

AN ABSTRACT OF THE THESIS OF

Shenoy Ravindra Umanath for the degree of Master of Science in Electrical and Computer Engineering presented on June 3, 1988.

Title: A CAD Noise Model for Chopper-Stabilized Switched-Capacitor Filters

Abstract approved: **Redacted for Privacy**

/ David J. Allstot

A modeling technique for simulating the noise performance of chopper-stabilized switched-capacitor filters using SPICE2 is presented. A method for selection of switched-capacitor network (SCN) stages to be chopped and an op amp topology for low noise applications are also outlined. Validity of the modeling is verified by the close agreement (3 dB) between the predicted and measured noise responses.

A CAD Noise Model for Chopper-Stabilized
Switched-Capacitor Filters

by

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A THESIS

submitted to

Oregon State University

in partial fulfillment of
the requirements for the degree of

Master of Science

Completed June 3, 1988

Commencement June 1989

APPROVED:

Redacted for Privacy

Professor of Electrical and Computer Engineering, in
charge of major

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Engineering

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Date thesis is presented October 25, 1982

Typed by Joan Higgins for Shenoy Ravindra U.

ACKNOWLEDGEMENTS

First and foremost, I wish to acknowledge the help, guidance and patience of Dr. David J. Allstot, my major professor--a great teacher and an excellent researcher. Many thanks are due to Magnavox Corporation which supported this project financially.

My parents and Shawneen gave me moral support, for which I am eternally grateful.

I also wish to thank my graduate committee and the faculty at Oregon State University for their consideration and their time.

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A CAD NOISE MODEL FOR CHOPPER-STABILIZED SWITCHED-CAPACITOR FILTERS

INTRODUCTION

Chopper-stabilization is a frequency modulation technique which has been successfully employed in Switched-Capacitor (SC) filters to eliminate the effects of low-frequency MOSFET flicker noise [1], [2]. Unfortunately, a simple technique for simulating the expected performance of chopper-stabilized SC filters has not been available. This thesis presents a modeling technique for simulating the noise performance of chopper-stabilized SC filters using the SPICE2 program.

The developed noise model is flexible enough to be used for both unchopped as well as chopped SC filters. All noise sources including op amp, capacitor switching and the effects of chopper-stabilization are included. Comparison of the predicted noise with the measured values verifies the model to be accurate to within 3 dB.

Voice-band communication systems often require

a very small input-referred noise figure even at the low frequency end of the noise spectrum. It is not enough to eliminate the $1/f$ noise component in these SC filters using the chopper-stabilization technique. Operational amplifiers with an inherently low white noise component have to be used in these SC systems. In the sections to follow, a modified cascode op amp topology suitable for use as a chopper amplifier and a method for selection of SC network stages to be chopped are also outlined.

DEVELOPMENT OF THE SWITCHED-CAPACITOR FILTER NOISE MODEL

Any noise model for SC filters should be flexible enough to be used both for the unchopped as well as the chopper-stabilized case. This is especially essential when analyzing for example a sixth-order SC filter where only three of the six stages are chopped. The noise model developed in the sections to follow is based on the previous work of Fischer [3] which works well for unchopped SC filters.

UNCHOPPED SWITCHED-CAPACITOR FILTER NOISE MODEL

The basic noise sources in any SC filter are as follows:

- (1) OP amp noise sources which are of two kinds:
 - (a) Flicker noise or $1/f$ noise
 - (b) Broadband thermal noise
- (2) Capacitor switching noise.

The op amp to be modeled is shown in Fig. 1. Referring to the input-referred noise plot of Fig. 1b, the point of inflexion is defined as the $1/f$ corner frequency of the op amp. Below the corner frequency, the dominant noise is the flicker noise and above the corner frequency, broadband noise dominates.

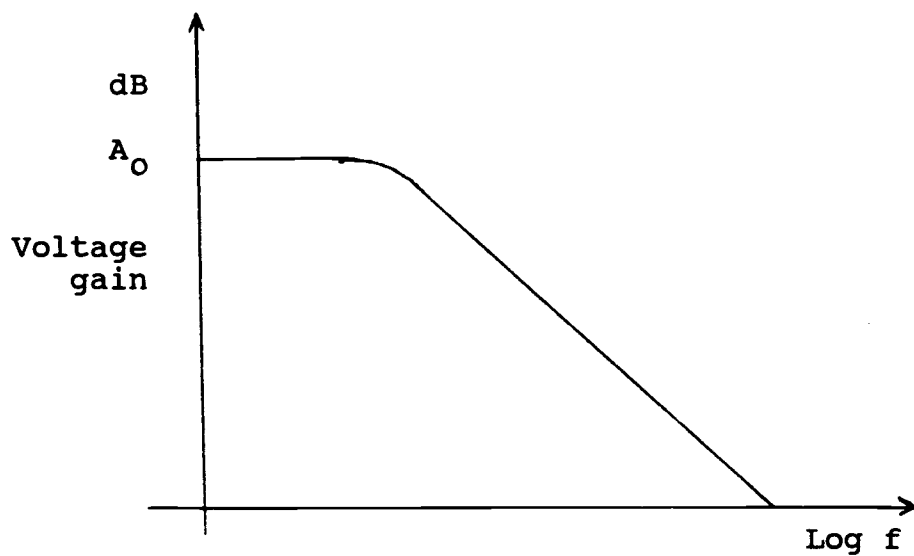


Fig. 1a. Magnitude plot of the op amp to be modeled.

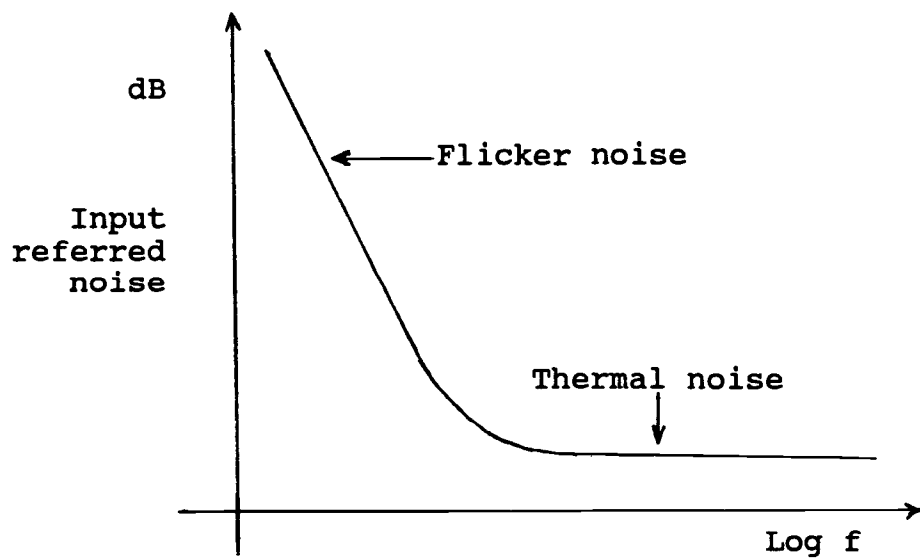


Fig. 1b. Noise plot of the op amp to be modeled.

Flicker noise is caused by the surface states in the channel regions of the MOSFETs. Therefore this noise source is inherently process dependent. The magnitude of flicker noise is also a function of the device geometries:

$$V_{N1/f}^2 = \frac{KF (I_{DS})^{AF}}{WLC_{ox}kf} \dots\dots\dots(1)$$

where KF is the flicker noise coefficient

AF is the flicker noise exponent

I_{DS} is the dc drain current of the MOSFET

WL are the device geometries

C_{ox} is the oxide capacitance

k is the MOSFET conductance, $k = \mu C_{ox}$

μ is the carrier mobility in the channel region.

To model the 1/f noise component of an op amp, a MOSFET is used. The device is diode-connected (for ease of biasing) to a pre-determined value of drain current by the current source I_{DS} . To assure that the device is in saturation, M1 is chosen as an enhancement mode type. Fig. 2 shows the basic op amp noise model, where M1 is flicker noise source. The noise voltage spectral density seen at the drain of M1 is given by:

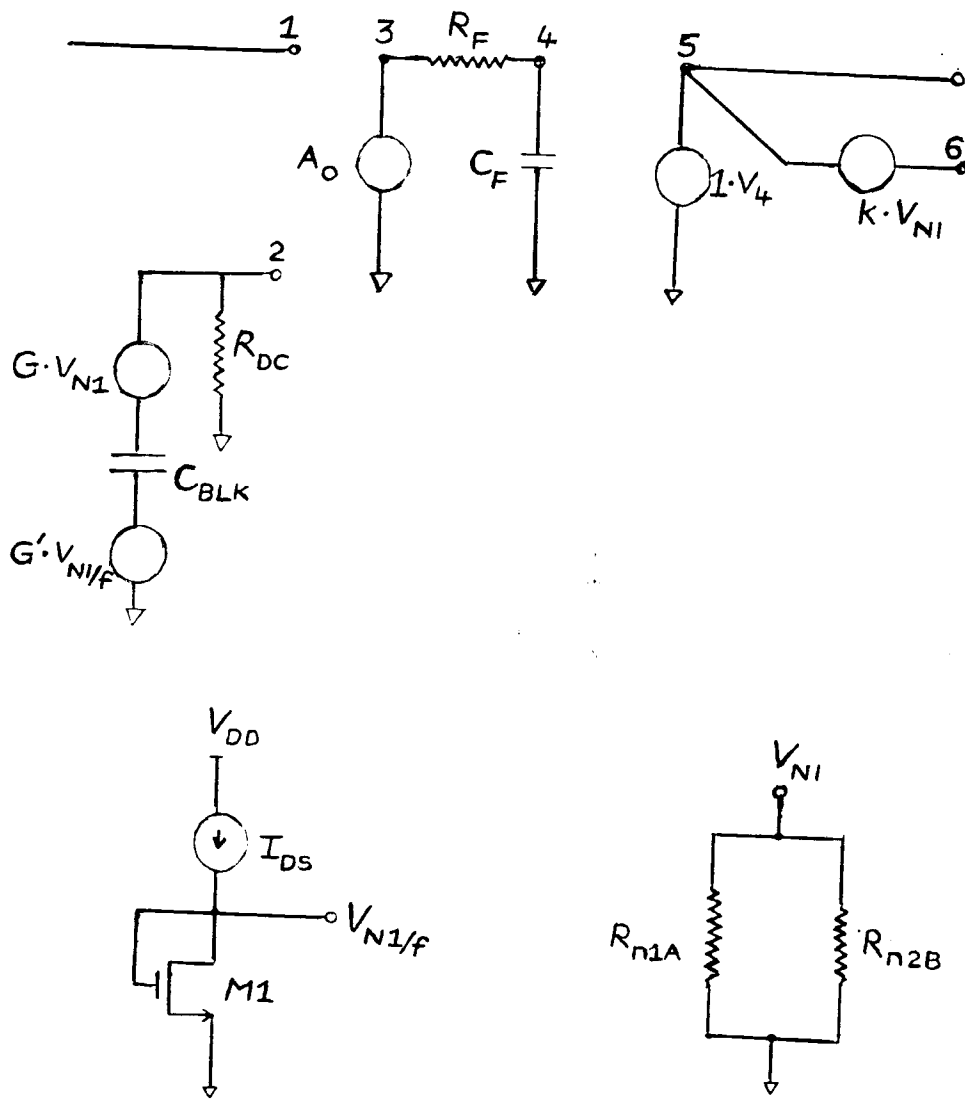


Fig. 2. Basic Op amp noise model.

A_O is the dc gain of the op amp

R_F, C_F form the low pass filter to simulate the
one pole response of the op amp

R_{DC}, C_{BLK} form the dc blocking circuit

$V_{N1/f}$ is the flicker noise source

V_{N1} is the broadband thermal noise source

$M1$ is the flicker noise generator

R_{n1A}, R_{n2B} are resistors to generate the white
noise

G is the gain factor of the broad-band thermal noise
source

k is the folded flat-band factor

node (1) is the inverting input of the op amp

(2) is the non-inverting input of the op amp

(5) is the continuous output of the op amp

(6) is the switched output of the op amp

Fig. 2, Basic op amp noise model.

$$V_N = \frac{(\text{Thermal noise component} + \text{1/f noise component})^{1/2}}{\sqrt{f}}$$

$$= \left[\frac{8kT}{3g_m} + \frac{KF(I_{DS})^{\alpha F}}{WLC_{OX}kf} \right]^{1/2} \dots\dots\dots(2)$$

Since this device is to serve as the 1/f noise source it is imperative to select the bias current and device size to assure that the g_m or transconductance term contributes much less noise than the 1/f noise term. With device parameters selected, adjust KF to match the input-referred 1/f noise curve (Fig. 1b) of the op amp.

