-capacitors (low-pass filter). In an L-C low-pass filter we have floating coils. Multiplication of all admittances by the complex frequency P (Bruton transformation) leaves the (voltage) transfer function invariant. Inductors are transformed to resistors, resistors to capacitors and capacitors to super-capacitors. The super-capacitors are connected to earth. They can be made with combinations of 2 operational amplifiers, 3 resistors and 2 capacitors. High-frequency roll-off effects can be compensated when the amplifiers are matched.

as in analog computers, using operational amplifier inverters and integrators. The first-order phase errors of the (nonideal) operational amplifiers can be cancelled [11]. We shall give some examples.

Figure 11 shows the classical example of an all-pole L-C low-pass filter and the corresponding signal-flow

Fig. 11. — Signal-flow graph filter (all-pole low-pass filter). A (leapfrog type) signal-flow graph is derived from the L-C prototype filter. It uses addition and integration as operations. The corresponding circuit is shown in the lower part of the figure. The phase errors of the positive and negative integrators can cancel (to a first-order approximation).

graph (leapfrog arrangement). It is implemented in the lower circuit with (phase-error-matched) positive and negative integrators. When we replace the integrators by differentiators we arrive at the corresponding high-pass filter. Insertion of resonators (Fig. 12) leads to band-pass filters (see also [15] and [16]). Figure 13

Fig. 12. — Signal-flow graph method (for band-pass filters). From the L-C prototype filter we derive a leapfrog type signal-flow graph. The integrations (see Fig. 11) are replaced by immittance functions of L-C resonance circuits. The integrators are replaced by electronic resonators.

Fig. 13. — Signal-flow graph filter (low-pass filter). The transmission zeroes (at finite frequencies) can be included in the design procedure by replacing the floating capacitances by pairs of earthed transcapacitances. We arrive at a circuit with inductive series branches and capacitive shunt branches. The signal-flow graph can be drawn. It is implemented as indicated in the lower part of the figure (compare with Fig. 11).

shows a more sophisticated example of a low-pass filter with transmission zeroes at finite frequencies.

3. Filter elements on silicon.

The main problems of integrating analog filters are that

- coils cannot be integrated,
- common integrated resistors have large temperature coefficients,
- integrated elements show high initial tolerances (deviations of the mean value from the design value).

Coils are replaced by capacitors and active circuits. Matching of elements on a chip and the application of variable (controlled) circuits solve the last-named problems.

In integrated circuits the resistor and the capacitor are elements of filter design in themselves. The transistor is not. Combinations of transistors and/or resistors (components) are used. Simple high-performance combinations are optimum for VLSI. Elements and components are combined to form filters.

We mention the following :

— elements	— inductive	— not on silicon,
	— capacitive	— junction capacitor,
		— dielectric capacitor,
	— resistive	— metal-film resistor,
		— diffused resistor,
		— transistor (trans)conductance,
— components	— nullor (operational amplifier),	
	— VCCS (transconductor),	
	— gyrator,	
	— NIC (negative immittance converter),	
	— integrator,	
	— resonance circuit,	
— filters	— low-pass, high-pass, band-pass (-stop),	
	— all-pass, (tapped) delay line.	

3.1 CAPACITORS. — Junction capacitors are present in all processes, gate-oxide capacitors are present in all MOS processes. A little extra effort creates capacitors between layers of interconnect (for switched-capacitor filters). The dielectric is either silicon-oxide or silicon-nitride (or aluminum oxide). Large dielectric capacitors (oxide-nitride sandwich) can be made on monolithic low-ohmic silicon, see figure 14 and [7]. Advanced etching techniques (U-groove) can significantly increase the effective area of capacitors (trench capacitors). It has been reported that 0.2-um-

Fig. 14. — Capacitor types. From the left to the right — capacitor type between two layers of interconnect (as commonly used for switched-capacitor filters) — capacitor between one interconnect layer and a low-ohmic diffusion — a junction capacitor (between 2 low-ohmic diffusions) — a trench type capacitor as has recently been made for application in memories. Some typical capacitor values have been indicated.

wide and 3- μ m-deep trenches filled with a 25-nm thin layer of silicon oxide and poly silicon have been made (for application in RAM's) [17].

Common capacitor values range from 500 pF/sq.mm to 2 000 pF/sq.mm. Leakage currents and nonlinear effects limit the application of junction capacitors. On the other hand, their variability can be of practical importance. Aluminum oxide is used as dielectric in GaAs processes [18]. Silicon oxide has lower leakage currents and a lower temperature coefficient compared to silicon nitride. Silicon nitride is less sensitive to electrical damage. The oxide/nitride combination has the advantages of both. The presence of silicon in the silicon oxide increases the dielectric constant [19]. Different capacitor types may be combined in parallel to increase the capacitance per sq.mm. A junction capacitor below a gate-oxide capacitor and a capacitor between two layers of interconnect may be shunted to arrive at a value of 3 000 pF/sq.mm. The trench capacitors have a much larger (3.7 times larger, as reported) effective area relative to the part of the chip that they take.

