Design of a Switched-Capacitor Filter for a Mobile Telephone Receiver

Li Ping, R. C. J. Taylor, R. K. Henderson, and J. I. Sewell

Abstract—**Conventional SC realizations of wide-band filters demand large capacitance spread and exhibit serious sensitivity problems. The UL–LU structure is a ladder type simulation which demonstrates superior sensitivity performance and maintains low capacitance spread. The design illustrates how the facilities of a modern filter compiler can be utilized to solve some quite difficult practical problems in a real application. Results from fabricated devices confirm the predicted properties.**

I. Introduction

The voice-band frequency range normally extends from 300 Hz to 4 kHz. The difficulty of designing filters in this range is that the relative bandwidth, defined by

$$\mathrm{RBW} = (\omega^+ - \omega^-)/\omega_m, \quad \omega_m = (\omega^+ \times \omega^-)^{1/2}$$

has a typical value of 3.5, which is quite high. Switched-capacitor (SC) realizations of such wide-band filters will have a large capacitance spread and serious sensitivity problems. In general, ladder-based realizations are known to produce low sensitivity solutions [1], though computer simulations [2] have shown that capacitance spread is particularly serious for bandpass leapfrog circuits [3] that use F-type damping. The sensitivity problem for cascade biquads is well known. It is also observed that coupled biquads [4] and LUD methods [5] suffer from high sensitivity at very low frequencies; in the case of low-pass filter filter designs, it has been proven [6] that this property will always result when the so-called right-hand matrix is not decomposed. Similar reasoning can be extended to the bandpass case.

In this paper, a practical voice-band filter design for a mobile telephone application is presented. A novel variant of the LUD method is utilized. It combines a low capacitance spread with low sensitivity, which are important in reducing silicon area and easing the design requirements of the amplifiers and switches compared to other solutions [7]. The strategy is to retain E-type damping (low capacitance spread) simultaneously with right-hand matrix decomposition (low sensitivity), which is not readily available with existing approaches. A modern filter compiler [9] has been used and exercised considerably to provide an optimum solution.

II. Wide-Band Filter Design

An audio receiving filter for mobile telephony applications typically has a wide passband extending from 300 Hz to 3 kHz and exhibiting a −20-dB/decade slope for frequency de-emphasis. A notch is included to remove unwanted mixed down frequencies which are close to the upper passband edge. The template can be seen in Fig. 1.

If the specification is met by a function with an elliptic-type zero distribution, a very large capacitance spread will be incurred. The finite zeros in the lower stopband create very large time constants that require large capacitors. Alternatively, if these zeros are avoided by choosing an all-pole approximation with zeros at zero and infinite frequency, the required order will be very high. A compromise is to use a function with all lower band zeros at zero frequency and elliptic-type zeros in the upper stopband. Since the lower band edge of the filter is at low frequency, the movement of the lower band zeros to the origin has only a slightly deleterious effect on the filter characteristic [2]. Such a transfer function which meets the required template is shown in Fig. 1. It is a tenth-order function with a third-order zero at the origin, a zero at infinity, a pair of imaginary axis zeros, and two pairs of mirror image real axis zeros. The latter zeros are specially placed to cause negative element values in the passive prototype which cancel components in the SC simulation [2]. The number of capacitors required to realize a pair of mirror image real axis zeros at $\pm 2f_s$ (f_s is the sampling frequency) is less than that to realize a pair of zeros on the imaginary axis. The capacitance spread can also be greatly reduced this way. Since f_s is much higher than the passband frequency, a pair of zeros at $\pm 2f_s$ has the same function as those at infinity. The approximator tidies up after these forcing modifications and ensures that the original specifications are still satisfied.