Because the flicker noise is to be injected into the non-inverting terminal of a high gain op amp, the dc bias voltage at the drain should be blocked. R_{DC} and C_{BLK} form the dc blocking circuit and a voltage-controlled voltage source (VCVS) is used as a buffer with a gain factor $G'=1$.

Broadband thermal noise is also referred to as flat-band noise because its magnitude remains constant and is independent of the frequency at which it is measured. Hence the thermal noise component of the op amp is modeled as a white noise source, which

can be easily simulated using a resistor in SPICE.

The resistance value is selected to match the measured or calculated input-referred flat-band noise of the op amp.

$$V_{N\text{flatband}}^2 = 4kTR \Delta f \quad \text{.....(3)}$$

where k is the Boltzman's constant

T is the absolute temperature

R is the matched resistor value

To satisfy the nodal requirements of SPICE, two resistors (R_{n1A} and R_{n2B} of Fig. 2), each twice the value of R , are paralleled so that all nodes have at least two components connected and a dc path to ground. These resistors in conjunction with the VCVS V_{N1} (with a gain factor of $G=1$) simulate the op amp broadband thermal noise component.

Sampling the SC filter at a frequency f_{samp} , produces a folded flat band noise component which is modeled using resistors R_{n1A} and R_{n2B} and the VCVS source V_{N1} . This noise source is placed at the output of the op amp and has a gain factor (folded flat-band factor) given by

$$k = \left[\frac{2BW_n}{f_{\text{samp}}} - 1 \right]^{1/2} \quad \text{.....(4)}$$

where BW_n is the equivalent noise bandwidth of the op amp.

An explanation of this folded flat-band effect is discussed in the Appendix.

To simulate the capacitor switching noise, the switched capacitors of the filter are replaced with resistors of value

$$R \approx \frac{1}{C f_{\text{samp}}} \quad \text{.....(5)}$$

where C is the value of the switched capacitor. To model the op amp magnitude plot of Fig. 1, a simple RC low pass filter and a VCVS are used as shown in Fig. 2. The VCVS sets the dc gain of the op amp and the low pass filter simulates the one pole response. The op amp model shown in Fig. 2 has two kinds of outputs. The switched output (S_0) is connected to all the switched capacitors connected to the output of the op amp to be modeled in a SC filter. The folded flat-band noise source, kV_{N1} , is placed at this output. The continuous output (C_0) is connected to the paths that are not switched.

The basic op amp noise model for unchopped SC filters has been modeled; in the next section it will

be extended to include the effects of chopper-stabilization.

B. CHOPPER-STABILIZED SWITCHED-CAPACITOR FILTER NOISE MODEL

For voice-band applications of SC filters, the dominant noise source is often the $1/f$ noise component of the op amp. 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 used in the past and works well in process technologies which have a low level of surface states at the outset. Still, this implies larger die areas which is uneconomical. A second approach is to use buried-channel devices so as to remove the channel from the influence of the surface states. This approach requires processing steps which are not included in the standard LSI technologies used to manufacture SC 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. In this section, this technique (chopper-stabilization) is analyzed and a noise model for chopped SC filters is developed. The concept of chopper-stabilization is shown in Fig. 3.

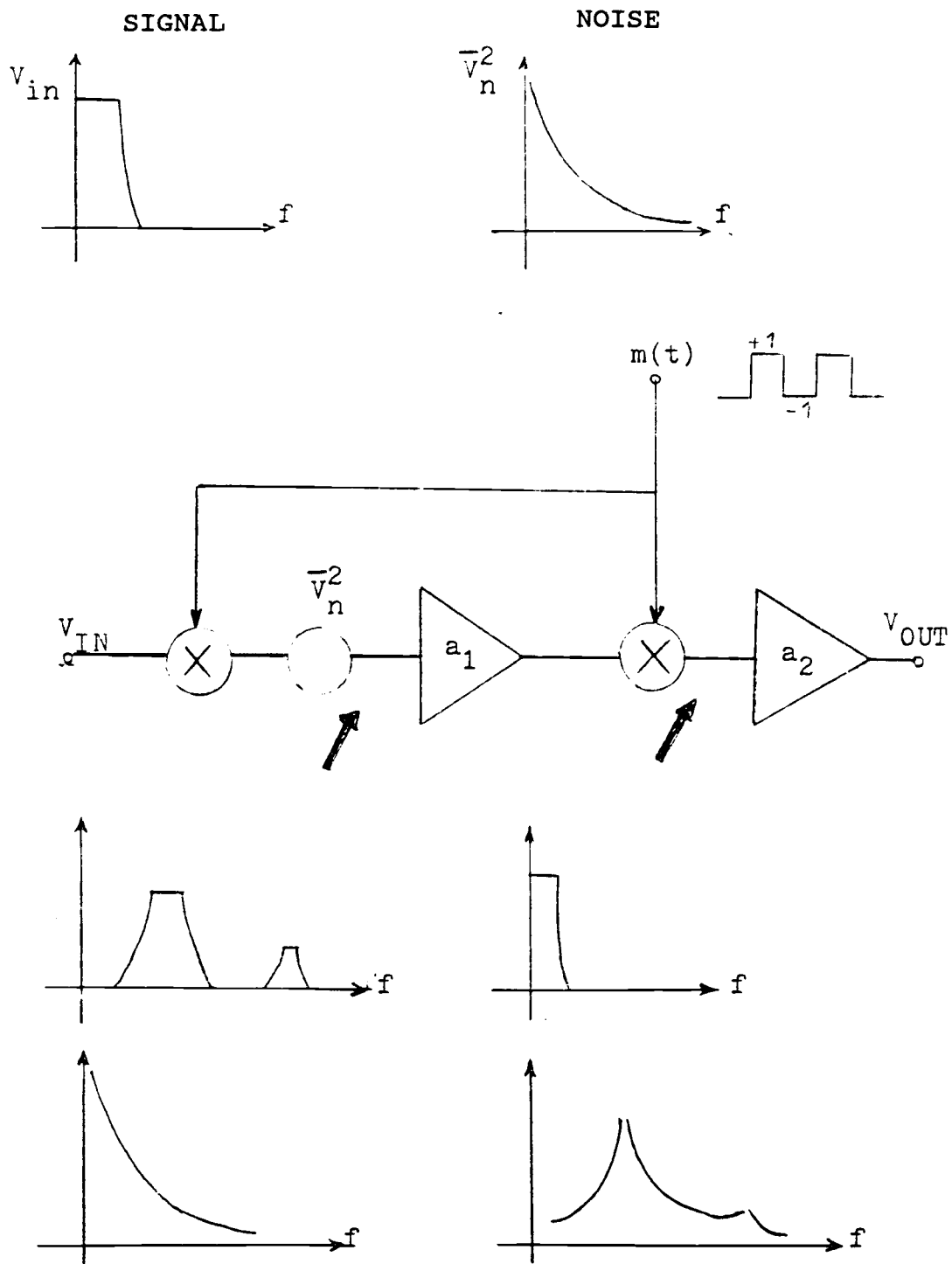


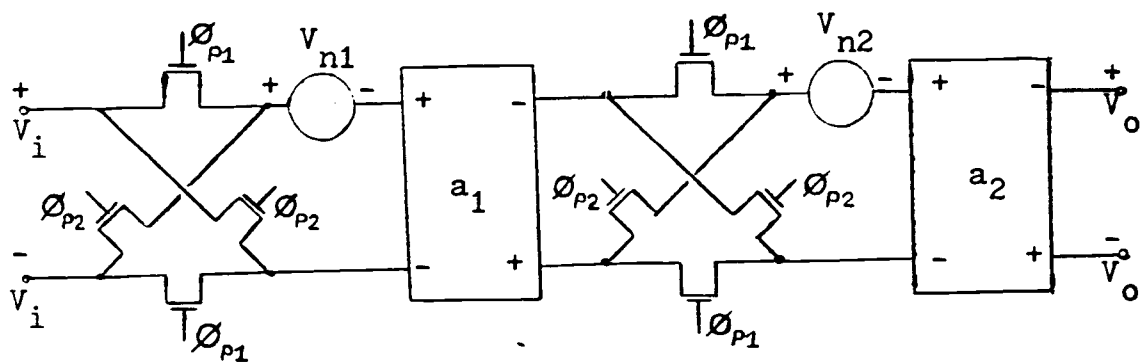
Fig. 3. Concept of chopper-stabilization.

Suppose the input signal has a spectrum limited to half the chopper frequency (to avoid aliasing) and that the amplifier has neither noise nor offset. The input signal is modulated by a square wave signal $m(t)$ with a period $T = 1/f_{\text{chop}}$. Modulation transposes the signal around the odd harmonics of the modulating signal. The signal is then amplified and demodulated back to the original bandwidth. Since the amplifier has a finite bandwidth, the output signal contains spectral components around the even harmonics of the chopper frequencies.

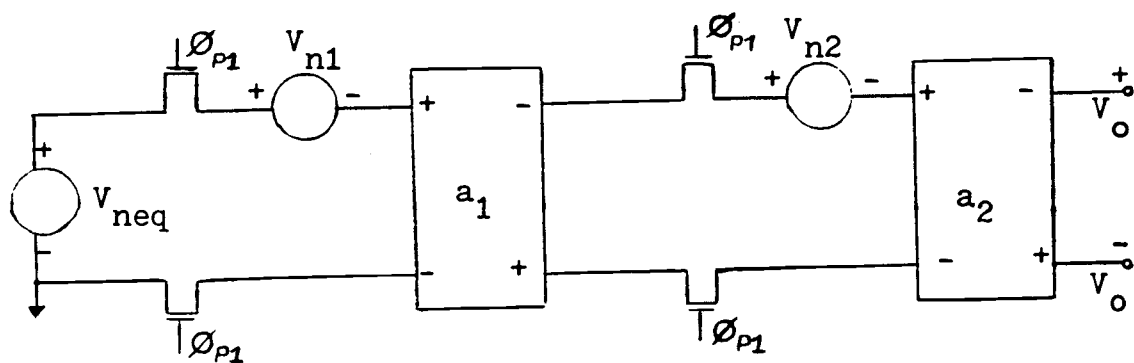
An excellent mathematical; treatment of the chopper technique provided in [4] and a similar approach is adopted in this section.

Fig. 4 shows a two-stage op amp which has a sufficiently high first stage gain a_1 . It is assumed that this amplifier is compensated to have a one pole response and that its transfer function is given by

$$A(f) = \frac{A}{1 + f/f_c} \quad \dots\dots\dots(6)$$

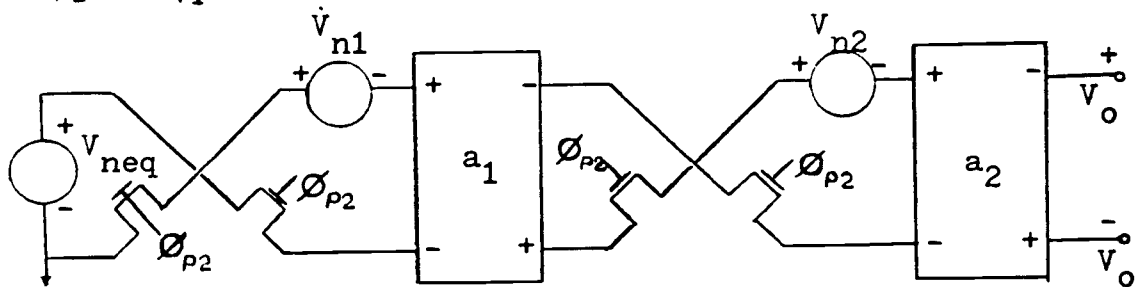


ϕ_{P1} ON, ϕ_{P2} OFF



$$V_{neq} = +V_{n1} - V_{n2}/a_1$$

ϕ_{P2} ON, ϕ_{P1} OFF



$$V_{neq} = -V_{n1} + V_{n2}/a_1$$

Fig. 4. MOS implementation of a chopper-stabilized operational amplifier.

where A_0 is the dc gain of the amplifier

$$A_0 = a_1 a_2$$

f_c is the 3dB cut-off frequency of the amplifier

The equivalent input noise of the amplifier can be expressed as

$$V_N^2 = S_N = S_{Nin} \left(1 + \frac{f_k}{f_{chop}} \right) \dots\dots\dots(7)$$

The noise and offset of the amplifier are only modulated once and translated to the odd harmonics of the chopping square wave. The output noise spectrum is given by

$$S_{Nout} = \left(\frac{2}{\pi} \right)^2 \sum_{\substack{n=1 \\ n \text{ odd}}}^{+\infty} \frac{1}{n^2} A(f-n/T) S_N(f-n/T) \dots\dots\dots(8)$$

where $S_N(f)$ is the amplifier equivalent input noise spectrum given by (7).