Initial tolerances (deviation of the mean capacitor value from the design value) are of the order of magnitude of $\pm 10\%$. Trimming is possible (laser trimming, zener zapping), but seldom applied. Capacitor ratios are used. Matching is important. Matching inaccuracy can be of the order of magnitude of 0.1 %, see [20] and [21].

Temperature coefficients can be of the order of 100 ppm/K and lower. All dielectric capacitors can have a low voltage dependence ($< 0.1\%/V$). Depletion layers in diffused monolithic or in poly silicon electrodes may be the cause of increased voltage dependence (when they are not sufficiently highly doped). Depletion layers may also be applied on purpose to make voltage variable capacitors (see, for example, [22]).

The ratio of the stray capacitance (to lower layers) to the desired capacitance is of the order of magnitude of 5-20 %. The percentage is invariant for capacitors on field oxide (LOCOS) and depends on the capacitor size and bias for capacitors on diffusions. Lower dopes and thinner field oxides in modern processes shifts the preference for capacitors on field oxide to capacitors on silicon (lower parasitics). In some cases the influence of the parasitic capacitors can be largely reduced by bootstrapping, see figure 15 and [7].

BOOTSTRAPPING

Fig. 15. — Bootstrapping of stray capacitors. In some cases the influence of stray capacitors can be eliminated by keeping the layer below the lower terminal at the same (signal) voltage. Often several filter capacitors are incident to a node. In that case only one voltage follower suffices for bootstrapping the corresponding stray capacitors.

The capacitor value per sq.mm in combination with the breakdown voltage yields the following estimates for the maximum charge density in the capacitors :

- junction capacitors (base/emitter junction) : 7 nC/sq.mm,
- dielectric capacitor (oxide/nitride sandwich) : 30 nC/sq.mm,
- trench capacitor : 60 nC/sq.mm ?

The trench type capacitor which has been developed for memory applications may be used for filters in the near future.

3.2 RESISTIVE ELEMENTS. — Resistors of various types are used :

- metal-film resistors,
- diffused resistors (monolithic or poly silicon),
- transistor (trans)conductances.

Only in the case of metal film resistors can one rely upon the absolute resistor value (on-chip trimming possible, low temperature coefficient). This technology is an (expensive) addition to standard processes (e.g. for A/D converters). In all other cases one relies upon matching rather than on absolute values. Obtainable (local) spreads are : thin film : 0.2 %, implanted : 0.3 %, diffused : 0.4 %. Dynamic matching can improve the above values by some orders of magnitude [23] at the cost of circuit complexity. Current trimming of heavily doped poly-silicon resistors can yield similar extreme accuracies [24].

Parasitic capacitances may be eliminated by bootstrapping. Nonlinearity caused by modulation of the width of diffused resistors can be reduced by distributed bootstrapping [25].

The (trans)conductances of (bipolar and field effect) transistors, too, are used as resistive elements. They are used as resistor (diode-type arrangements) as well as in transconductor applications. The feature of the transconductor (voltage controlled current source (VCCS)) of having separate voltage input and current output is used in filters.

Whereas metal-film and diffused (or implanted) resistors are fixed resistors, the transistor (trans)conductances can be varied by some bias voltage (and/or bias current). In integrated analog filters the capacitive and/or resistive elements are commonly variable so as to be able to control the filter time constants. This makes on-chip trimming unnecessary.

The influence of the nonlinearity of the transistor conductances is reduced by antimetric excitation of symmetric arrangements (cancellation of all even-order harmonics) and by employing more sophisticated transistor combinations. Examples are given below.

3.3 EXAMPLES FROM LITERATURE. — Let us consider some examples as found in the literature.

The first example uses fixed resistive elements (transconductors) and variable (junction) capacitors (Fig. 16). The combination has been used for resonance circuits and short analog delay lines in the MHz region (video frequencies) [26, 27].

Fig. 16. — Junction capacitor — gyrator combination. It is used for video resonance circuits and short analog (luminance) delay lines, see [26, 27]. The transconductors have a fixed value. The tuning bias for the junction capacitors is controlled by a reference frequency (frequency-locked loop).