A comparison of sensitivities is available from the filter compiler PANDDA [9]. This has been made for a number a different realizations [3]–[6] and the results are shown

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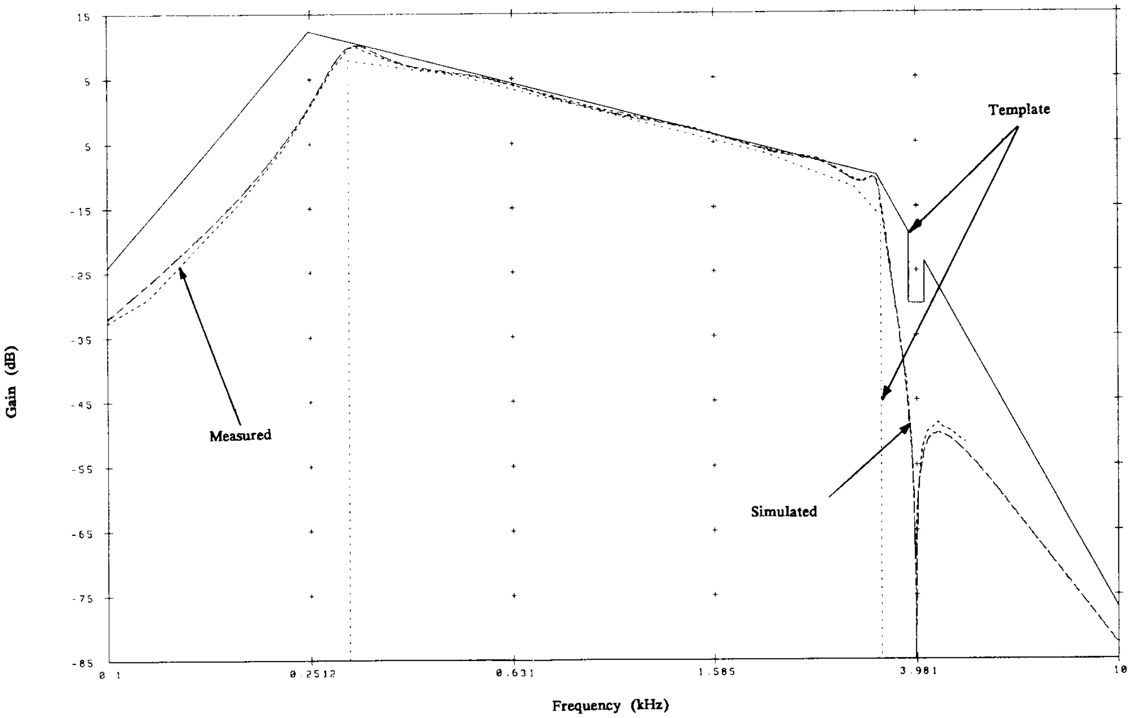

Fig. 1. Template and frequency responses of tenth-order filter.

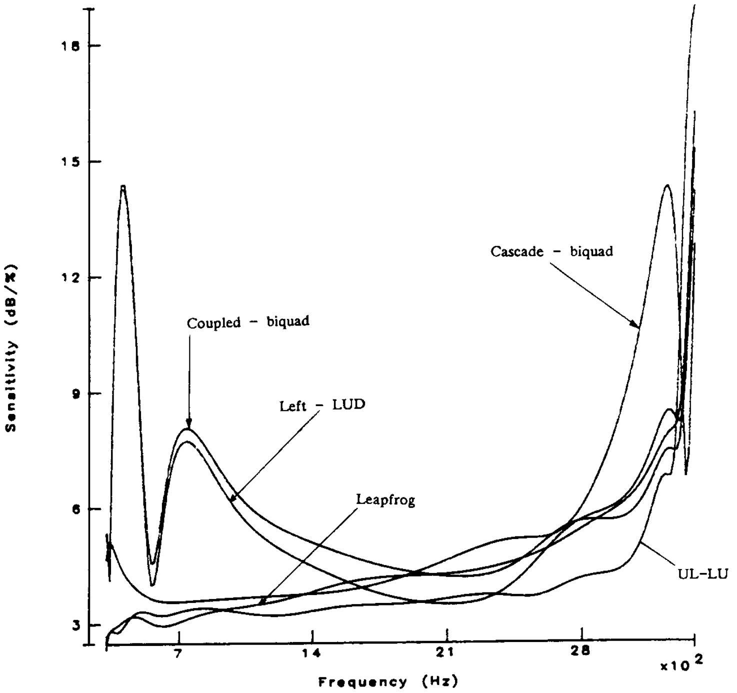

Fig. 2. Passband sensitivity comparison for tenth-order filter realizations.