For a single pole amplifier as in this case with a transfer function given by (6), the low frequency output noise given by the above expression can be summed and approximated by

$$S_{Nout} = A_0^2 S_{Nin} \left(1 + \frac{17f_k}{2\pi^2 f_{chop}} \right) \dots\dots\dots(9)$$

with the assumption that the amplifier has a sufficient overall gain at the chopper frequency.

The MOS implementation of the chopper-stabilization is shown in Fig. 4. The square wave modulator is realized by two cross-coupled switches which are controlled by two non-overlapping clocks. When ϕ_{p1} is on and ϕ_{p2} is off, the overall 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 overall equivalent input noise is equal to the negative of this instantaneous value. If the voltage gain of the first stage is high enough, the contribution from the second stage can be neglected and the sign of the overall equivalent input noise changes periodically.

From equation (9), it is apparent that for a chopper frequency $f_{\text{chop}} \gg f_k$, the equivalent low frequency input noise of the chopper amplifier is equal to the original amplifier white noise component. It would appear that for effective elimination of the $1/f$ noise component, the necessary condition would be that the chopper frequency be much greater than the corner frequency. However, in

reality there exists a finite amount of residual offset due to the coupling between the modulator, signal source and the amplifier. This residual offset increases as the chopper frequency moves higher on the frequency scale [4].

An increase in the residual offset limits both the sensitivity and precision of the amplifier. Making the chopper frequency almost equal to the corner frequency is an essential trade-off between effective noise reduction and the residual offset value [4].

Since the chopper eliminates the $1/f$ noise, the value of the gain factor G' for the chopper-stabilized SC filter noise model is zero. The expression for the output noise of the chopper reveals that the input or original amplifier white noise component is increased by a factor proportional to the ratio of f_k to f_{chop} . Therefore, in the chopper SC noise model, the gain factor for the broadband thermal noise is

$$G = \left(1 + \frac{17 f_k}{2 \pi f_{\text{chop}}} \right)^{1/2} \dots\dots\dots (10)$$

The mathematical analysis of the chopper reveals that the white noise is not aliased by the chopping process. Hence there is no change in the folded flat-band factor for this noise model as compared to the unchopped noise model. Table I lists the values of the gain factors for both the noise models.

TABLE I
GAIN FACTORS OF THE TWO SC NOISE MODELS

GAIN FACTORS	UNCHOPPED SC NOISE MODEL	CHOPPED SC NOISE MODEL
G	1	$\frac{(1 + 17f_k)}{2^2 f_{\text{chop}}})^{1/2}$
G'	1	0
k	$(\frac{2BW_n}{f_{\text{samp}}} - 1)^{1/2}$	$(\frac{2BW_n}{f_{\text{samp}}} - 1)^{1/2}$

A MODIFIED CASCODE OP AMP TOPOLOGY FOR CHOPPER-STABILIZED SC FILTERS

The implementation of the chopper technique is shown in Fig. 3, where a_1 and a_2 are the gains of the two stage op amp. Among the existing CMOS op amp topologies, the Two-Stage and the Folded-Cascode are the most widely used [5]. The former can be used as a chopper and furthermore it has a much lower input noise and offset as compared to a folded-cascode op amp of similar dimensions but it has to be compensated to achieve a one pole response. Compensation is achieved either by the Dominant-pole or the Pole-Splitting compensation techniques. In either case a capacitor is involved which is nominally 15 picofarads. This capacitor consumes a significant amount of die area which is normally limited and since capacitance values usually vary by $\pm 10\%$, the op amp may not be properly compensated.

On the other hand, the folded-cascode needs no compensation but has a one stage configuration which rules out the chopper implementation shown in Fig. 4. The two topologies (without the biasing network) are shown in Fig. 5.

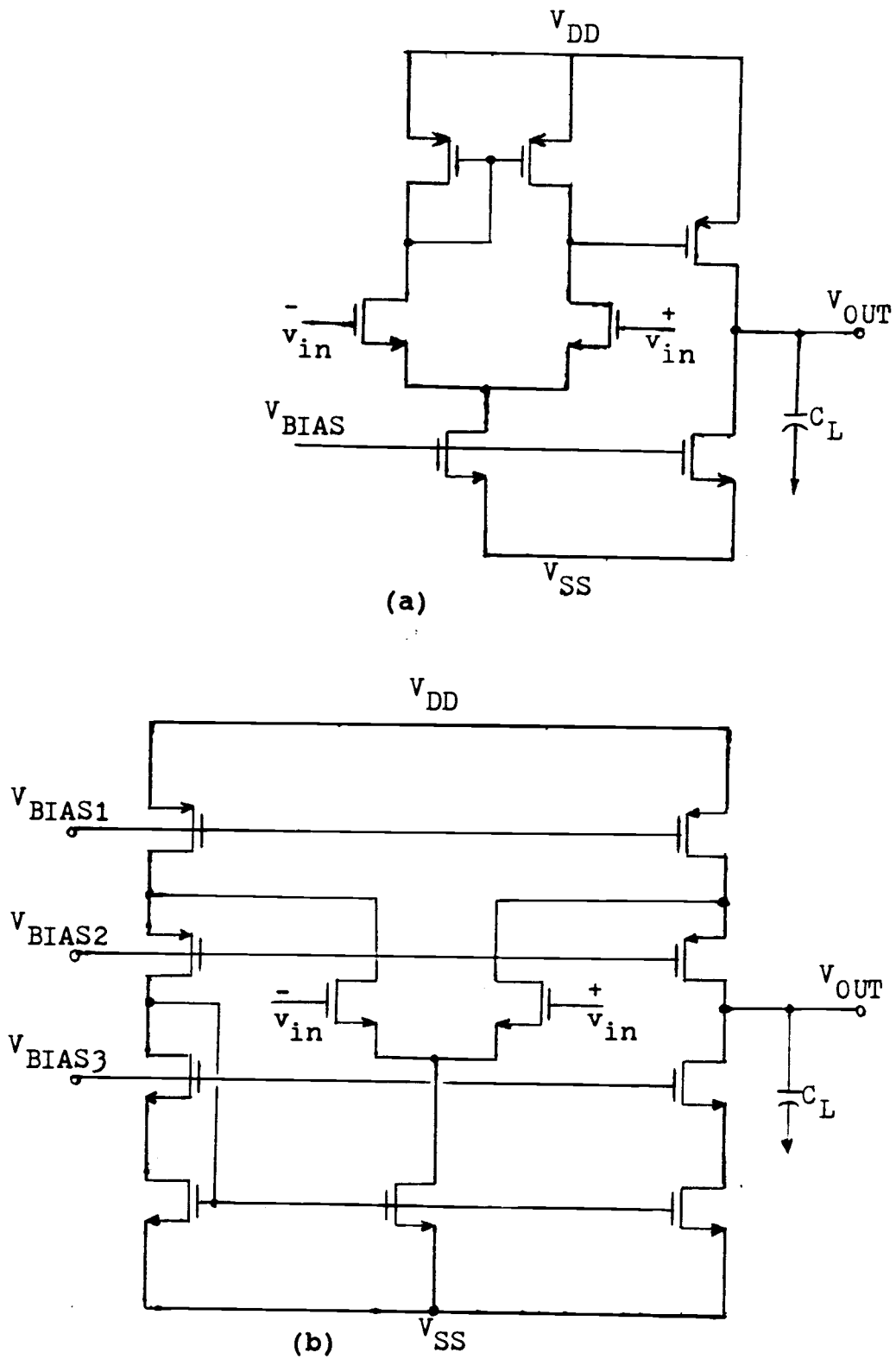


Fig. 5. CMOS op amp topologies.
 (a) Two-Stage op amp;
 (b) Folded-Cascode op amp.

The solution to this dilemma is presented in Fig. 6, where the traditional folded-cascode is modified to introduce a second stage which also serves as a current mirror to reflect the input stage current into the output stages. In this cascode op amp, the structure formed by transistors M2, M3, M4 and M9 allows the principle of input and output chopping of the first stage to be implemented.

For low noise applications it is not enough to use a chopper amplifier, as is evident from the expression for the effective output noise of the chopper amplifier. The input-referred noise is dependent on the following factors:

- (1) S_{Nin} , the original white noise component of the amplifier;
- (2) The ratio of the corner frequency to the chopper frequency.

Therefore, the design of the op amp is critical and the arbitrary use of the chopper is not the solution, in SC applications where a very low value for the input-referred noise is required even at the low frequency end of the noise spectrum.

In the case of the modified cascode op amp,

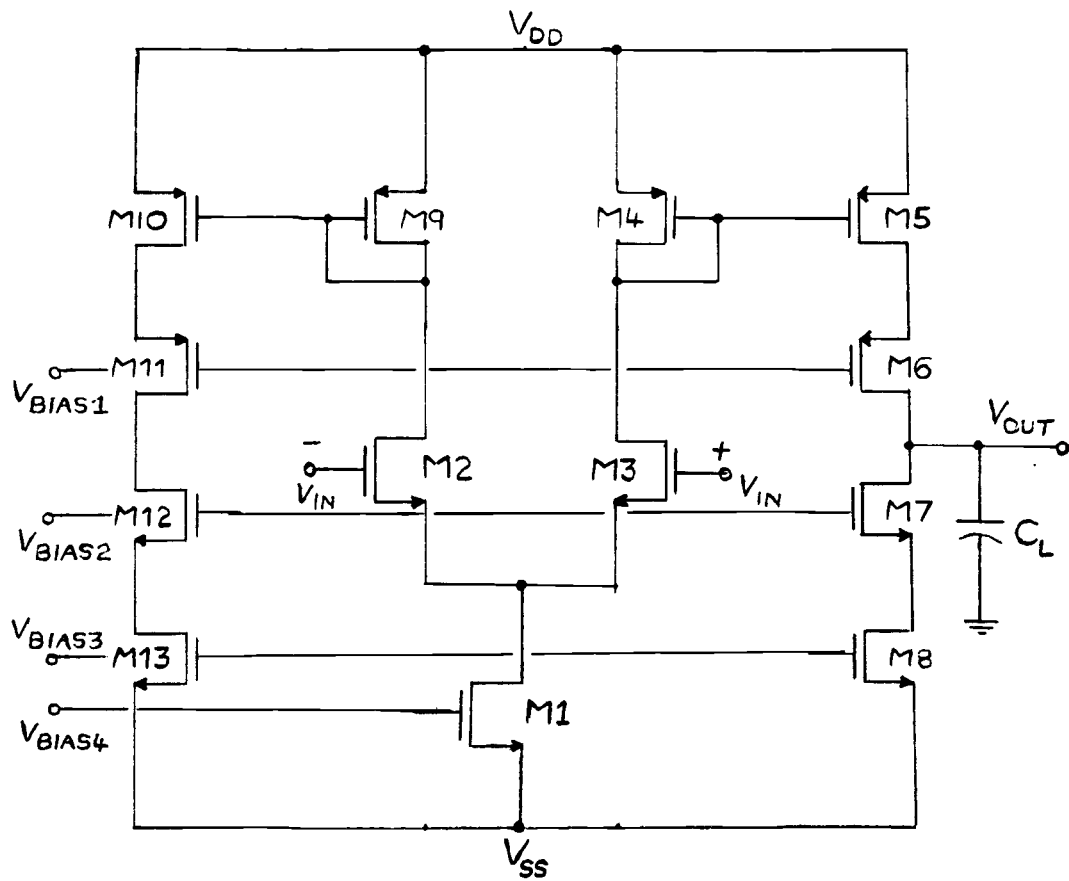


Fig. 6. Modified cascode op amp used to implement the chopper concept.

the main contributors to the thermal noise are the input pair, the current mirror transistors (M4, M9) and the terminal devices in the cascode structure. The thermal noise generated by a MOSFET is given by

$$V_{N \text{ thermal}}^2 = \frac{8kT}{3g_m} \Delta f \quad \text{.....(11)}$$

The g_m term or the transconductance of the MOSFET is dependent on the device size (specifically the W/L ratio). Therefore, one way to decrease the thermal noise value is to increase the device g_m , i.e., to increase the W/L ratio. Having the dominant white noise contributing devices of significant dimensions is the only way to design an op amp with an inherently small value of S_{Nin} . Though this represents an increase in the die area, it is a trade-off between area and performance.

