A Low-Noise Chopper-Stabilized Differential Switched-Capacitor Filtering Technique

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Abstract—This paper describes the implementation of a wide dynamic range voiceband switched-capacitor filter using a differential chopper-stabilized configuration. The noise behavior of switched-capacitor filters is discussed qualitatively, and the effects of the chopper stabilization on the noise performance is analyzed. Experimental results from a fifth-order low-pass voiceband prototype are presented.

I. Introduction

SWITCHED-CAPACITOR filtering techniques have been widely applied to voiceband applications requiring dynamic range on the order of 85 dB. Depending on details of design and fabrication, these filters have been limited in dynamic range by operational amplifier noise, thermal noise in the transistor switches, or a combination thereof. In addition, the realization of very high levels of power supply rejection ratio (PSRR) has proven to be a difficult task, with PSRR on the order of 40 dB typical in work reported to date.

This paper describes a switched-capacitor filtering technique which is aimed at improving the dynamic range and power supply rejection of such filters. It differs from conventional approaches in two respects. First, the signal path throughout the filter is fully differential rather than single-ended as in the conventional case. This results in reduced injection of power supply and clock-related signals into the signal path, and also increases the dynamic range since the effective signal swing is doubled. Second, the first stage of the operational amplifier is chopper-stabilized, which reduces one component of the operational amplifier noise, the $1/f$ noise. Assuming that the other sources of noise in the filter are reduced by suitable design of the operational amplifier and choice of capacitor values, this allows the dynamic range of such filters to be extended to beyond 100 dB.

In Section II, the important sources of noise in switched-capacitor filters are discussed from a qualitative point of view. This discussion is included so that the relative importance of $1/f$ noise in the overall noise of a switched-capacitor filter can be better appreciated. In Section III, various circuit alternatives for the reduction of $1/f$ noise are discussed, and the effect of chopper stabilization on the noise spectrum of the switched-capacitor integrator is analyzed. In Sections IV and V, implementation of the differential filter configuration is described, and the design of an experimental fifth-order filter is discussed. In Section VI, experimental results from this circuit are presented.

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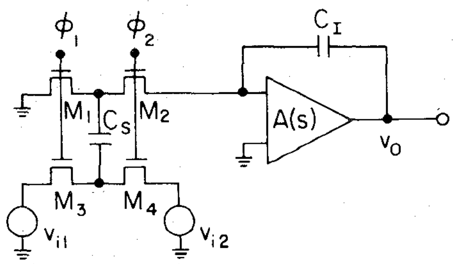

Fig. 1. Typical bottom-plate switched-capacitor integrator.

II. Noise in Switched-Capacitor Integrators

The circuit technique described in this paper is aimed at the reduction of the low frequency noise contributed by the operational amplifier. Other important sources of noise are the wide-band thermal noise in the operational amplifiers and the thermal noise in the channels of the transistor switches making up the filter. While the $1/f$ noise is often dominant, its relative importance in the total filter noise is a function of the details of the design of the operational amplifier and the choice of the sampling and integrator capacitor sizes in the filter. The qualitative discussion of noise in switched-capacitor filters below is included so that the relative importance of $1/f$ noise in overall filter noise for a particular design can be better appreciated.

A. Thermal Noise in the MOS Switches

Thermal noise in the MOS transistor switches represents a fundamental limitation on the amount of noise added to the signal as it passes through the integrator for a fixed value of integrating capacitance. These switches correspond to transistors $M1$, $M2$, $M3$, and $M4$ in the example switched-capacitor integrator shown in Fig. 1. It can be shown [5]–[7] that under the assumption that the clock frequency is much larger than the frequency of interest, that the effective baseband equivalent input noise spectral density of the switched-capacitor integrator is exactly the same as that of a continuous time integrator in which a continuous resistor is substituted for the switched-capacitor resistor. Under the same assumption, it can further be shown that the total input referred mean-squared noise voltage between dc and the unity gain frequency of the integrator is simply equal to kT/C_I where C_I

is in this case the value of the integrator capacitor. Thus, the total in-band noise is determined by the integrator capacitor, which is the larger of the two capacitors in Fig. 1 for voice-band switched-capacitor filters with high sampling rates. As a result, this noise source represents a basic limitation on the dynamic range which can be achieved for a given total capacitance.