in Fig. 2. The sensitivity measure being used is

$$s(\omega) = \left\{ \sum_i \left[\frac{c_i}{|H(\omega)|} \frac{\partial |H(\omega)|}{\partial c_i} \right]^2 \right\}^{1/2}.$$

It can be seen that both left-LUD and coupled-biquad circuits suffer from a low-frequency sensitivity peak, whereas the leapfrog and cascade-biquad circuits exhibit poorer sensitivity performance at the higher band edge.

TABLE I
COMPARISON OF CAPACITANCE COSTS FOR FILTER REALIZATION

Structure	Total Capacitance	Capacitance Spread
UL-LU	746.4	113.1
LUD	703.3	112.5
Coupled-E-Biquad	756.4	112.5
Leapfrog	1734.7	957.8
Biquad	728.8	122.5

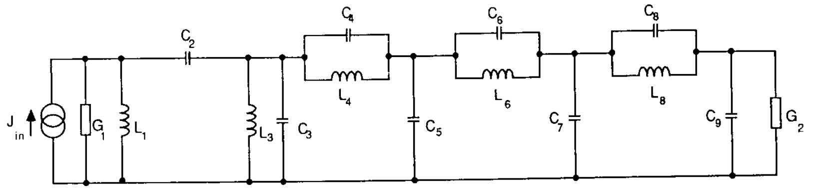

Fig. 3. Tenth-order passive prototype.

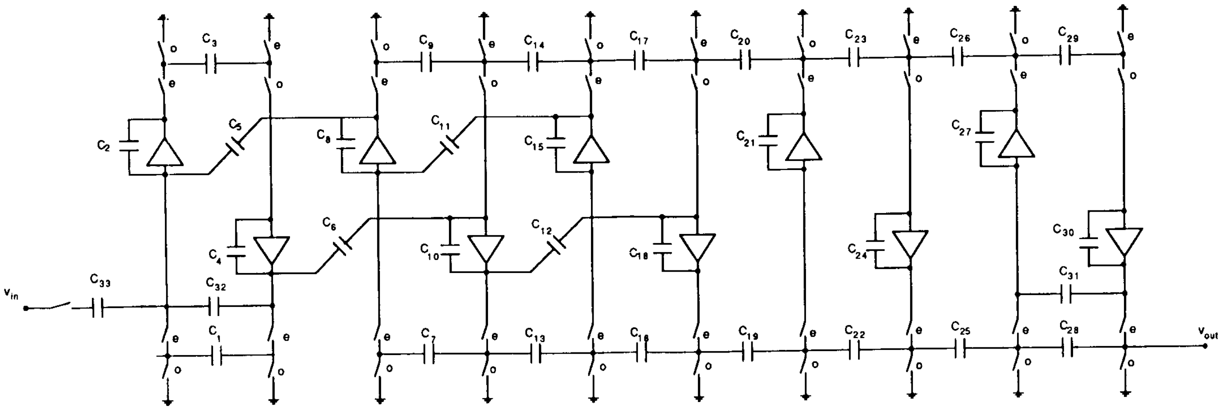

Fig. 4. UL-LU SC circuit realization.

The new UL-LU structure (detailed in the next section) maintains a low sensitivity over the whole passband. Table I gives a comparison of capacitance cost and highlights the penalty of choosing a leapfrog realization. The conclusion of these studies is that the UL-LU structure offers the best solution to this filtering problem. Fig. 3 shows the structure of a passive prototype synthesized from this transfer function. Note that the position of a series capacitor between the first and second nodes is required to ensure E-type terminations [7] in the SC circuit realization shown in Fig. 4. The simpler structure of the filter towards the output is due to the cancellation of feedthrough capacitors by the specially positioned real axis zeros (capacitors C_6 and C_8 in Fig. 3 therefore have negative values).

