

examples which illustrate the good results obtained with the new method. The mean-square errors (MSE) and the modified

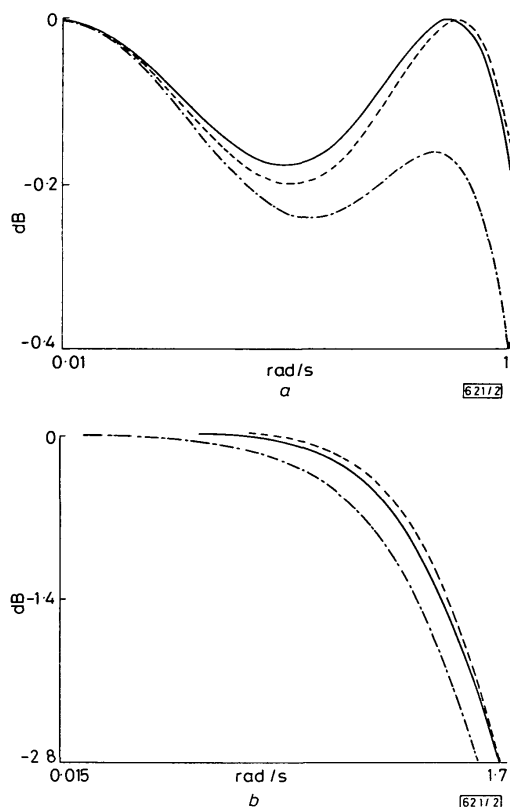


Fig. 2 Examples of 3rd-order filters with $\omega_d/\omega_p = 8$, ripple = 0.177 dB and $\tau = T$

a Elliptic filter: MSE (without correction) = 1.35×10^{-4} , MSE (with correction) = 3.57×10^{-6} , $a_1 = 0.82955$, $a_{c1} = 0.79885$
b Butterworth filter: MSE (without correction) = 1.33×10^{-4} , MSE (with correction) = 4.8×10^{-6} , $a_1 = 1.6988$, $a_{c1} = 1.5756$

coefficients are shown for each example. The MSE was calculated using 80 points uniformly distributed from $0.01\omega_p$ to ω_p , and defined as

$$\text{MSE} \triangleq \frac{\sum_{j=1}^{80} |m_c(j) - m(j)|^2}{\sum_{j=1}^{80} |m(j)|^2} \quad (8)$$

where $m(j)$ and $m_c(j)$ are the magnitudes of $H(\omega)$ and $H_{cw}(\omega)$, respectively.

Conclusions: A new analytical method which is extremely simple to apply has been proposed for the compensation of $(\sin x)/x$ distortion. It consists of the modification of one coefficient of each quadratic factor in the denominator of the transfer function of the reconstruction filter. Despite the simplicity of the method, significant improvements in the simulated filter responses were obtained. These improvements were verified both graphically and by the mean-square error calculated in the passband.

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CURRENT COPIER CELLS

Indexing terms: Current copying, Current sources, Analogue integrated circuits, Integrated circuits, Circuit theory and design

Analogue circuits are presented that can sample, hold, and make practically identical multiple copies of current signals, without the need for accurate matching of active or passive circuit components.

Introduction: Conventional current mirror techniques are eventually limited by the achievable matching of transistors used as current sources. A corresponding limit is thus imposed on the accuracy of data converters and signal processors using current mirrors. In this letter we propose a technique for sampling and holding a current, and for making practically identical copies of it without the need for accurately matched components.¹ The technique may be applicable to A/D and D/A converters as well as signal processors. The use of current rather than voltage may help reduce the problem of limited voltage swing encountered in modern VLSI processes because of scaling.²

Current copiers: In contrast to the present approaches, the current copier technique does not require matching of transistors for its operation. The circuit requires a noncritical MOS transistor, a noncritical capacitor and switches. To copy the current I_{ref} , switches S_1 and S_2 are closed (Fig. 1) feeding the current I_{ref} to the transistor and capacitor. The capacitor