B. Operational Amplifier Noise

The equivalent input noise spectrum of a typical MOS operational amplifier is shown in Fig. 2. The noise added to the signal path by the operational amplifier can be divided into two components. Flicker noise, or $1/f$ noise, is concentrated at low frequencies and arises from surface states in the channel of the MOS transistors. The magnitude of the low frequency noise component is dependent on the process used, the design of the operational amplifier used, and on the size of the input transistors used in the operational amplifier. Under the assumption that the noise energy associated with the $1/f$ noise lies far below the sampling rate, the effects of this noise component can be analyzed disregarding aliasing effects. Because the $1/f$ noise source is both integrated and translated by the integrator circuit, its effects can be conveniently represented by two noise sources equal in magnitude to the $1/f$ noise of the operational amplifier, one in series with the input of the switched-capacitor integrator and one in series with the output. This is shown in Fig. 3. The objective of the chopper stabilization technique described in this paper is the reduction of this noise source by shifting the noise energy to a higher frequency outside the passband.

The third important noise contribution in the filter is the broad-band thermal noise of the operational amplifier, corresponding to the flat portion of the noise curve shown in Fig. 2. The analysis of the thermal noise of the operational amplifier is complicated by the fact that the bandwidth over which this noise energy lies is much broader than the sampling rate. As a result, the sampling of the next integrator stage causes a portion of the high frequency components of this noise to alias into the passband. The effect of this noise can most easily be visualized by dividing the frequency spectrum into two parts: the range from dc to the unity-gain frequency of the operational amplifier, and the range above the unity-gain frequency of the operational amplifier. For the former, the operational amplifier behaves like a low impedance voltage source when viewed from its own output with an equivalent noise voltage equal to the equivalent input thermal noise voltage of the operational amplifier. Thus, when sampled by the switch and sampling capacitor of the next stage, this energy will be aliased into the passband, and under the assumption that the equivalent input noise is flat in this frequency range and that the time constant of the sampling circuit is long compared to the operational amplifier bandwidth, the effective equivalent input noise density can be found approximately by simply multiplying the broad-band noise energy density by the ratio of the unity-gain frequency of the operational amplifier to the sampling frequency. The effective in-band noise can be represented by inserting a white noise source of this value in series with the output of the integrator, to

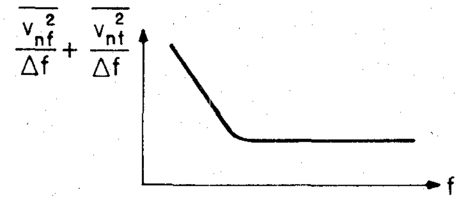

Fig. 2. Typical input equivalent noise voltage spectrum for an MOS operational amplifier.

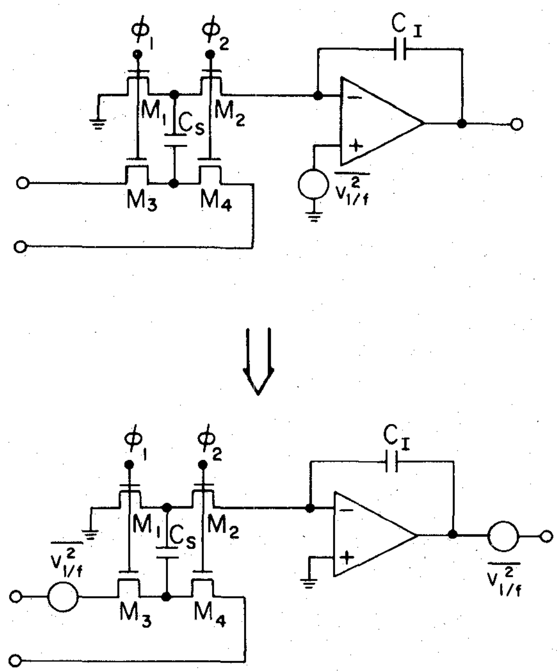

Fig. 3. Representation of operational amplifier $1/f$ noise in the switched-capacitor integrator.

represent the aliasing due to the sampling of the next integrator, and a second source of the same value in series with the input of the integrator. The latter results from the fact that the broad-band noise appears at the summing node of the operational amplifier as well as at the output, and this results in a aliasing of the broad-band noise by the sampling capacitor of the integrator itself.

For frequencies beyond the unity-gain frequency, the amount of noise aliasing that occurs depends on the design of the operational amplifier. In this range, the gain of the operational amplifier is less than unity, and the effect of the feed-band loop around the operational amplifier is negligible. Thus, viewed from its output, the operational amplifier has an output impedance and a noise equivalent resistance which is determined by the circuitry between the compensation point and the output node in the amplifier. If the operational amplifier can be designed so that in this range the output resistance and the noise equivalent resistance at the output are the same, then the increased thermal noise is offset by the fact that the bandwidth over which the noise is sampled by the next stage is reduced by the output resistance, and no additional noise contribution over and above that of the kT/C noise. This situation can be achieved or approximated in one and two stage operational amplifiers which have no output stage and drive the output directly from a high impedance compensation