Similarly, fixed capacitors have been combined with variable resistive elements (Figs. 17-22). The tuning bias for the variable elements comes from an auxiliary circuit locked to a reference frequency (frequency-locked loop or phase-locked loop) or from a stabilizer with (external) reference resistor (one adjustment), see figure 23. Accurate matching of the auxiliary circuit to the filter circuits provides automatic correction of the time constants of the filter.

Transconductors, integrators and resonators are shown in the examples (no filters). They can be used as building blocks, which can be combined to form filters.

Variable voltage-to-current conversions (transconductors) are made, either using fixed linear resistors followed by some scaling (e.g. by current multiplication, see figure 17 and [7, 28, 29]) or as nonlinear

Fig. 17. — Transconductor-multiplier combination. The transconductor value is controlled by multiplication of the signal current by a ratio of bias currents, see figure 23 and [7] (one adjustment). The capacitors have a silicon oxide/nitride sandwich dielectric (between aluminum and emitter diffusion).

Fig. 18. — Differential integrator with junction-FET transconductor (in bipolar technology). Proposal for application in signal-flow graph filters. The tuning bias is controlled by a reference frequency (phase-locked loop) [30].

Fig. 19. — A combination of two long-tailed pairs as a simple linearized transconductor (VCCS) of bipolar transistors. It is employed in transconductor-capacitor filters and in short (tapped) analog delay lines. It can be applied at low supply voltages (1 V) and up to high frequencies (10 MHz). The tuning bias current has to be proportional to the absolute temperature [31].

variable elements where transistor properties (transconductances) are used deliberately. In general the latter type of transconductor is simpler. Figure 18 (see [30]) shows an integrator with junction field-effect transistors. Methods for linearization have been proposed for bipolar transistor differential stages (Fig. 19, [31]) as well as for MOS transistor differential stages (Fig. 20, [32]).

MOS transistors have been used as variable resistors in the proposals in figures 21 and 22. Antimetric

Fig. 20. — Cross-coupling of MOS long-tailed pairs yields a linearized transconductor, too. Capacitive coupling of fully balanced resonators gives a band-pass filter [32].

Fig. 21. — Antimetric excitation of symmetric MOS transistor resistors yields accurate cancellation of even-order harmonics [33]. Earthed (virtually earthed) resistors are controlled (at gate and back bias) by half the signal voltage superimposed on the bias voltages. The method can be employed for signal-flow graph filters.

Fig. 22. — Fully balanced differential integrators. The application of fully balanced differential integrators is an alternative to the method in figure 21 for cancellation of all even-order harmonics [34]. The method has been proposed for application in signal-flow graph filters.

harmonic excitation ($+V_s$ and $-V_s$ at source and drain, see figure 21, V_c is a bias voltage) of symmetric MOS transistors, yields a current which is free from harmonics of even order. Addition of a voltage V_s to all terminals does not change the current. In the implementation (an integrator), the gate as well as the back bias are controlled by half the signal voltage superimposed on the bias voltage (see also [33]). Elimination of all even-order harmonics is also obtained when the symmetric integrators in figure 22 (see [34]) are used.

Just as for transconductors, it has been tried for resistors also to make combinations of MOS transistors to improve the linearity (see [35]). Poly silicon resistors have been modified to MOS transistors to obtain a high variable resistance on a small chip area [36].

There are a large number of methods for analog filter design, all with their advantages and disadvantages. In high-performance biCMOS processes, all combinations can be made and compared. Usually the choice of the process is not fixed by the filters and only a restricted set of filter types can be made.

3.4 ADDITIONAL REMARKS.

— We have considered filter design methods, filter elements on silicon and a number of combinations. Modifications and different combinations are in fact conceivable.

— Although we have mainly considered capacitive and resistive elements, the choice (and design) of the circuits for the control of the time constants is as important (see Fig. 23).

— Design problems concerning latch-up, overflow-limit-cycle oscillations and stability at high frequencies have not been considered in this survey (see [10]).

— Active-R filters, which use the dominant pole of operational amplifiers for filter design (see [38]), are considered to be special examples of integrator-based filters.

— We have stated that coils cannot be integrated. This is not correct for very-high-frequency applications (GHz frequencies), on high-ohmic GaAs substrates, in particular, see [39].

— Distributed R-C structures have not been included in the survey because (1) general design methods are lacking and (2) they cannot be designed as lossless two-ports between resistive terminations [37].