To avoid aliasing, in SC filters the sampling frequency must be at least twice the maximum signal frequency. For telephony systems, f_{signal} is limited to 3.2kHz and $f_{\text{samp}}/2$ or $f_{\text{samp}}/4$.

If the SC circuit is to be designed for an application where parameters such as the maximum signal frequency, sampling frequency and chopping frequency are pre-specified then the op amp for this

SC circuit has to be designed so as to minimize S_{Nin} . The output noise of the chopper decreases as the ratio of f_k to f_{chop} decreases. However, the higher the chopper frequency, the larger the residual offset of the chopper [4], implying a trade-off between offset and noise. An optimum value for f_{chop} would be:

$$f_k = f_{chop} + \text{maximum signal frequency} \dots\dots\dots(12)$$

Having chosen the chopper frequency and thereby the corner frequency, the design of the op amp has to be manipulated so as to achieve

$$v_{Nthermal}^2(f_k) = v_{Nflicker}^2(f_k) \dots\dots\dots(13)$$

Since both the flicker noise and thermal noise depend on the device size, to design a low noise op amp, die area has to be sacrificed. Reference [6] provides additional information on low noise op amp design.

SELECTION OF SWITCHED-CAPACITOR FILTER STAGES TO BE CHOPPED

SC networks could be one stage or multiple stage filters. Not all stages of a SC filter need be chopped to reduce the input-referred noise. This is because in some networks only one stage might be the dominant noise contributor. Also the chopper only removes the $1/f$ noise component of the op amp; it does not eliminate the capacitor switching noise or the noise produced by the coupling capacitors of the filter. If these terms are significant noise contributors, then chopping does not reduce the noise very much. In this section some SC networks will be examined and the relative merits/demerits of chopping these networks will be discussed.

(a) Circuit #1, High Pass Response

The SC network is shown in Fig. 7 and its noise equivalent (with capacitor C_2 replaced with noise resistor R_2) is shown in Fig. 8. It is evident from the unchopped and chopped input-referred noise plots (Fig. 9) of this circuit that the $1/f$ noise component of the op amp is the main noise source. Since the flatband noise does not increase

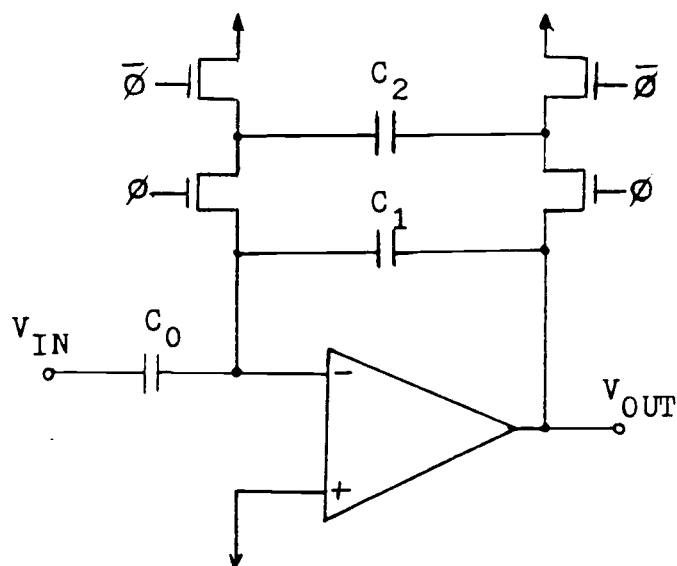


Fig. 7. Circuit #1.

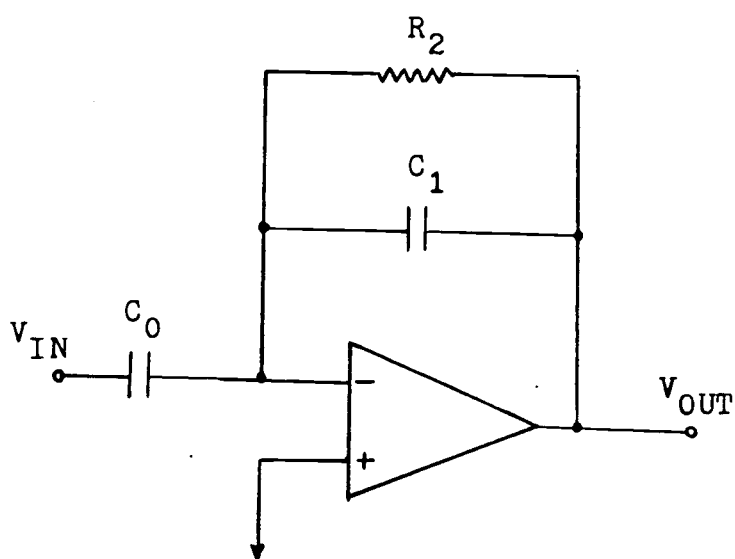


Fig. 8. Equivalent noise network of circuit #1.

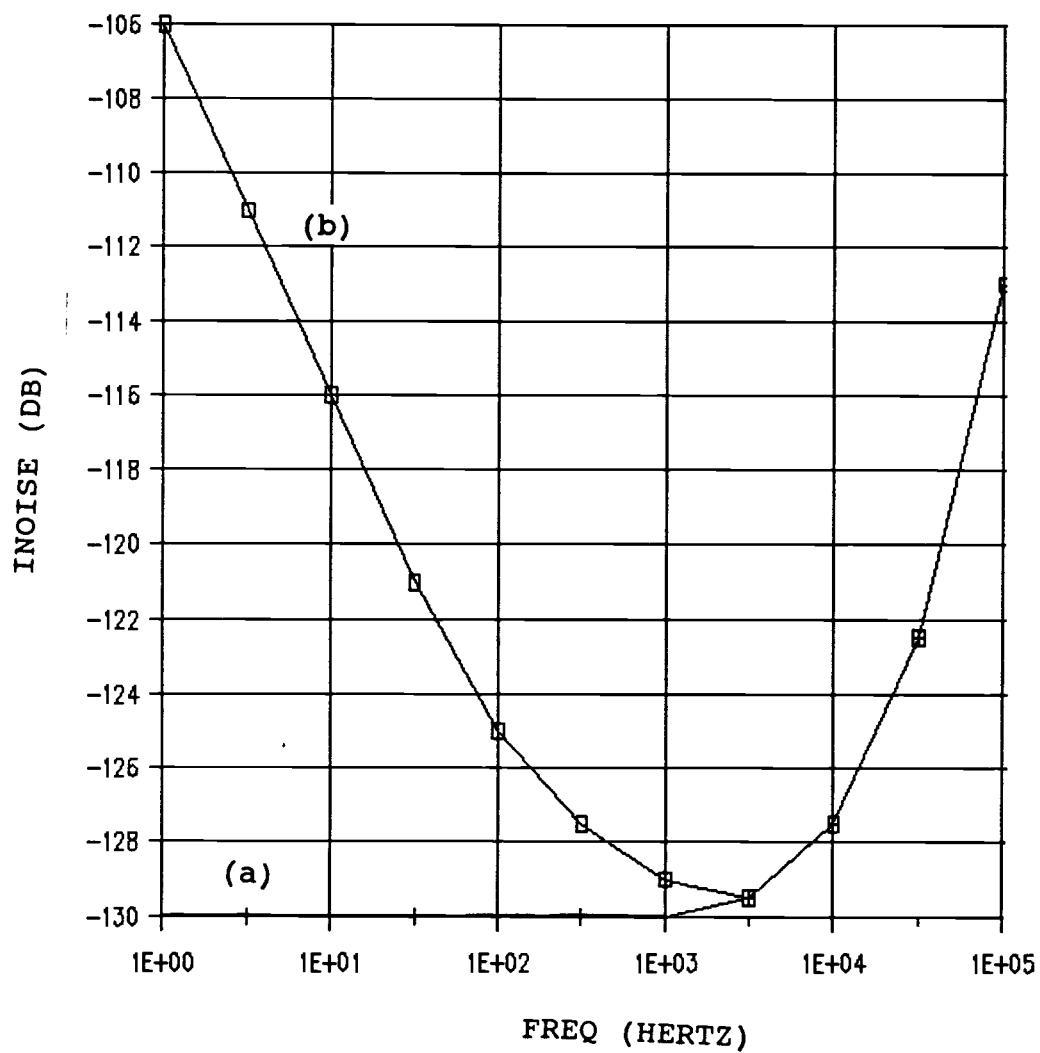


Fig. 9. Simulated input-referred noise plot of circuit 1. (a) Unchopped case; (b) Chopped case.

due to chopping, the second largest contributor of noise is the capacitor switching noise source. However, there is a significant improvement in the noise plot due to the chopper.

(b) Circuit #2, Lossy Integrator

In this circuit shown in Fig. 10, the parasitic-insensitive switching scheme is implemented [7]. Since it is an integrator, it has a low pass response and the input-referred noise plots are shown in Fig. 12. Once again the $1/f$ noise source dominates and there is a considerable reduction in the noise due to chopping. The attractive feature of this circuit is that it has a uniform input-referred noise value from 1 Hz to 6 kHz for the chopped case. This brings to light the fact that for a SC filter to have a flat input-referred noise curve, it has to have a low pass type magnitude plot.

(c) Circuit #3, Low pass Biquad

Shown in Fig. 13, this circuit is a Low pass Biquad, with the gain of the second stage greater than the gain of the first stage. The noise plots (Figs. 14 and 15) indicate the primary noise source as the $1/f$ noise component of the first stage. The

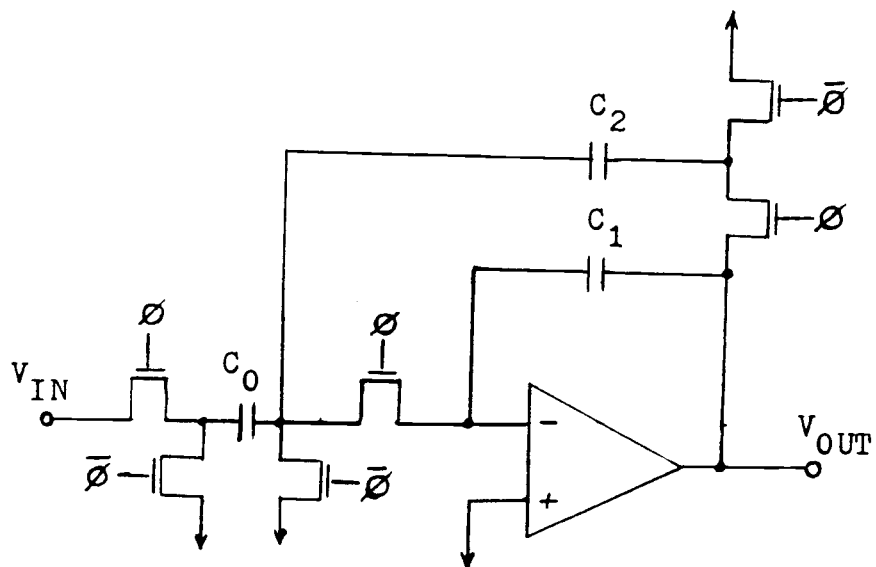


Fig. 10. Circuit #2, Lossy Integrator.