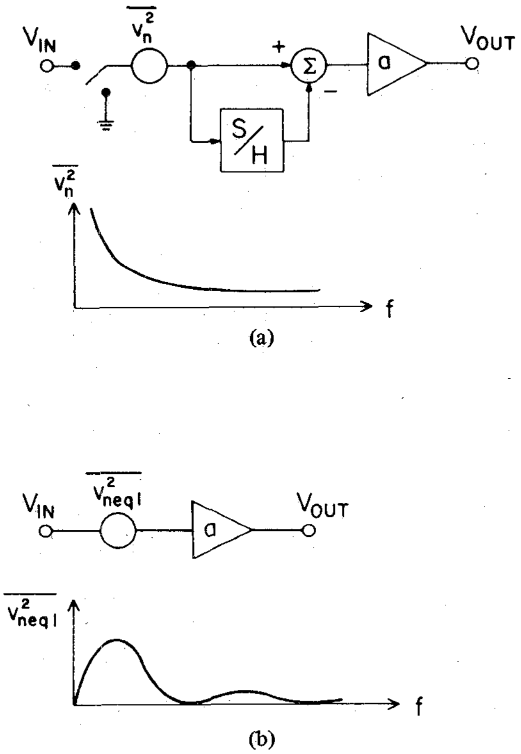

Fig. 4. (a) Concept of correlated double sampling noise reduction. (b) Equivalent input noise of the circuit in (a).

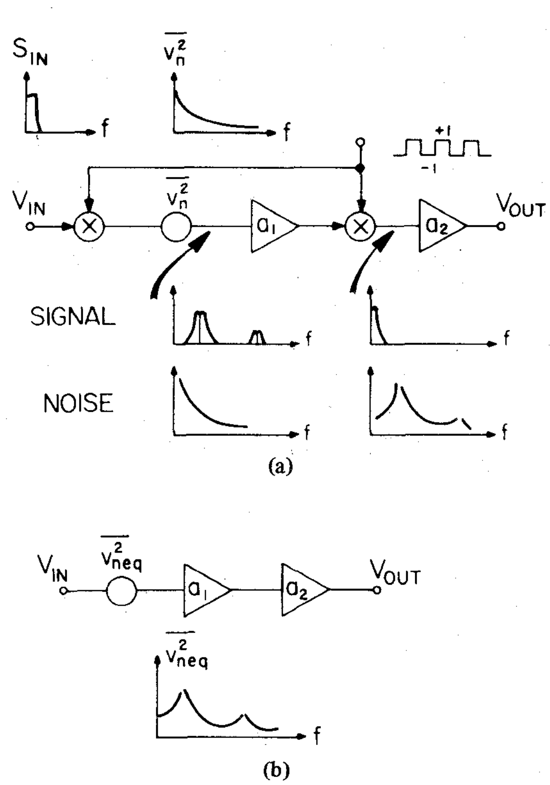

Fig. 5. (a) Concept of chopper stabilization. (b) Equivalent input noise for the circuit in (a).

node. However, in operational amplifiers which utilize some form of output stage, the output resistance is often lower than the equivalent noise resistance in this frequency range, and in this case the operational amplifier thermal noise contribution can be greatly increased because the output noise can be sampled with a large bandwidth.

III. Techniques for the Reduction of $1/f$ Noise

For voiceband applications of switched-capacitor filters, the dominant noise source is often the $1/f$ noise component of the operational amplifier. The $1/f$ noise can be reduced by a number of different methods. One approach is to simply use large input device geometries to reduce the $1/f$ noise associated with these devices. This approach has been widely used in the past, and works particularly well in process technologies which have a low level of surface states at the outset. For processes which have high surface state densities, however, this approach can give uneconomically large input transistor geometries for applications requiring extremely high dynamic range. A second approach is to use buried channel devices so as to remove the channel from the influence of surface states. This approach requires process steps which are not usually included in the standard LSI technologies used to manufacture switched-capacitor filters in high volume. A third approach is to use circuit techniques to translate the noise energy from the baseband to some higher frequency so that it does not contaminate the signal. This approach is the subject of this paper.

One technique for reducing the $1/f$ noise density at low frequencies is the correlated double sampling (CDS) method [8]. This technique is illustrated conceptually in Fig. 4. If the sample/hold function and the subtractor could be easily incorporated in the operational amplifier without adding additional energy storage elements, then this would be a very desirable approach to $1/f$ noise reduction. However, a practical problem with CDS is that the equivalent input noise is most easily obtained by shorting the input and output nodes of the operational amplifier. This operation requires that the output node of the amplifier slew back and forth between the signal level and the initialized level each clock period. This puts a severe requirement on the operational amplifier settling time.