4. Performance, limitations and challenges.

The following is a set of yardsticks for estimating and comparing the performance of analog filters :

- signal-handling capacity (S/N ratio, distortion),
- efficiency (dissipation),
- chip area (costs),

Fig. 23. — Control methods. Upper left — An auxiliary resonance circuit is locked to a reference frequency (f_0) in a frequency-locked loop (FLL) arrangement. The tuning bias is also applied to the filter (tracking). Upper right — An auxiliary oscillator is synchronized in a phase-locked loop (PLL) arrangement. Lower left — This circuit corresponds to the filter method in figure 17. The integrated filter resistor R_k is multiplied by the ratio of two bias currents, $I_1/I_2 = R_{ext}/R_{int}$. The effective resistance $(R_k/R_{int}) R_{ext}$ is the ratio of 2 integrated resistors (constant factor) multiplied by the reference resistor. Lower right — The stabilizer provides a temperature proportional bias current to the transconductor in figure 19. The effective resistance $(25/8) (kT/qI)$ becomes $(25/8) R_{ext}/\ln(m \cdot n)$ (temperature-independent).

- accuracy (yield),
- range of application,
- simplicity of design (VLSI),
- electromagnetic compatibility (EMC).

We define the signal-handling capacity of a filter as the distance between the maximum signal (limited by distortion requirements) to the noise (at the output of the filter). For a « symmetric » resonance circuit of a gyrator with two capacitors (Fig. 24) the noise voltage is : $\sqrt{kT/C} (1 + FQ)$, where C is the capacitor value, Q is the quality factor of the resonance circuit and F is a(n) (excess) noise factor [40]. In passive implementation $F = 0$. For an electronic implementation (when we assume that the gyration resistors are noisy as diodes) $F = 1$. In most practical cases $F > 1$. We define the efficiency (η) of the gyrator as the ratio of the maximum gyrated signal power to the dissipation (supply power consumption) of the gyrator. Typical values are of the order of 1 %. An expression can be given for the dissipation of the gyrator in terms of the above parameters (Fig. 24). The dissipation increases with increasing signal handling, narrow bandwidth, high noise factor, low gyrator efficiency and high frequencies.

Fig. 24. — Noise of a « symmetric » gyrator resonance circuit. For a passive L-C resonance circuit the noise voltage is $\sqrt{kT/C}$ (noise of loss resistances). The same value applies to the resonance circuit with a passive gyrator (passive implementation) with noiseless gyration resistances. Assuming that the noise currents of the gyration resistances of an active gyrator correspond to diode noise, the noise currents are $1 + Q$ times larger [40]. An excess noise factor F makes the factor $1 + FQ$. The gyrator efficiency (η) is defined as : (maximum gyrated signal power)/(supply power consumption). Now the supply power consumption of the gyrator can be expressed in a set of useful parameters (including the signal-handling capacity).

Fig. 25. — Supply power consumption of band-pass filters (rough estimate). The result in figure 24 has been generalized. Approximations and practical considerations have been used. In the graph, $kT = 4 \times 10^{-21}$, $F = 4$ and $\eta = 1\%$. High frequencies, narrow bandwidths and large S/N ratios increase the dissipation of RC-active filters. Signal level adaptive methods and class A/B operation may reduce the average dissipation (but at the cost of increased EMC problems).

A similar (semi-empirical) formula can be derived for band-pass filters, see figure 25 and [10]. Methods to reduce the average dissipation are :

- control of the supply current proportional to the (instantaneous) signal level (e.g. implemented in the adaptive gyrator [10]), and
- class-B operation (see also [41]).

The reduction in dissipation is obtained at the cost of increasing EMC problems (e.g. cross-talk *via* supply lines and substrate).

The accuracy of the filters determines the part of the chips which will be within the specifications. It is determined by the accuracies of

- the reference (frequency, resistor, C-value),
- the comparison (PLL, FLL, stabilizer),
- the matching (C 's : 0.1 %, R 's : 0.2 %, transistors : 0.3 %).

Whether a filter type can be applied or not also depends on the supply voltage and on the frequency range, see figure 26.

RANGE OF APPLICATION

Fig. 26. — Range of application. Analog filters can be made for supply voltages > 2 V; and with reduced accuracy (or increased complexity) for supply voltages > 1 V. The frequency range for practical applications is (roughly) from 10 Hz to 10 MHz. In general, high frequencies and low supply voltages lead to bipolar transistor circuits, large time constants (control loops) to CMOS circuits (lower transconductance). Electronic multiplication of time constants reduces the chip area.

Common supply voltages are 18 V, 12 V, 5 V, 3 V, 1.8 V, 1.2 V. Below 5 V it is becoming increasingly difficult to design filters for a high signal-handling capacity (say > 80 dB, combined with a small chip area and a low current consumption). Below 2 V the filter accuracy tends to decrease (assuming the use of simple circuits) (see, for instance Fig. 27 and [42]).