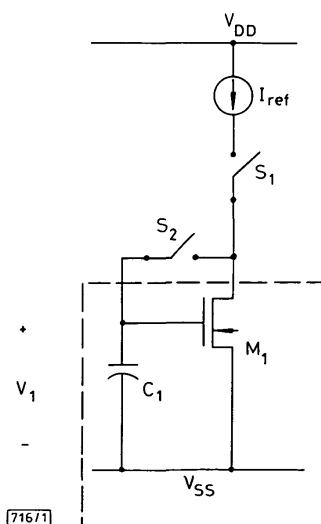


Fig. 1 Simple current copier

charges to whatever gate voltage is needed by the transistor to support a current equal to the reference current I_{ref} . When the switches S_1 and S_2 are open, the transistor and capacitor are disconnected from the current source; thereafter, the transistor and capacitor combination is capable of sinking a current I_{ref} when connected to a load. This process can be repeated using several 'current copier cells', providing multiple copies of I_{ref} . The simple cell of Fig. 1 can be modified in a

straight-forward manner to allow the copying of bidirectional currents.

In a practical current copier cell, the effects of channel length modulation, charge feedthrough from the switches,³ and parasitic capacitive coupling from the drain to the gate of the transistor, must be addressed.

Because of channel length modulation, the circuit shown in Fig. 1 will not work properly if the drain to source voltage V_{DS} of the transistor M_1 during current copying is not the same as that during cell use (i.e. when the cell is connected to a load). This problem can be avoided by arranging for both current copying and cell use to be performed under constant V_{DS} conditions, as shown in Fig. 2. During copying (ϕ_{1a} and

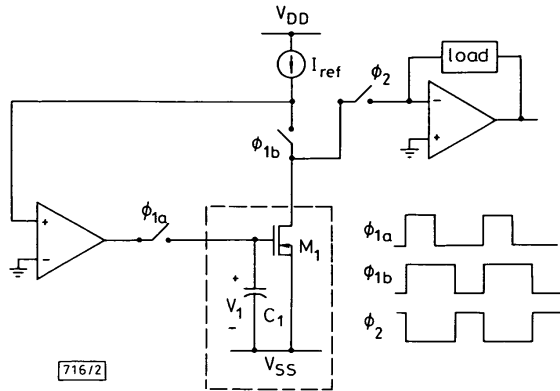


Fig. 2 Op-amp current copier

ϕ_{1b} are both high) the op-amp forces V_{DS} of M_1 to be equal in magnitude to the supply voltage V_{SS} . After copying, the cell is disconnected from the op-amp and is ready to be used as a current source. When used in this way (ϕ_2 is high), V_{SS} should again be forced across M_1 to ensure the proper current is extracted. This is easily done by again using an op-amp, also shown in Fig. 2; a single op-amp can be time-shared for both functions. Note that several current copier cells can be connected to the op-amp negative input on ϕ_2 allowing current summing through the load.

The use of an op-amp could be avoided if the V_{DS} of the storage transistor can be held constant by other means. One way to accomplish this is to add a cascode device (Fig. 3) to the output of the current copier cell and a source follower to the input of the cell. The cascode transistor acts to hold V_{DS} of

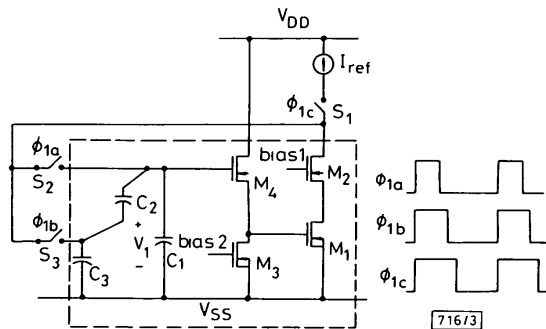


Fig. 3 Cascode current copier

M_1 constant for a given current. But by only adding a cascode transistor the ability to connect the output of the cell to the input is lost, since the potential at the output of the cell would normally be higher than the desired potential at the input. A source follower added to the input of the cell makes the two potentials compatible, allowing a simple scheme for the current to be 'memorised' by the cell. A second reason for using a source follower is to isolate the input of the cell from changes at its output. With the simple current copier (Fig. 1) this coupling can occur through the gate to drain capacitance C_{gd} , with a resulting change in the voltage stored on capacitor C_1 .