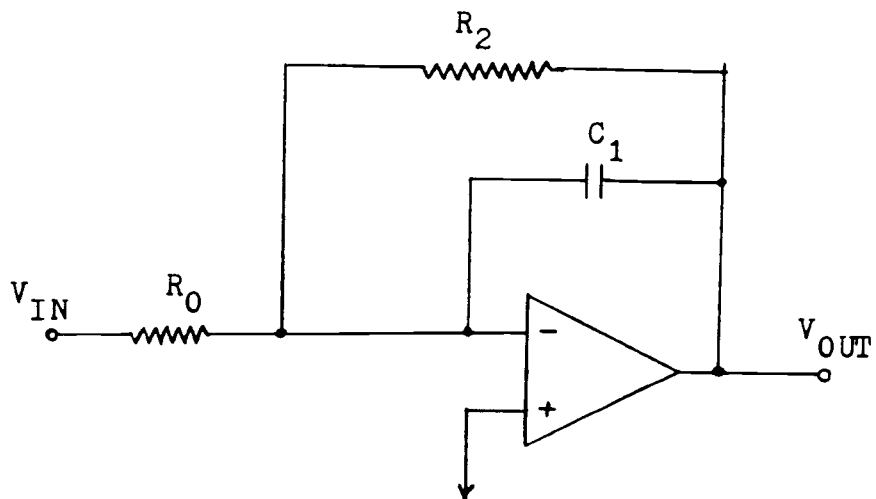


Fig. 11. Equivalent noise network of circuit #2.

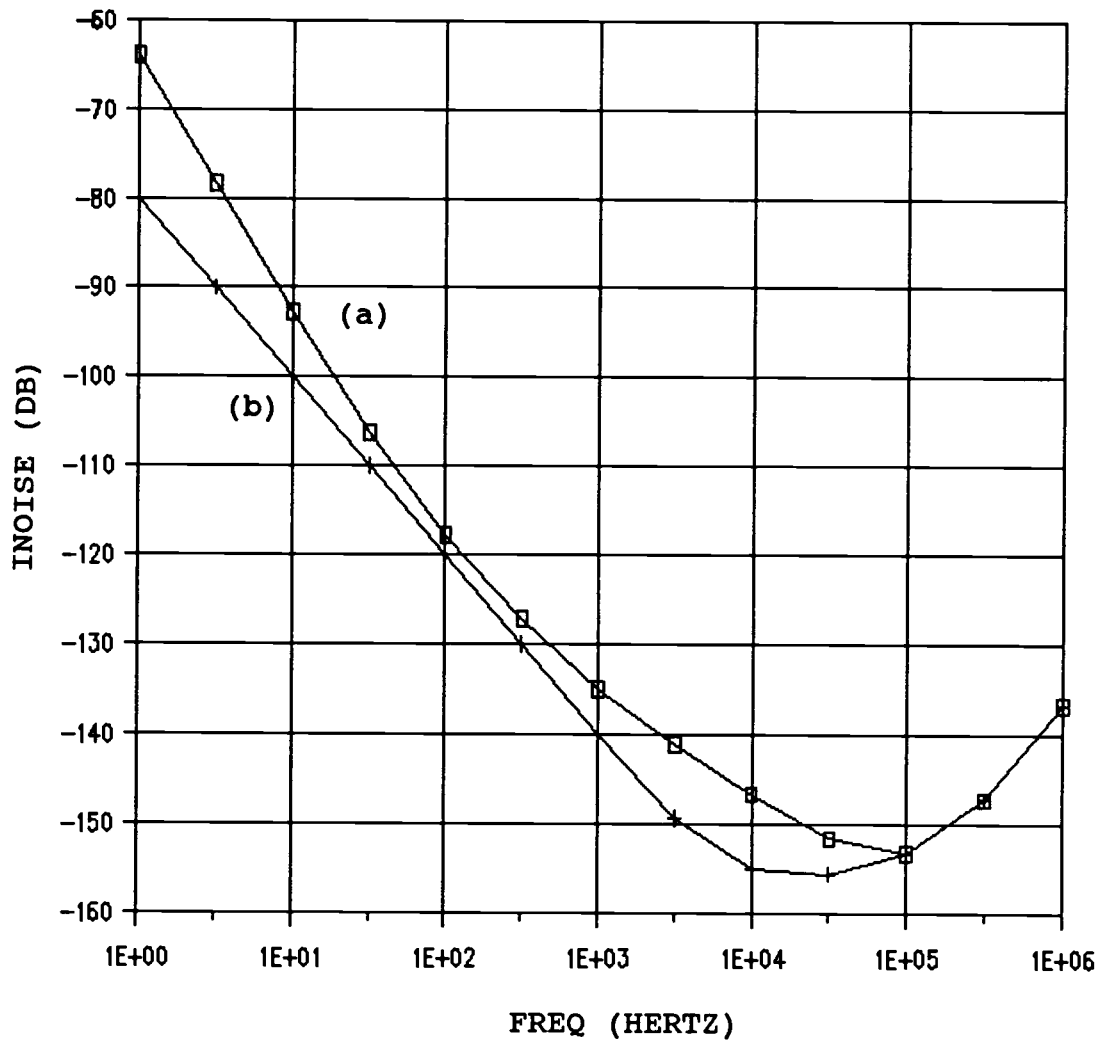


Fig. 12. Simulated input-referred noise plot of circuit #2. (a) Unchopped case; (b) Chopped case.

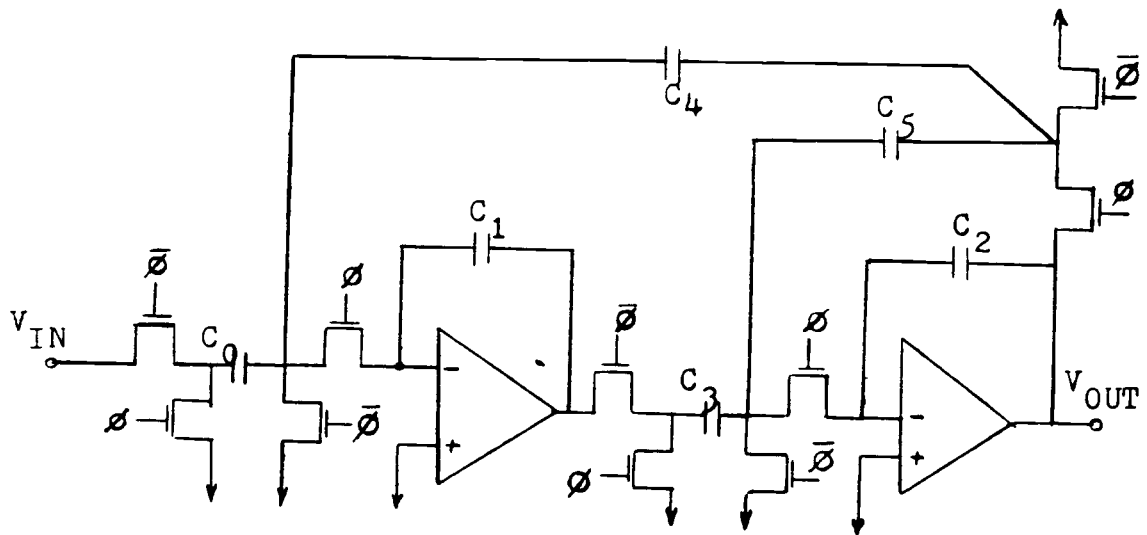


Fig. 13. Circuit #3, Low pass Biquad.

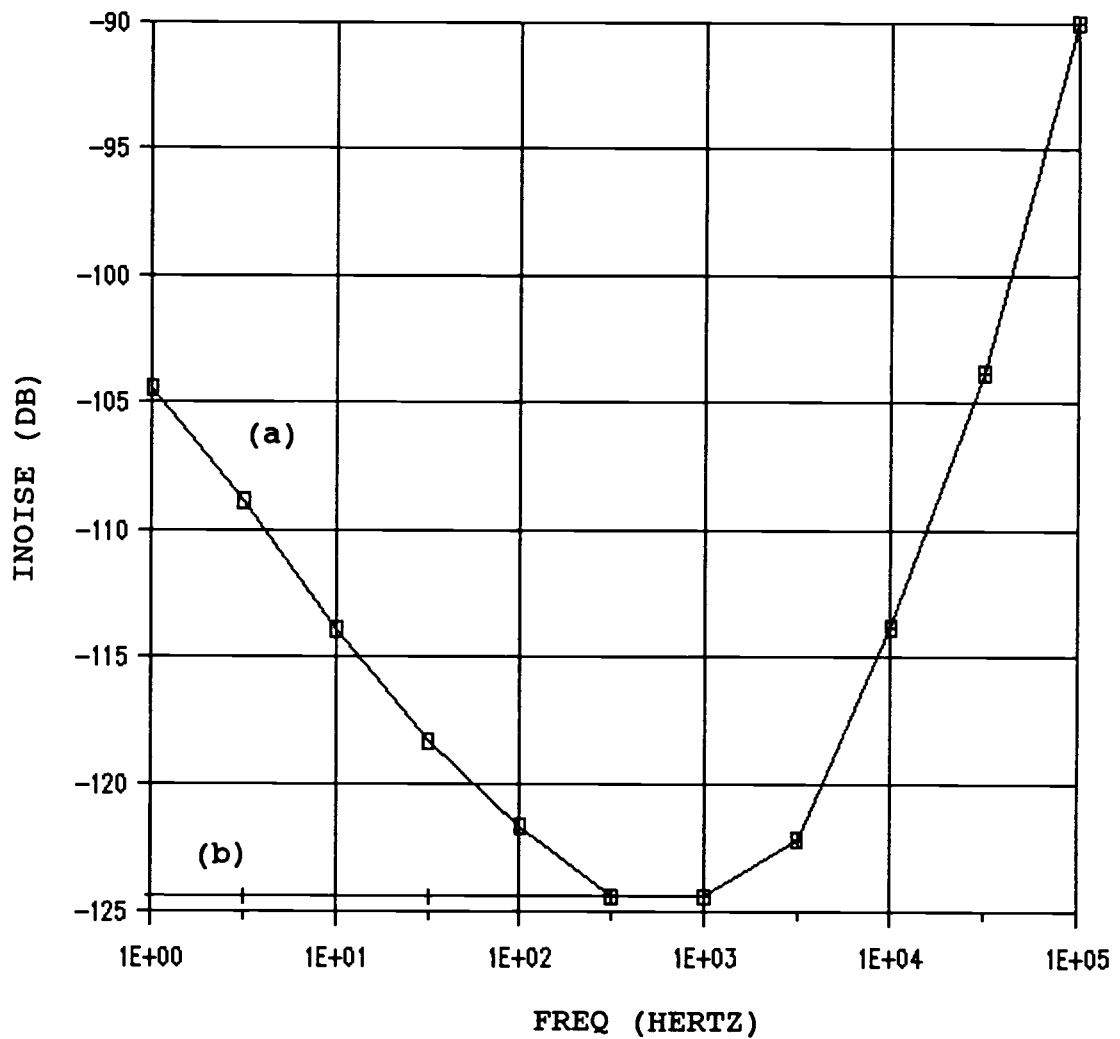


Fig. 14. Simulated input-referred noise plot of circuit #3. (a) Both stages unchopped; (b) First stage chopped, second stage unchopped.