An alternate approach to $1/f$ noise reduction is chopper stabilization. This technique has been used for many years in the design of precision dc amplifiers. The principle of chopper stabilization is illustrated in Fig. 5. Here, a two-stage amplifier and a voiceband input signal spectrum are shown. Inserted at the input and the output of the first stage are two multipliers which are controlled by a chopping square wave of amplitude +1 and −1.

After the first multiplier, the signal is modulated and translated to the odd harmonic frequencies of the chopping square wave, while the noise is unaffected. After the second multiplier, the signal is demodulated back to the original one, and the noise has been modulated as shown in Fig. 5. This chopping operation results in an equivalent input noise spectrum which is shown in Fig. 5(b), where the $1/f$ noise component has been shifted to the odd harmonic frequencies of the chopping square wave. The $1/f$ noise density at low frequencies is now equal to the "folded-back" noise from those harmonic $1/f$ noise components. Therefore, if the chopper frequency is much higher than the signal bandwidth, the $1/f$

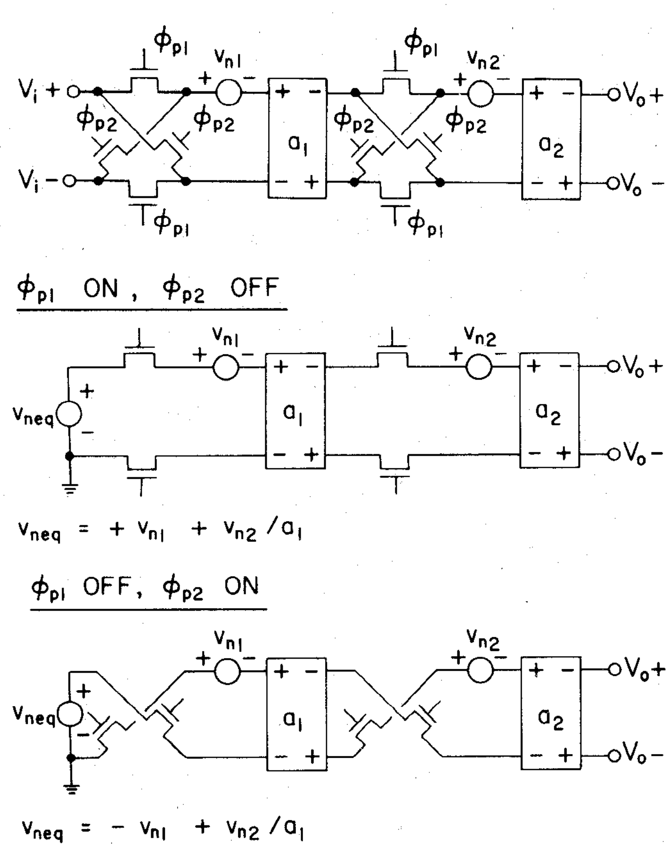

Fig. 6. MOS implementation of a chopper-stabilized operational amplifier.

noise in the signal band will be greatly reduced by the use of this technique.

An MOS implementation of the chopper stabilization technique is shown in Fig. 6. The multipliers described before are realized by two cross-coupled switches which are controlled by two nonoverlapping clocks. When ϕ_{p1} is on and ϕ_{p2} is off, the equivalent input noise is equal to the equivalent input noise of the first stage plus that of the second divided by the gain of the first stage. When ϕ_{p1} is off and ϕ_{p2} is on, the equivalent input noise is equal to the negative of this instantaneous value. If the voltage gain of the first stage is high enough, the noise contribution from the second stage can be neglected and the sign of this equivalent input noise changes periodically.

IV. Differential Filter Implementation

The dynamic range of a switched-capacitor filter is determined by the ratio of the maximum signal swing giving acceptable distortion to the noise level. Thus, improvements in signal swing result in direct improvement of the dynamic range. In the experimental filter described in this paper, the effective signal swing is doubled relative to a conventional switched-capacitor filter by the use of a differential output integrator. The differential configuration used has the additional advantage that because the signal path is balanced, signals injected due to power supply variations and clock charge injection are greatly reduced.

An example of a bottom-plate fully-differential switched-capacitor integrator is shown in Fig. 7. Two differential input signals are connected to four input nodes of this integrator. Signals v_{i1}^+ and v_{i1}^- form one differential input, and v_{i2}^+ and v_{i2}^- form another differential input. Assuming that the two differential inputs have no common-mode component, the common-mode voltage at the operational amplifier input is set by V_B. This potential can be set with on-chip circuitry to a value which is convenient from the standpoint of the design of the operational amplifier, and can be used to eliminate the necessity of a level shift function within the operational amplifier. In addition to the fact that dynamic range and power supply rejection are improved with this configuration, the fact that both output polarities are available makes the design of elliptic ladder filters much easier.