The circuit uses a clock frequency of 128 kHz and has been fabricated onto silicon using a 3-μm single-metal, double-poly, 5-V process. The amplifiers used are single-stage folded cascode without compensation capacitors, stability being ensured by the capacitors in the filter; the input devices are laid out as cross-coupled common centroid to help with matching and to reduce offsets. The area-to-perimeter ratio of unit to nonunit capacitors is kept constant so that parasitic peripheral capacitance effects are cancelled. The size of the filter is 3027 μm × 894 μm, which is relatively small considering the complexity of the response. Passband details of the measured frequency response can be seen in Fig. 5. The response meets the template very well and over a number of devices there was hardly any deviation. This illustrates the low sensitivity of the filter to process variations and hence the robust nature of the design. For commercial production this is an important factor in ensuring an increased yield and reliability of the devices. The noise floor at 1 kHz was −60 dB and was approximately level at this value over the entire passband; no harmonic level could be detected above the noise floor.

III. DERIVATION OF THE UL-LU STRUCTURE

The derivation utilizes a UL-LU decomposition together with a bilinear-LDI ladder design [2]. Starting from

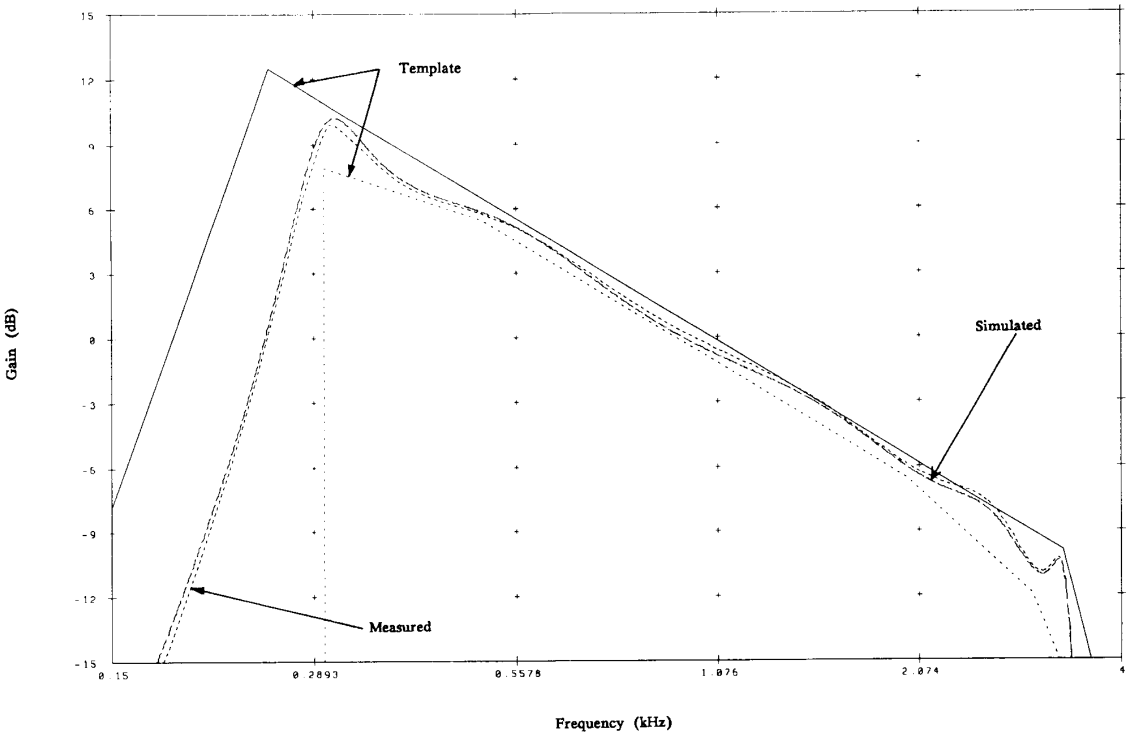

Fig. 5. Passband detail of tenth-order filter.