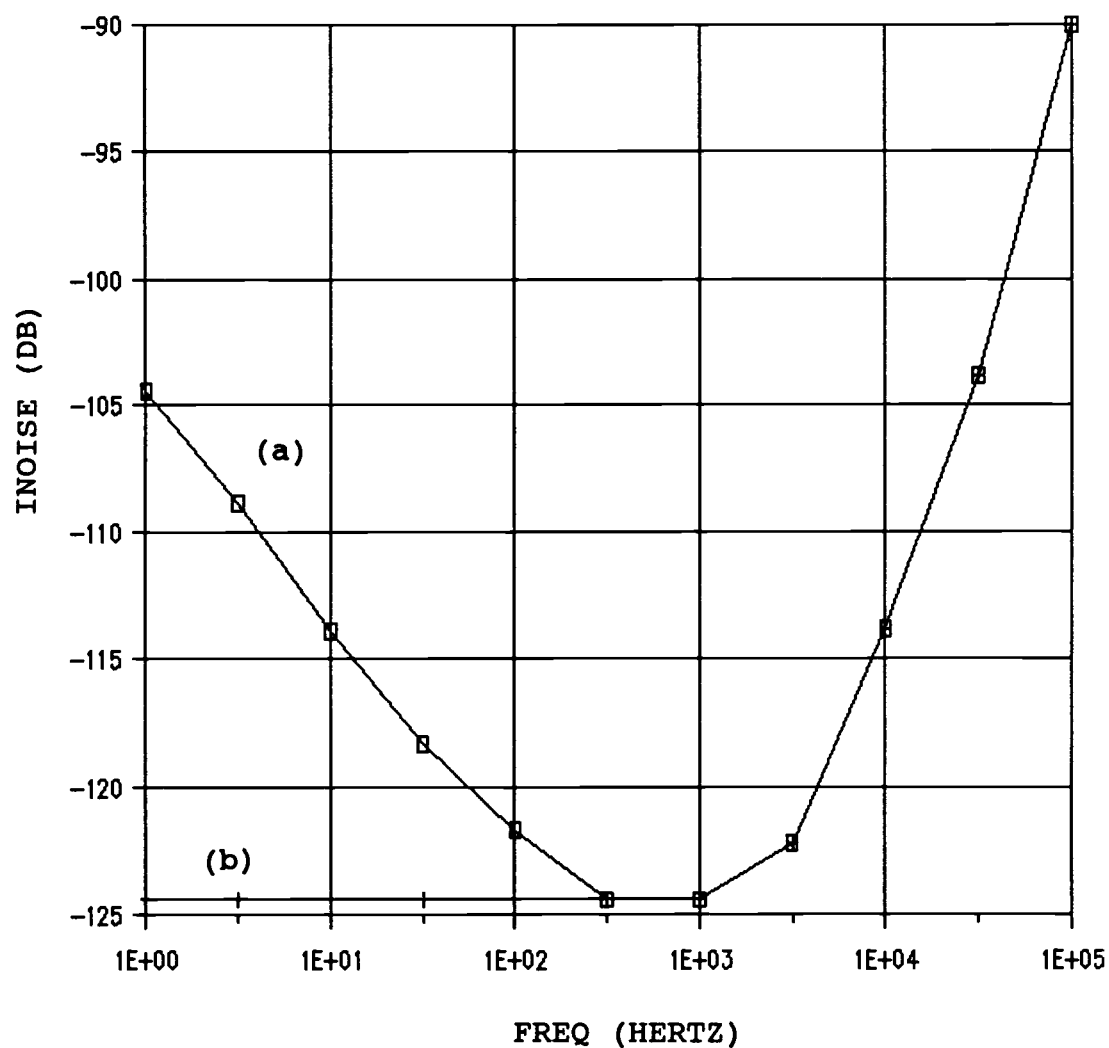


Fig. 15. Simulated input-referred noise plot of circuit #3. (a) First stage unchopped, second stage chopped; (b) Both stages chopped.

capacitor switching noise of the circuit is greater than the noise contribution of the second stage; consequently there is no advantage in chopping the second stage. It is enough to chop the first stage to achieve a significant reduction in the input-referred noise.

The circuit parameters for the SC filters analyzed in this section were:

- (1) Sampling frequency $f_{\text{samp}} = 120 \text{ kHz}$
- (2) Chopper frequency $f_{\text{chop}} = 30 \text{ kHz}$
- (3) Op amp dc gain $A_0 = 102 \text{ dB}$
- (4) Op amp topology - modified cascode
- (5) Op amp unity bandwidth $w_0 = 6.28\text{E}5 \text{ rad/sec}$
- (6) Op amp corner frequency $f_k = 40 \text{ kHz}$.

RESULTS

In this section, the validity of the developed noise model is tested for a specially designed second-order SC filter shown in Fig. 16. The output of the filter is the continuous output terminal of the first (high pass) stage.

The circuit parameters for this filter are:

- (1) Input signal limited to 3kHz
- (2) Sampling frequency $f_{\text{samp}} = 60 \text{ kHz}$
- (3) Chopper frequency $f_{\text{chop}} = 30 \text{ kHz}$

The op amp used in this filter is a conventional Two-Stage with the following specifications:

- (1) Unity gain bandwidth $W = 7.0\text{E}6 \text{ rad/sec}$
- (2) Dc gain $A_0 = 750$
- (3) Corner frequency $f_k = 550 \text{ kHz}$

The magnitude and noise plots of this op amp are shown in Figs. 17 and 18 respectively.

The computed values of the gain factors for both the unchopped and chopped noise models of this filter are shown in Table II. The reason for large values of the gain factors of the chopped SC noise model is that the Two-Stage op amp is not optimized

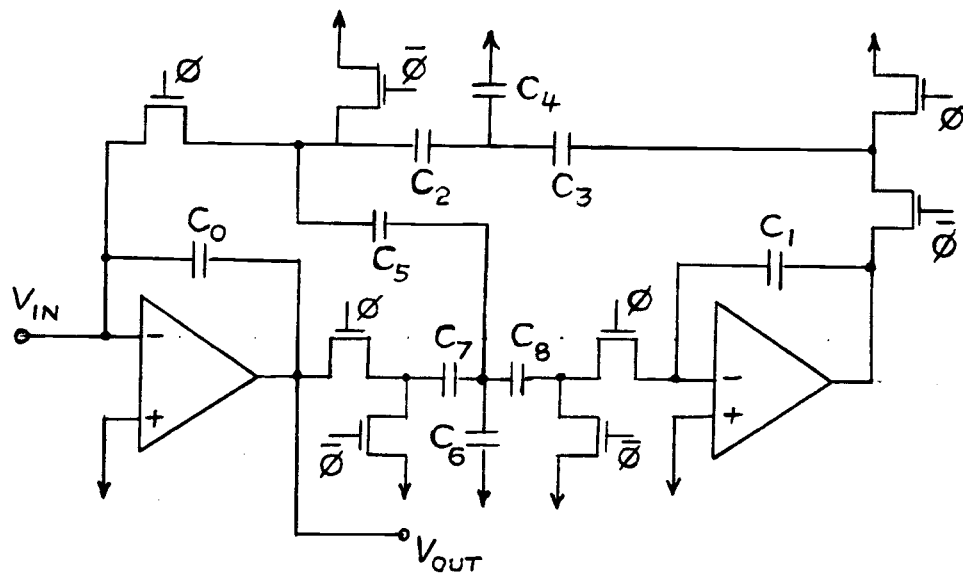


Fig. 16. Test circuit to verify the validity of the noise models.

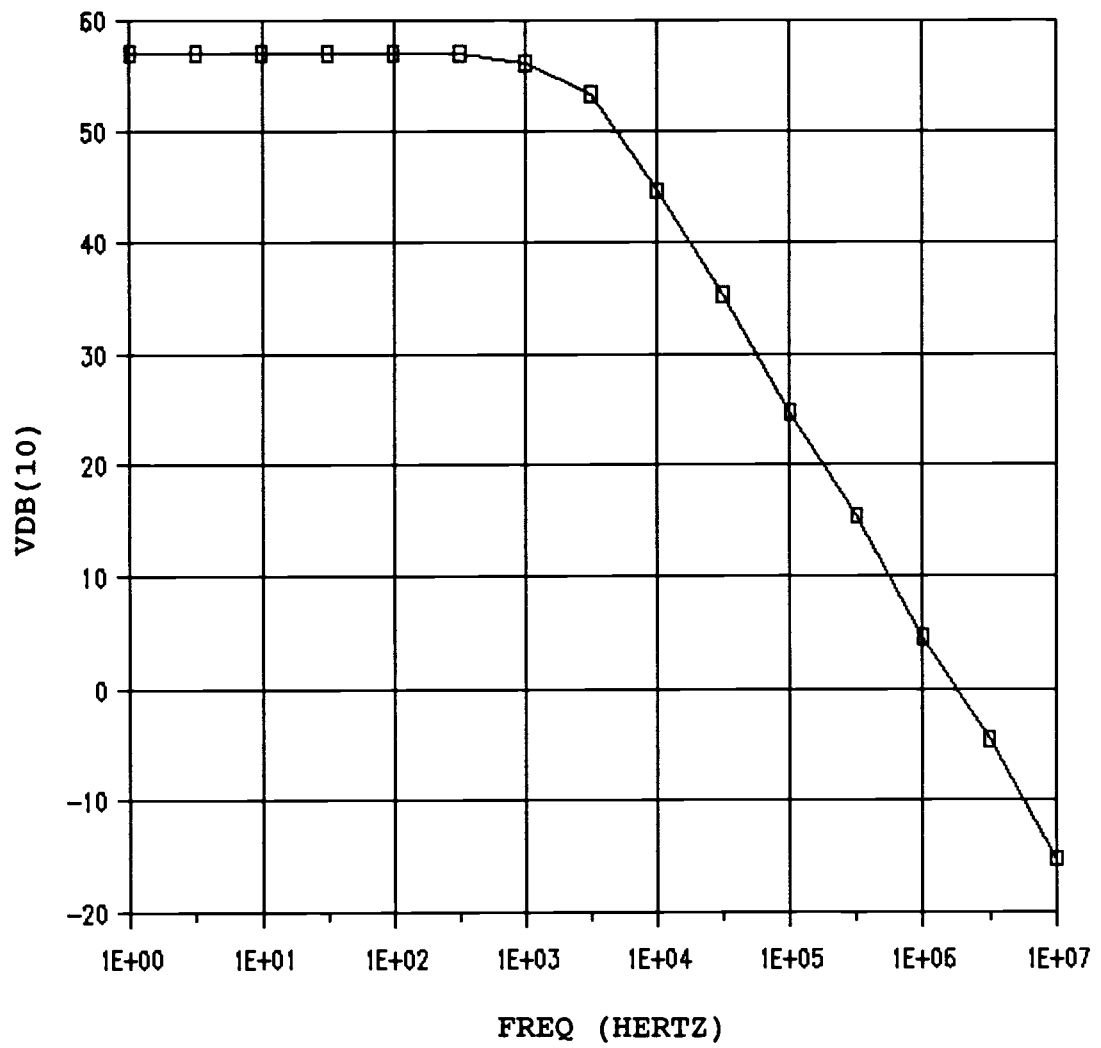


Fig. 17. Simulated magnitude plot of the Two-Stage op amp.

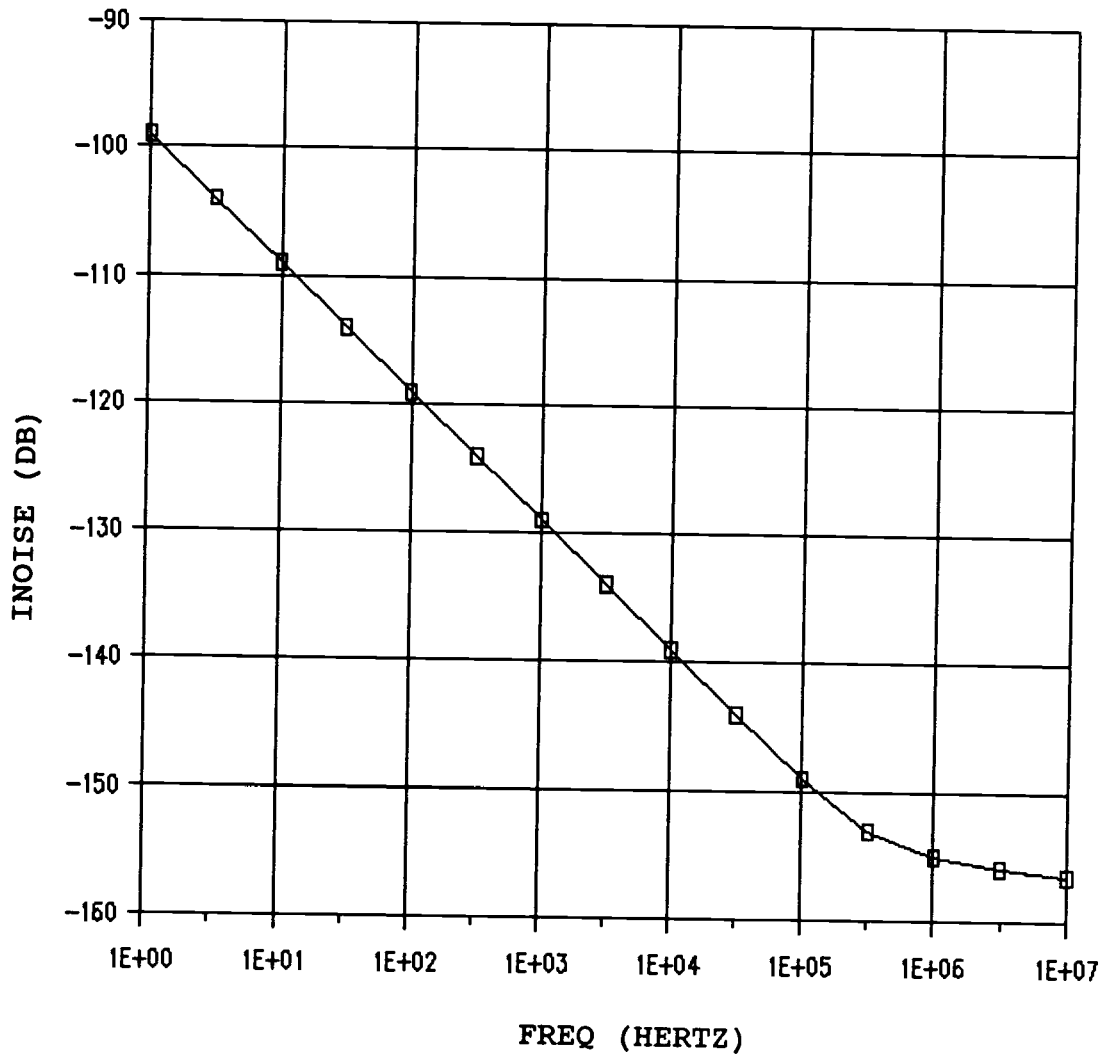


Fig. 18. Simulated input-referred noise plot of the Two-Stage op amp.