In addition to modifying the cell to reduce channel length modulation effects, changes to the switching configuration can also improve the performance of a current copier cell. The

action of turning off an MOS switch dumps charge from the switch's inversion layer onto the nodes it connects. This additional charge will change the voltage across C_1 by a small amount ΔV_1 , resulting in a small change in the current 'memorised' in the cell:

$$\Delta I = g_m \Delta V_1$$

where g_m is the transconductance of M_1 evaluated at the current I_{ref} . To reduce this effect, feedthrough reduction techniques similar to those used in switched capacitor circuits can be used. An example of such a circuit, using feedthrough reduction techniques explained elsewhere,⁴ is shown in Fig. 3. Typically,

$$C_1 = C_3$$

$$C_2 < C_1$$

The result is a reduction in the error ΔI , to $C_2/(C_1 + C_2)$ of its value without this technique.

Simulation results: To illustrate the operation of the current copier circuit, the circuit of Fig. 3 was simulated using the ADVICE circuit simulator program. The simulations used the process parameters for a 1.75 μm CMOS process with two levels of poly for the capacitors, with $C_1 = C_3 = 4 \text{ pF}$ and $C_2 = 0.4 \text{ pF}$.

To evaluate the range of the current copier and how mismatch of devices influences operation of the cell, a reference current ranging from 1 μA to 1 mA was 'memorised' by a copier cell. The current reference was then disconnected from the cell, and a load connected to its output, with 1 MHz clocks controlling the switches. The results from the simulations are grouped into two types of errors. One, an absolute error as shown in Fig. 4a, is the difference between the reference current and the memorised current, while a second error, the relative error, is the one most important in applications such as D/A conversion, and is shown in Fig. 4b. This error is

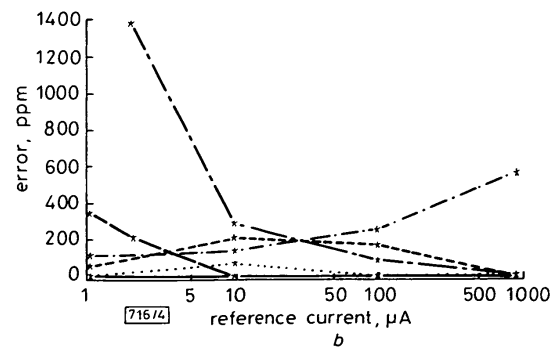
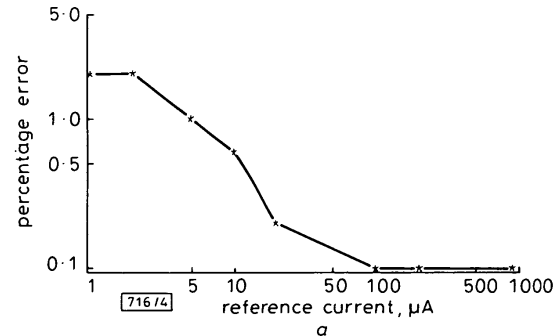


Fig. 4

a Current copier absolute error
b Current copier relative error

— switch 3 mismatch
— M_1 mismatch
— junction leakage
— load mismatch
— M_2 mismatch
— C_1 mismatch

the difference in the memorised currents of two copiers, derived from the same reference current, assuming a 1% mismatch between the devices indicated. The current copier technique, not dependent on device matching, has very small relative errors, measured in parts per million. In contrast, use of the conventional current mirror technique would result in errors of close to 1%.

One other effect must be included when considering current copier cells. Just as in switched capacitor circuits, junction leakage from the MOS switches can change the value of a voltage stored across a capacitor. This leakage places a time limitation on how long a charge can be stored on a capacitor (and how long a current can be memorised by a copier cell). To explore this junction leakage effect Fig. 4b also includes a curve comparing the current memorised by a copier cell after one clock cycle to the current memorised by a cell five clock cycles later. The source-drain diode junction currents of the MOS switches as modelled as ideal diode currents with saturation currents of 1 pA.

The cascode current copier of Fig. 3 was breadboarded using commercially available parts. Measurements on this breadboard were consistent with both absolute and relative error simulation results over similar current ranges and device mismatches.

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INTEGRATED-OPTIC ACOUSTICALLY TUNABLE INFRA-RED OPTICAL FILTER

Indexing terms: Integrated optics, Acousto-optics, Wavelength filters, Optical filters

We demonstrate a narrowband (≈ 1 nm) LiNbO_3 acoustically tunable optical filter with a tuning range adequate to span the 1.3-1.55 μm wavelength range, exploiting the considerable advantages of an unconventional x-cut design.