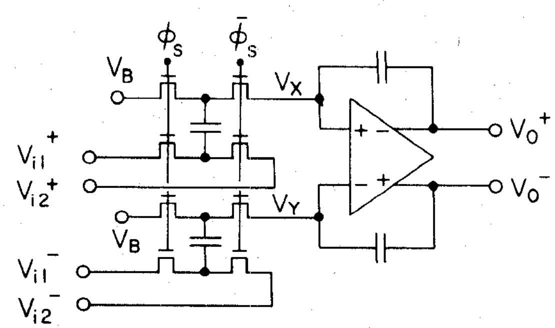

Fig. 7. Fully differential switched-capacitor integrator.

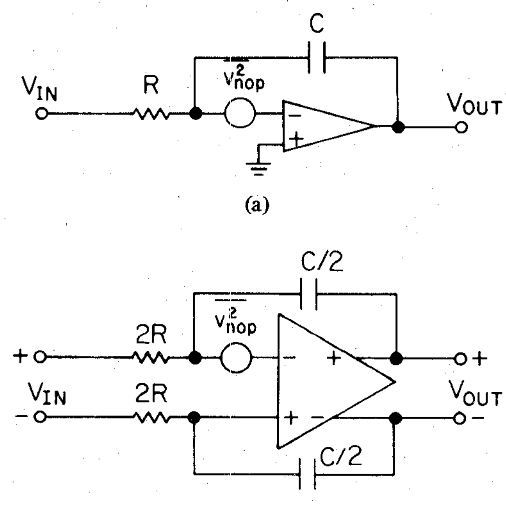

Fig. 8. Signal-to-noise comparison between single-ended and differential integrators.

Two disadvantages to this technique are that the interconnection problem makes circuit layout more complex, and that a differential to single-ended conversion may be necessary in some applications.

The improvement in dynamic range resulting from the use of the differential technique is illustrated in Fig. 8. In both single-ended and differential cases, the transfer functions for the signal from the input to the output are identical. For the operational amplifier noise, the transfer functions from the noise source to the output of the integrator are identical. If the operational amplifier noise is dominant, the result is a 6 dB improvement in S/N ratio, because, for the same power supplies, the effective voltage swing doubles in the differential case. If the thermal noise in the resistors is dominant, the

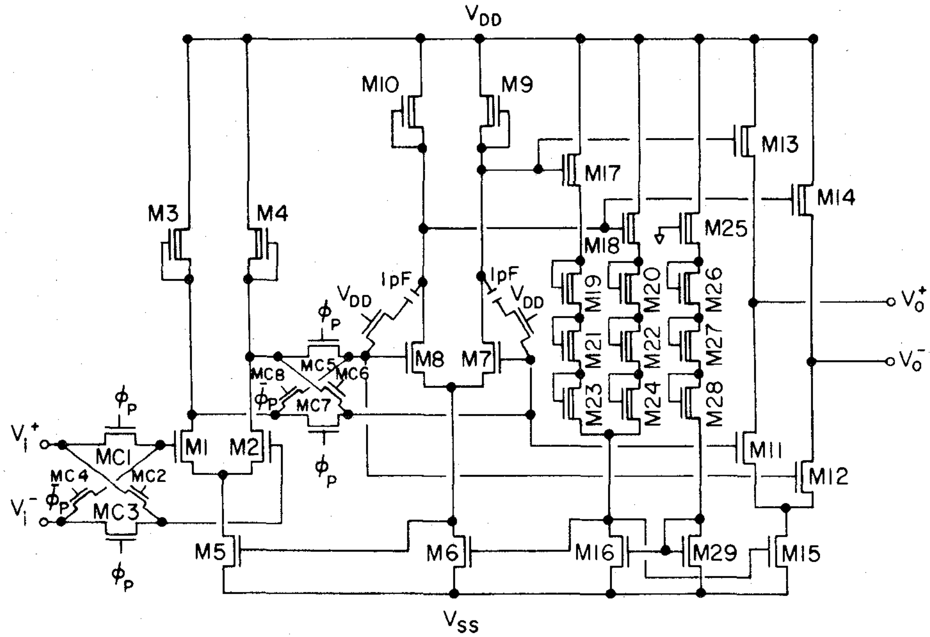

Fig. 9. Schematic diagram of differential chopper-stabilized operational amplifier.

dynamic range is the same for the same amount of total integrating capacitance in both cases.