TABLE II
GAIN FACTORS FOR THE TWO STAGE OP AMP

GAIN FACTORS	UNCHOPPED SC NOISE MODEL	CHOPPED SC NOISE MODEL
G	1	4
G'	1	0
k	9.6	9.6

to match the circuit parameters i.e. $f_k \neq f_{\text{chop}} + f_{\text{signal}}$.

The measurements were performed using the HP32582A RF spectrum analyzer. The resolution of this analyzer is 0.5% of the frequency span which can be set from 1Hz to 25 KHz in a 1-2.5-5-10 sequence. Calibration of the instrument was performed using standard resistor values and incorporating a correction factor in all the measurements.

The frequency range of interest is the 1Hz to 1KHz spectrum. However, the 1Hz to 90Hz span is not accurate for noise measurements as it is dominated by the sidebands of the 60Hz power supply noise. Figs. 19 and 20 show the output noise plots of the test circuit for two cases--both stages of the filter chopped and both stages chopped respectively.

For both cases 17 sample measurements were taken at 100, 200, 500 and 1000Hz. Using the sample points, standard deviation from the median was computed and error bars were placed on the output noise plots at these frequencies. Deviation from predicted (simulated) values is less than 3dB for both cases, which proves the validity of the model.

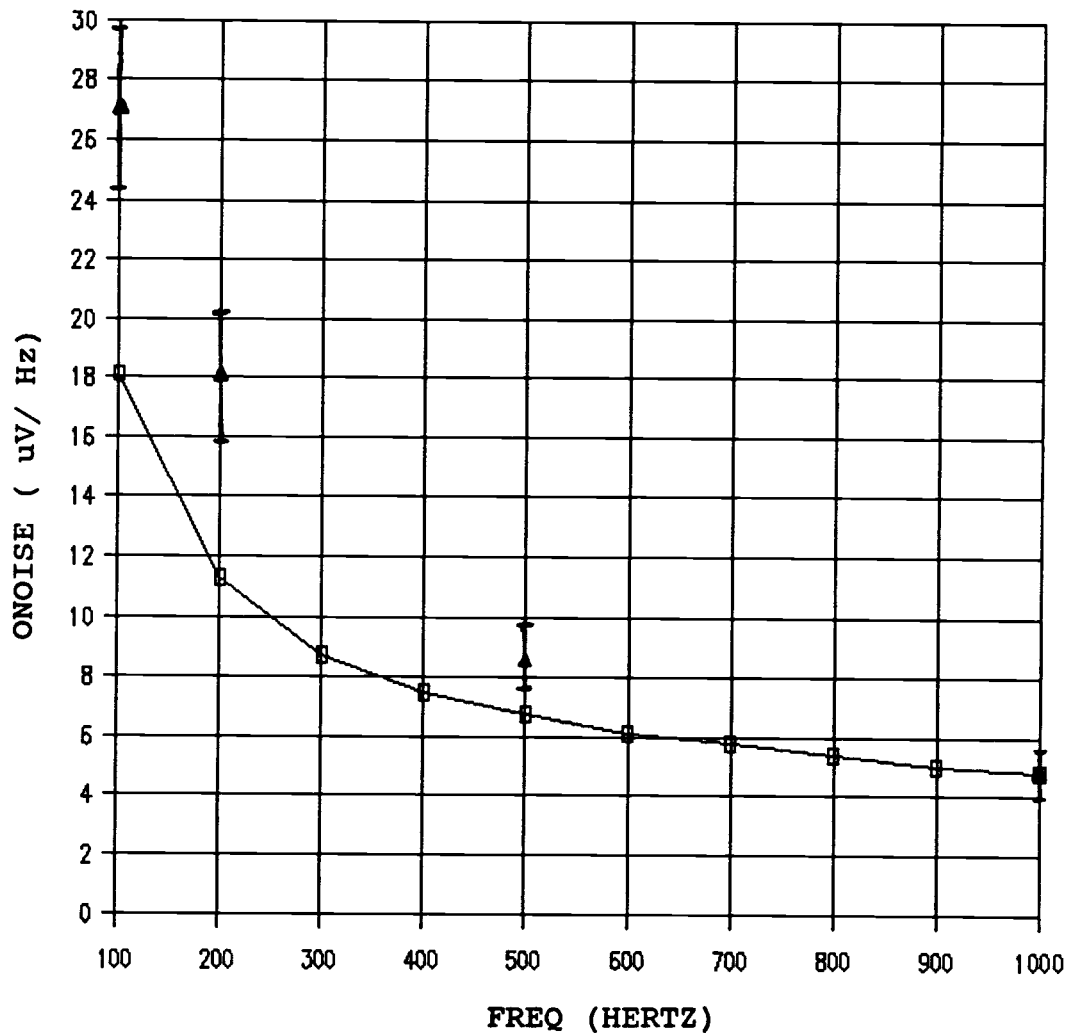


Fig. 19. Output noise plot of the test circuit, both stages unchopped. ---- simulated curve, median of measured values and the corresponding error bar.

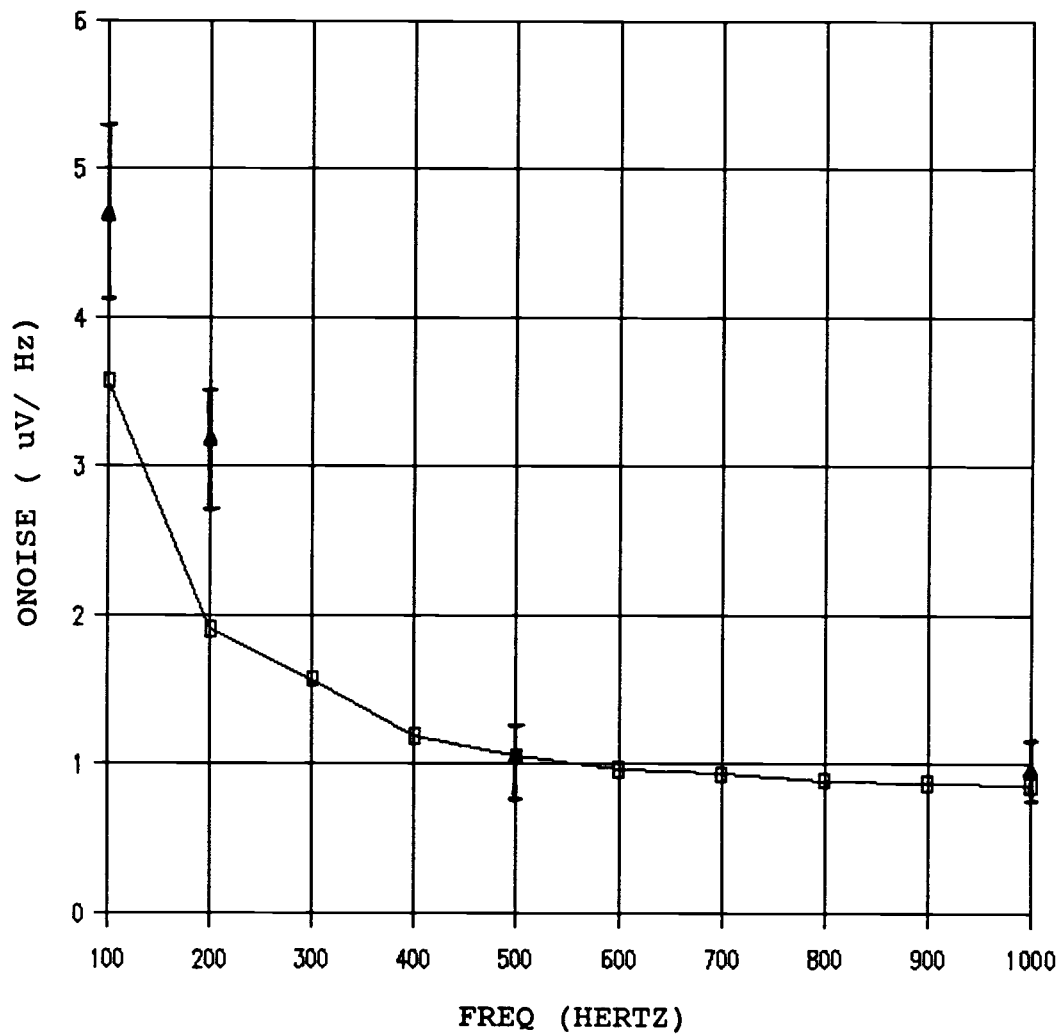


Fig. 20. Output noise plot of the test circuit, both stages chopped. — simulated curve, ▲ median of measured values and the corresponding error bar.

The accuracy of the measurements could have been significantly improved had an audio spectrum analyzer (normal frequency span 1Hz - 5KHz) and a noise meter been used. But both these instruments were not available at hand. The reason that an audio spectrum analyzer performs better than an RF spectrum analyzer in this instance is because the former has a better resolution for the range of frequencies of interest. The noise meter would have improved the calibration and could have also served as a second source for measurements.