the nodal equation of a passive prototype ladder

$$(s\boldsymbol{C} + s^{-1}\boldsymbol{\Gamma} + \boldsymbol{G})\boldsymbol{V} = \boldsymbol{J}, \tag{1}$$

applying a bilinear transformation, and separating out a pair of LDI integration operators $\Phi = 1/(1 - z^{-1})$ and $\Psi = z^{-1}/(1 - z^{-1})$, equation (1) becomes

$$\left(\frac{1}{\Psi}\boldsymbol{A} + \Phi\boldsymbol{B} + \boldsymbol{D}\right)\boldsymbol{V} = (1 + z)\boldsymbol{J} \tag{2}$$

where $\boldsymbol{A} = 2/T\boldsymbol{C} + T/2\boldsymbol{\Gamma} + \boldsymbol{G}$, $\boldsymbol{B} = 4T\boldsymbol{\Gamma}$, and $\boldsymbol{D} = 2\boldsymbol{G}$.

A simplified UL–LU form has $A = U_a L_a$, $B = L_b U_b$, $\boldsymbol{W}_a = \Psi^{-1}\boldsymbol{L}_a \boldsymbol{V}$, $\boldsymbol{W}_b = \boldsymbol{U}_b \boldsymbol{V}$, and $\boldsymbol{D}_s = \boldsymbol{D}\boldsymbol{U}_b^{-1}$. Again the upper triangular matrix $\boldsymbol{U}_{as}$ and lower triangular matrix $\boldsymbol{L}_{bs}$ are defined to satisfy the identity $\boldsymbol{U}_{as}\boldsymbol{L}_a = \boldsymbol{L}_{bs}\boldsymbol{U}_b$. Then (2) can be linearized in terms of the LDI operators as

$$\boldsymbol{U}_a \boldsymbol{W}_a = -(\Phi\boldsymbol{L}_b + \boldsymbol{D}_s)\boldsymbol{W}_b - (1 + z^{-1})\boldsymbol{J} \tag{3a}$$

$$\boldsymbol{L}_{bs}\boldsymbol{W}_b = \Psi\boldsymbol{U}_{as}\boldsymbol{W}_a. \tag{3b}$$

The scheme described by (3) is a variation on those derived in [2]. In general $\boldsymbol{D}_s$ is less sparse than $\boldsymbol{D}$, as the entry $\boldsymbol{D}_{11}$ when multiplying the first row of $\boldsymbol{U}_b^{-1}$, which is the upper triangle, would produce a full nonzero row. However, in this design, the first row of $\boldsymbol{U}_b^{-1}$ has only one nonzero entry, resulting from the fact that the first inductance L_1 in the prototype is separated from the other inductors. The beneficial effect of this is to ensure E-type damping at the input of the SC realization and this is consistent with E-type damping already implicit at the output. The virtue of E-type damping is to reduce capacitance spread in the termination sections of the SC realization. Notice now that the output is w_{bn}; as $\boldsymbol{U}_b$ is the upper triangular, w_{bn} differs from the output by only a constant. The realization procedure for (3) by an SC circuit follows matrix methods [2], [9] and yields the circuit shown in Fig. 4. A further point of interest is to note that the input factor $1 + z^{-1}$ in (3a) has been implemented implicitly by prewarping the original specification by a $1 + z^{-1}$ function. This is conveniently combined with the sin $(x)/x$ correction, resulting in tan $(x)/x$ prewarping.

IV. CONCLUSIONS

In this paper we have examined the problem of designing voice-band SC filters with wide-band specifications. Such filters do not have satisfactory realizations by conventional design techniques due to excessive area requirements or sensitivity to component value deviations. Low-frequency notches, which cause the large component spread, are eliminated by designing a transfer function with lower band zeros at the origin. A new ladder simulation structure has been proposed to overcome the sensitivity and capacitance spread problems of other realizations. A tenth-order filter with sloping passband response has been fabricated and the measured results verify that difficult audio frequency responses can be met practically using a relatively small area of silicon. This particular design illustrates the potential of modern filter compilers and their application to difficult practical design problems.

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