Introduction: We report the first use of the x-cut orientation of LiNbO_3 in making an acoustically-tuned optical filter (ATOF). The integrated-optic ATOF devices reported to date have all employed a geometry in which the optical and acoustic waves propagate collinearly in the x-direction on a y-cut substrate (Y-X).¹ The X-Y orientation has perhaps been avoided because acoustic anisotropy causes a y-directed acoustic beam to walk off at an appreciable angle from a

y-propagating optical wave, limiting the acousto-optic interaction length. This putative limitation can be easily circumvented and there are significant advantages to using the x-cut orientation over the popular y-cut, namely an enhanced photoelastic effect (by 16 dB) as well as a stronger piezoelectric coupling for surface acoustic wave (SAW) excitation.

Device: The waveguides were made by diffusing 675 Å of titanium into x-cut lithium niobate at 1050°C for 6.5 h in a wet, flowing O_2 ambient. We used a closed platinum crucible for the diffusion.³ White light spectroscopy indicated that the waveguides of interest (6-8 μm) were single-moded between 1.23 and 1.55 μm . The 1500 Å gold interdigital transducers were fabricated by evaporation and liftoff. The SAW transducer consisted of four pairs of 20 μm period electrodes (5 μm lines and spaces), yielding a transducer bandwidth of 30 MHz about a centre frequency of 170 MHz, which is adequate for tuning in the 1.3-1.55 μm range. A transducer width of 1.7 mm was selected to prevent appreciable diffraction of the acoustic beam over three or four cm of interaction length. In all of the experiments to be reported in this paper, the transducers were fabricated in pairs centred on the crystal, allowing for RF insertion loss measurements to be made and thus, acoustic power to be evaluated. Because of this placement, the largest possible interaction length on the 4 cm long, 1 cm wide crystal was 2.5 cm. The transducer period of 20 μm was only approximately correct for 1.523 μm , since at the laser wavelength the birefringence beat length is $\Lambda = \lambda/\Delta n = 20.7 \mu\text{m}$. The consequence of this mistuning, considering the 30 MHz transducer RF bandwidth, was that the one-way SAW transducer efficiency was reduced from its centreband value of -5.5 dB to -7 dB at 1.52 μm and to around -17 dB at 1.30 μm , where other measurements were made.

Experiment: Fig. 1 shows the layout of the filter bandwidth measurement experiment, with a detail of the ATOF itself.

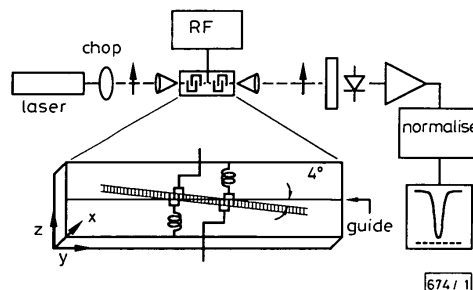


Fig. 1 ATOF experiment shown in band-reject configuration used for conversion efficiency measurements

Inset: detail of crystal layout

For most measurements, a 1.52 μm HeNe laser beam was chopped, passed through a crystal polariser and then endfire-coupled into the ATOF crystal. The output beam was passed through another crystal polariser and the image of the waveguide under study was apertured to reduce unguided light contamination and measured by a Ge detector through an interference filter. The RF drive to the SAW transducer was itself chopped at another, lower frequency to allow (instantaneous) comparison of the filtered and unfiltered response. This information allows one to normalise out the effects of laser power drift and, more importantly, to compensate for changes in the endfire coupling efficiency caused by crystal heating due to the RF power. The normalisation network used to produce transmission plots was cross checked with direct measurements of RF-on and RF-off optical power on an oscilloscope. We evaluated filter bandwidth and conversion efficiency at both 1.52 μm (HeNe) and at 1.299 μm (diode laser) for waveguide widths between 7 μm and 9 μm . Measurements were made in either the bandpass mode (input and output polarisers orthogonal) or in the band-reject mode (parallel polarisers). Conversion efficiencies were evaluated in the band-reject or dip mode.

Results: Because the SAW travels in both directions away from the transducer, it was necessary to suppress one direc-