Fig. 21 shows the simulated output noise plot of the same SC filter using a modified cascode op amp designed using the principles discussed in the previous sections. It is evident that this op amp performs much better than the Two-Stage op amp of Fig. 17 since it has been designed to match the circuit parameters. The magnitude and noise plots of the modified cascode op amp are shown in Figs. 22 and 23 respectively. This op amp has a die area comparable to that of the Two-Stage and has the following specifications:

- (1) Dc gain $A_0 = 102 \text{ dB} = 1.26\text{E}5$
- (2) Unity gain bandwidth $W_0 = 6.28\text{E}6 \text{ rad/sec}$

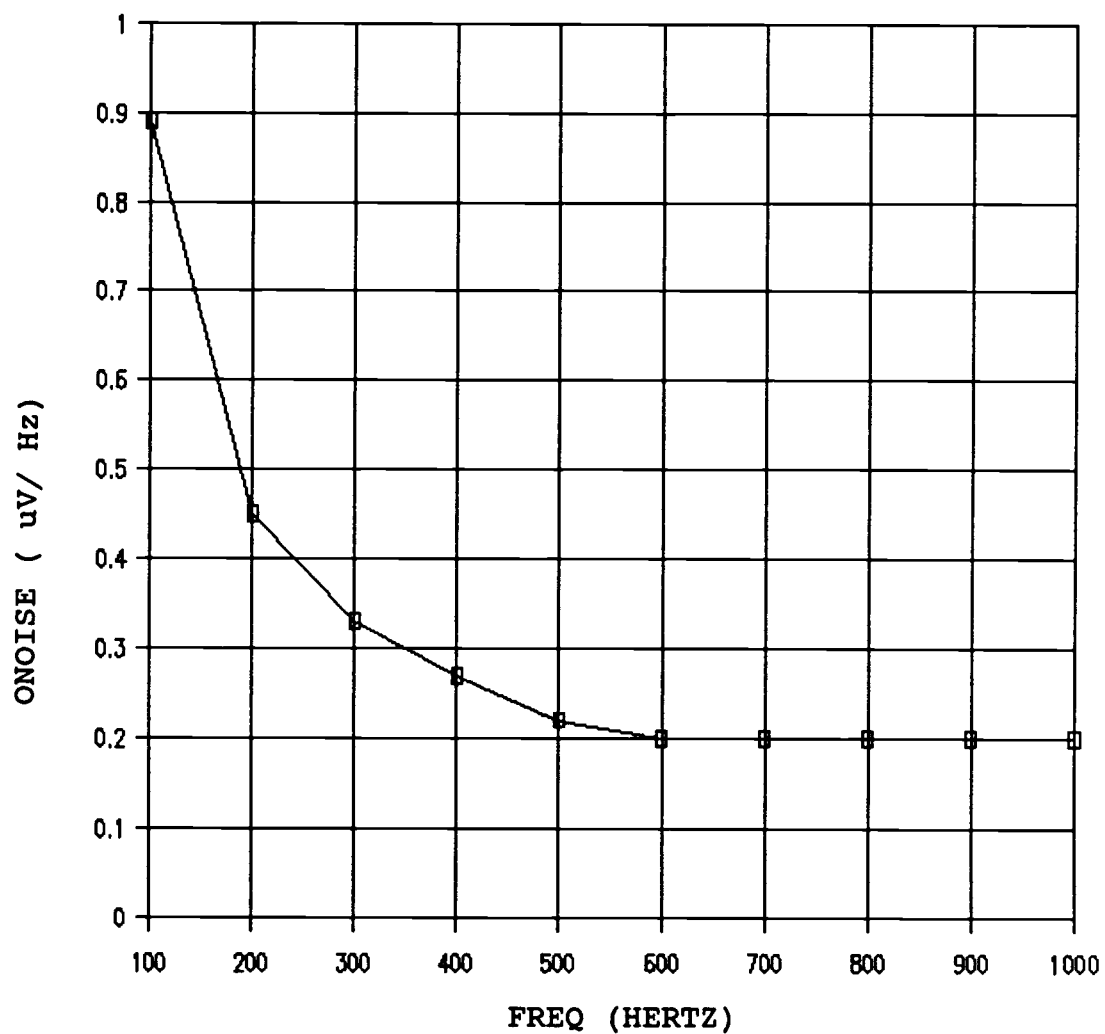


Fig. 21. Simulated output noise plot of the test circuit, both stages chopped, op amp used---- modified cascode.

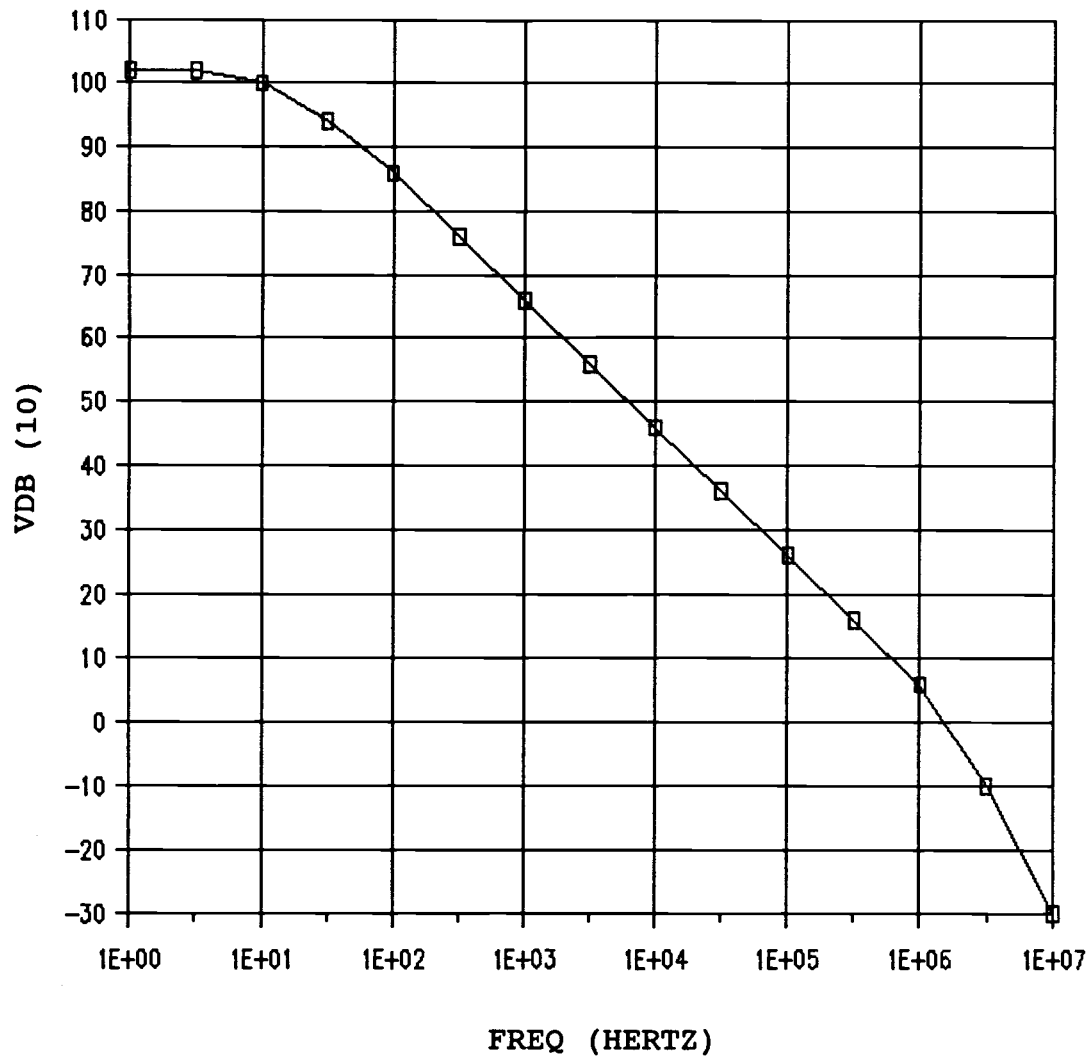


Fig. 22. Simulated magnitude plot of the modified cascode op amp designed for the test circuit.

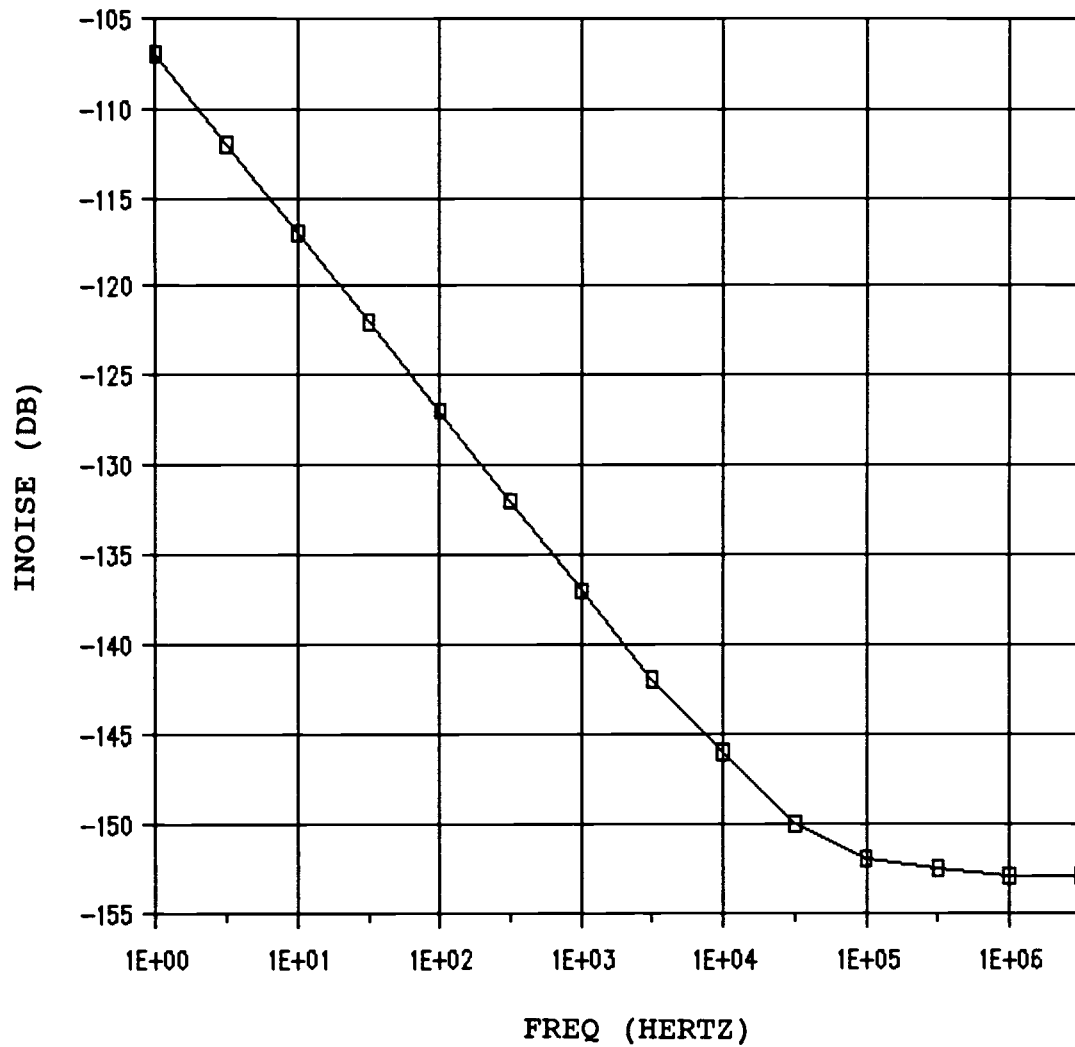


Fig. 23. Simulated input-referred noise plot of the modified cascode op amp designed for the test circuit.

- (3) Corner frequency $f_k = 40 \text{ kHz}$
- (4) Input referred noise at 100kHz = -154 dB.

CONCLUSIONS

A highly flexible noise model based on SPICE2 for SC filters has been described. The model allows analysis of both chopped and unchopped SC networks which is especially useful in noise prediction of multiple order SC filters in which only some stages might be chopped. Measured and simulated noise values differ by only 3 dB which verifies the accuracy of the model.

For SC applications which require a low value of the input-referred noise even at the low frequency end of the noise spectrum, it is not enough to implement the chopper-stabilization technique. An op amp with an inherently small thermal noise component has to be used. A modified cascode op amp topology in CMOS has been discussed, that is suitable for use with a chopper.

A method for selection of SC network stages for optimal use of the chopper has been outlined and is used to determine which stages of a SC network need be chopped to reduce the input referred noise.

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APPENDIX

APPENDIX

To avoid aliasing in SC filters, the signal frequency is limited to less than half the sampling frequency. However, the output noise of the op amp is aliased since it has a bandwidth much greater than the sampling rate. If we define BW_n as the equivalent noise bandwidth of the op amp where BW_n is that bandwidth required to contain the same noise power as the op amp but with a uniform spectral density.

$$BW_n = \frac{\pi}{2} BW_u \quad \dots\dots\dots(14)$$

where BW_u is the unity gain bandwidth of the op amp. If the input signal is the thermal noise of an op amp, the sidebands produced due to aliasing will be uncorrelated and the increase in the thermal noise spectral density is given by:

$$K = \frac{(2 BW_n - 1)}{f_{\text{samp}}} \quad \dots\dots\dots(15)$$

Since SPICE works with noise voltages, the noise voltage gain due to sampling or the folded flat-band factor is obtained by taking the square root of K.

Sampling of the $1/f$ noise component also produces uncorrelated sidebands which add to the baseband $1/f$ noise density. However, it has been found [2] that for sample rates of 60kHz or higher, the contribution of the sidebands is less than 20% of the baseband noise density. For sample rates of 100kHz or higher, the foldover effect of the $1/f$ noise component can be neglected. In the developed noise model, we have chosen to ignore this contribution.