At low frequencies (say, 1 Hz) large time constants ($R \cdot C$ products) have to be made. The wish for a small chip requires : small capacitors (say, 100 pF) and hence nonrealistic resistor values (1 G Ω). Electronic multiplication techniques (e.g. by application of the Miller effect) can solve this problem. The resulting limitations are found to be of a more practical nature (DC offset and capacitor leakage). Low frequencies are of importance for the integration of control loops. On the other hand, one should be aware that, particularly at low frequencies, digital solutions may well be more economical.

For filters to be used at high frequencies dissipation is one of the main limiting factors. It will often be better not just to replace an existing passive filter by its active counterpart, but to reconsider the system (filtering can

Fig. 27. — Filter circuits for low supply voltages (1 V). The right-hand side of the figure shows two linearized transconductors. The left circuit is a bias current stabilizer (all current source transistors have equal collector-emitter voltages — compensation of Early effect). The circuit in the centre is a (NIC-type) low-ohmic filter earth circuit.

sometimes be transformed to lower frequencies). System modifications may reduce the dissipation and the accuracy requirements. In applications for higher frequencies one should avoid (slow) pnp and pMOS transistors in the signal paths. A minor phase shift can

TRANSCONDUCTOR DELAY

- (1) $G_k \rightarrow G_k / (1 + p\tau)$
- (2) $G_k \rightarrow G_k / (1 + p\tau)$
- (3) $p \rightarrow p(1 + p\tau)$
- (4) $A + j\varphi = -\ln H\{p(1 + p\tau)\}$
- (5) CAUCHY - RIEMANN

$$\begin{aligned} dA &= -\omega^2 \tau \frac{\partial \varphi}{\partial \omega} \quad (\text{ENHANCEMENT}) \\ d\varphi &= +\omega^2 \tau \frac{\partial A}{\partial \omega} \\ d\left(\frac{\partial \varphi}{\partial \omega}\right) &= +\omega^2 \tau \frac{\partial^2 A}{\partial \omega^2} \end{aligned}$$

GROUP DELAY

COMPENSATION

$$pC_k \rightarrow pC_k / (1 + p\tau)$$

SERIES RESISTANCES τ/C_k

Fig. 28. — Influence of uniform transconductor delay. We consider a transconductor-capacitor filter with uniform transconductor delay. The delay can be represented by an extra time constant (1). Multiplication of all admittances in the filter by $(1 + p\tau)$ leaves the (voltage) transfer function invariant (2). The transconductor delay proves to be equivalent to a frequency transformation (3). With $p = \sigma + j\omega$ we arrive at $d(\sigma) = 0$, $d(\omega) = \omega \cdot \sigma$, as a first approximation. The (negative of the) natural logarithm of the transfer function $H(p)$ is analytic almost everywhere. The real part « A » is the attenuation (in Neper) and the imaginary part « phi » is the phase (in radians) (4). Application of a Taylor expansion and the Cauchy-Riemann relations yields the lower relations. One of the effects proves to be an enhancement of the gain proportional to the group delay of the filter. The effects can be compensated by addition of (low) resistances in series with the filter capacitors.

be as important as the loss angle ($\tan(\delta)$) of a capacitor. Delay in the resistive elements, too, can degrade the filter characteristics (see Fig. 28). The corresponding gain enhancement proves to be proportional to the group delay of the filter. N.B. this is also a grave limiting factor for the design of longer analog delay lines. Compensation is possible by adding (small) resistors in series with the filter capacitors. In general we have to remain far below the 3 dB cut-off frequency of the electronic elements, making an exception for the operational amplifiers, the delays of which can often be cancelled out [11]. Nowadays, analog filters can be made up to approximately 10 MHz. It is of importance to arrive at still higher frequencies so as to be able to

- apply filters at still higher frequencies and
- simplify the design at somewhat lower frequencies (for VLSI).

Together with a higher density on the chips, crosstalk and interference problems increase. In particular, sensitive analog circuits cannot be put near large-swing digital or switched-capacitor circuits. Shielding methods have to be developed. EMC problems within the VLSI chip are becoming more and more important.

5. Conclusion.

We may conclude by stating that

- analog filters are needed in the interfaces between digital processors and the analog outside world (transmission),
- design methods (synthesis) of analog filters are well established,
- VLSI-oriented technology is improving rapidly (opening new possibilities to and posing new problems for designers),
- all resulting in improved analog integrated filters (which has not yet run its course).

The challenges to designers are to achieve :

- higher S/N ratios at lower supply voltages,
- higher frequencies,
- large time constants (for control loops),
- further improved accuracy (standard cells, simplified filter design), and
- more sophisticated applications (controlled, programmable, adaptive filters).

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