V. Experimental Differential Chopper-Stabilized Operational Amplifier and Filter

An experimental differential switched-capacitor low-pass filter was fabricated to investigate the use of chopper stabilization for low frequency noise reduction. The operational amplifier will be described first, then the filter configuration used.

The key design problem in the implementation of a filter of the type described above is the realization of an operational amplifier which has differential outputs with a well-defined common-mode voltage, and which incorporates the chopper stabilization without undue complexity. While in principle chopper stabilization could be implemented without the simultaneous use of the differential configuration, the differential configuration is particularly amenable for the implementations of the chopper circuitry using balanced cross-coupled analog switches as shown in Fig. 9. Transistors *MC*1-*MC*4 and *MC*5-*MC*8 form two cross-coupled choppers which are controlled by two nonoverlapping clocks. Transistors *M*1-*M*5, *M*6-*M*10, and *M*11-*M*15 are the input, gain, and output stages, respectively. The operating points of the input stage are biased by the common-mode feedback loop through the gain stage. The common-mode output dc voltage and the operating points of the gain stage are set from common-mode feedback circuit *M*16-*M*24, which is biased by a grounded replica reference string consisting of transistors *M*25-*M*29, and uses depletion transistors *M*17-*M*24 as a level shifter to ensure a well-defined common-mode output dc voltage for large differential output swings. The common-mode dc voltage at the input of the operational amplifier can be obtained through the bias voltage V_B. Pole splitting compensation in the gain stage and a feedforward path to the output stage (*M*11 and *M*12) are used to improve the stability of the operational amplifier when driving a large capacitance load.

The schematic diagram of a bottom-plate fifth-order differential low-pass ladder switched-capacitor filter is shown in Fig. 10. This configuration is used to implement a differential Chebyshev low-pass filter with cutoff frequency at 3400 Hz. This circuit is fully differential-in and differential-out. Five differential chopper-stabilized operational amplifiers with chopper frequency 128 kHz and master clock frequency of 256 kHZ are used. Thus, the $1/f$ noise is translated by the chopper to exactly the Nyquist rate of the filter clock. As a result, no aliasing effect of the $1/f$ noise back into the passband occurs. All the $1/f$ noise components will appear in the frequency domain, equally spaced between two adjacent integer multiples of the filter clock frequencies. The odd harmonic frequencies of the chopper clock, where the shifted $1/f$ noise located, do not interfere with the master filter clock frequency. This choice of chopping frequency can be shown to be optimum in terms of maximum reduction in baseband $1/f$ noise.

The noise at the output of a switched-capacitor filter can be calculated by knowing the transfer function from each noise source (or equivalent noise source) to the output of the filter. This calculation was carried out for the experimental filter by first measuring the equivalent input noise spectrum of the operational amplifier, and then numerically summing the various contributions weighted by the transfer function from that point in the filter to the output. Aliasing of broad-band noise was taken into account as described in Section II. The resulting predicted noise spectrum is shown in Fig. 11, with and without chopper switches operating. As expected, the dominant noise mechanism without chopper stabilization is the $1/f$ noise component. With chopper operating, the dominant noise in this particular filter is the operational amplifier thermal noise. This results primarily from the choice of cur-

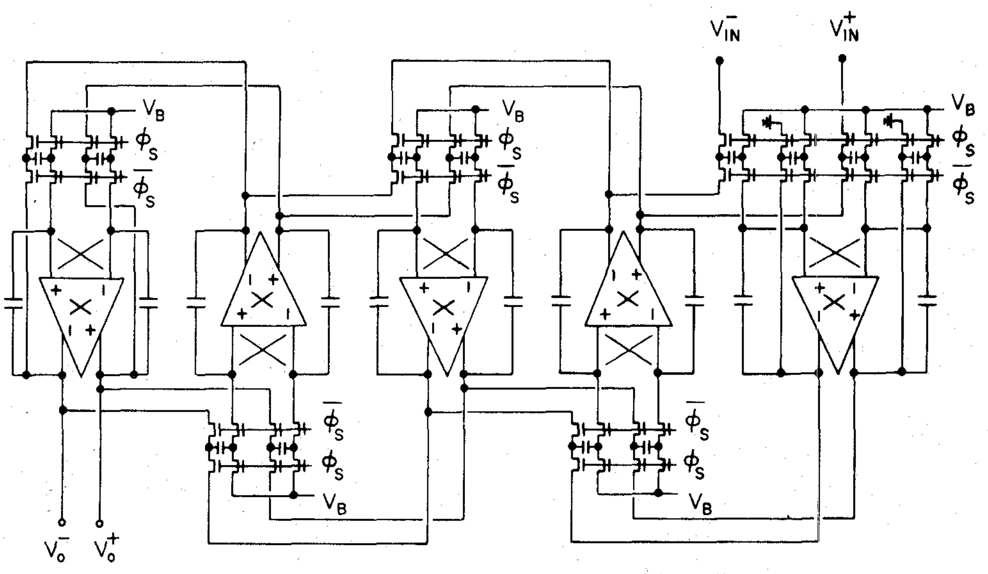

Fig. 10. Differential fifth-order chopper-stabilized filter.

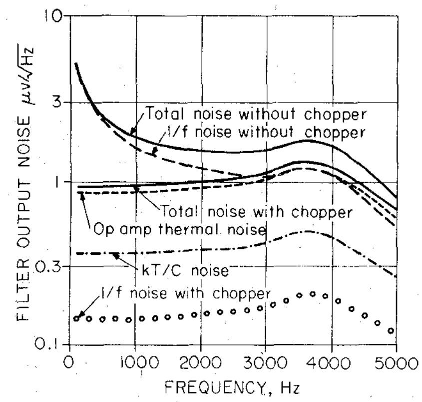

Fig. 11. Theoretical noise performance of the experimental filter, showing each of the three noise contributions with and without chopper stabilization.

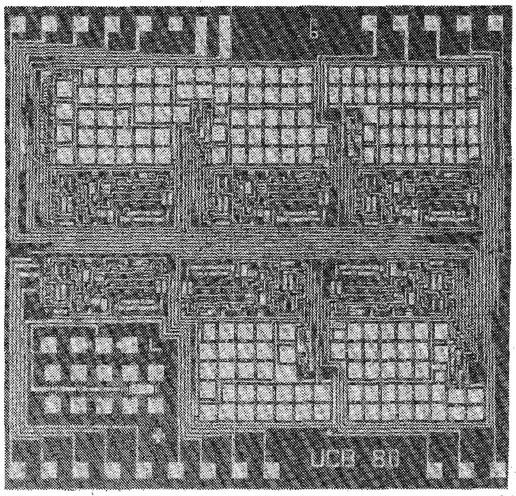

Fig. 12. Die photomicrograph, experimental fifth-order filter.

rent level and W/L ratio in the input transistors of the operational amplifiers, giving a relatively low g_m and a relatively large value of noise equivalent resistance in the frequency range from dc up to the operational amplifier cutoff frequency of 15 MHz. A straightforward modification of this design would give a reduction in this noise source of 8-10 dB, giving the desired result that filter noise be dominated by thermal noise in the switches.

Note that the on resistance of the switches in the chopper in series with the input of the operational amplifier do contribute to the thermal noise of the operational amplifier. This contribution could be significant in a case in which the inherent noise equivalent resistance of the amplifier was very low.

VI. Experimental Results

An experimental prototype differential chopper-stabilized fifth-order Chebyshev filter was designed and fabricated using a metal-gate n-channel MOS process with depletion load. The minimum transistor gate length in the process used was 15 μ. One objective of the experimental filter was to explore the maximum achievable dynamic range, and toward that end a relatively large integrating capacitance averaging 100 pF integrator was chosen. The filter area is about 9600 mils^2. A die microphotograph is shown in Fig. 12.

The experimental equivalent input noise response of the on-chip test operational amplifier is shown in Fig. 13. The dashed curve is the input referred noise response without chopper stabilization (i.e., the chopper is kept in one of the two possible states). The bottom solid curve is the input referred noise response with chopper frequency at 128 kHz. Notice that in this case the noise at 1 kHz is primarily due to the first folded-back 128 kHz harmonic $1/f$ noise and the thermal noise of the operational amplifier. The total is about 40 dB (100 times in power) less than that without chopper stabilization.

In order to demonstrate further the translation of the $1/f$ noise, the noise was measured with the chopper frequency at one eighth of its nominal value, or 16 kHz, with the result shown as the upper solid curve in Fig. 13. Notice that the noise at 1 kHz is about 12 dB (16 times in power) less than that without chopper stabilization. Also, the $1/f$ noise peak has been shifted to the odd harmonic frequencies of the chopper clock, which are 16 kHz and 48 kHz in Fig. 13. The

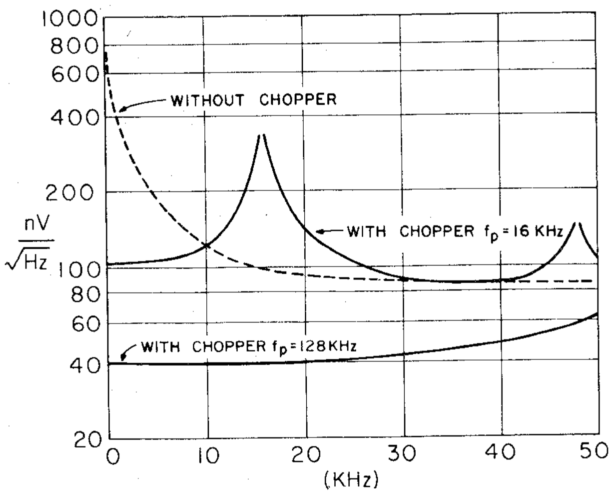

Fig. 13. Experimentally observed operational amplifier equivalent input noise for various chopping frequencies.

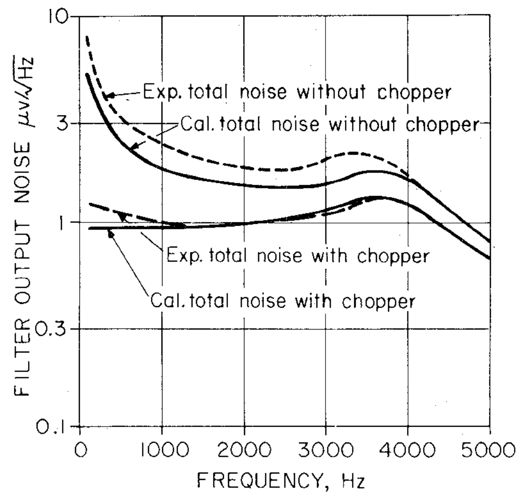

Fig. 15. Experimentally observed filter output noise with and without chopper, compared to theoretically predicted total noise.

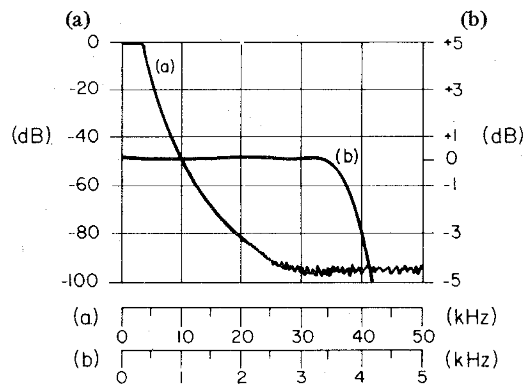

Fig. 14. Experimental filter frequency response. (a) Overall response. (b) Passband response.

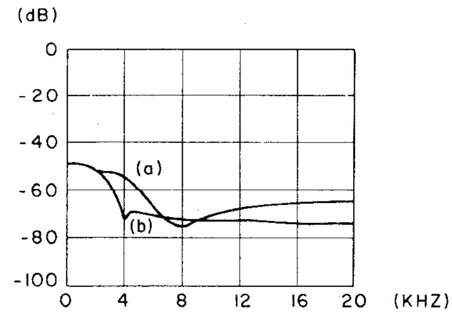

Fig. 16. Experimentally observed PSRR. (a) From positive power supply. (b) From negative power supply.

operational amplifier displayed a gain-bandwidth product of 15 MHz, and power dissipation of 4 mW with + and -7.5 V supplies.

The observed filter frequency response is shown in Fig. 14. The experimental filter output noise response with and without chopper stabilization is shown in Fig. 15. The noise reduction at 1 kHz is about 10 dB. This experimental result agrees closely with the predicted total output spectrum, which is shown on the same graph. The $1/f$ noise is dominant in the filter without chopper operation, and the thermal noise of the operational amplifier is dominant with chopper operation. Despite a 40 dB noise reduction at 1 kHz in the operational amplifier noise when the chopper is operated, only 10 dB of improvement is obtained in the filter noise. This occurs because the aliased thermal noise of the operational amplifier becomes the dominant noise mechanism when the chopper is on.

The experimental positive and negative power supply rejection ratios (PSRR's) for frequency up to 20 kHz are shown in Fig. 16. The minimum PSRR is 50 dB for both supplies in this frequency range. The filter power dissipation is 20 MW. For plus/minus 7.5 V power supplies at 25°C, the experimental filter had a maximum differential output signal swing at 1 kHz and 1 percent total harmonic distortion (THD) equal to 5 V rms. The filter output noise with C-message weighted was 40 μV rms, giving a dynamic range of 102 dB.

VII. Conclusions

A differential chopper-stabilized switched-capacitor filtering technique has been described, which extends the dynamic range of the switched-capacitor filters beyond 100 dB. The chopper stabilization greatly reduces the $1/f$ contribution to total filter noise, and is most applicable in applications where very large dynamic range is an objective. The differential configuration results in substantial improvements in power supply rejection ratio. This improvement is of considerable practical significance since power-supply related noise and crosstalk is a major problem in critical applications.

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