

A CMOS Channel-Select Filter for a Direct-Conversion Wireless Receiver

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Abstract— A highly selective and linear switched-capacitor channel-select filter is fabricated in 1- μ m CMOS for a direct-conversion wireless receiver operating in the 902–928 MHz ISM band. The filter selects a 230-kHz wide channel and attenuates by at least 50 dB from 320 kHz to 57 MHz. The input IP3 is +30 dBm, the input-referred noise in the passband is 70 nV/ $\sqrt{\text{Hz}}$, and the circuit takes 4.6 mA from a 3.3 V supply. Direct subsampling of the 915 MHz RF input signal by the filter front-end is also demonstrated with only a small degradation in linearity. The input noise voltage is halved in a redesign while keeping the current drain unchanged.

Index Terms— Biquadratic filters, CMOS analog integrated circuits, elliptic filters, frequency shift keying, low-pass filters, surface acoustic wave filters, switched capacitor circuits, switched capacitor filters, UHF radio communication, UHF receiver.

I. INTRODUCTION

RECENT research shows the feasibility of CMOS RF circuit blocks in the 1 GHz frequency range [1]–[8]. These elementary RF building blocks have recently been assembled into subcircuits such as a receiver front-end and a spread-spectrum transmitter [9]. On-chip filtering remains key to a totally integrated wireless receiver, and this is one of the first reports on a CMOS filter for a wireless receiver.

The channel-select filter must satisfy some exacting requirements. The first and most important is *selectivity*. A filter with a sharp cutoff should select the desired channel and attenuate all other adjacent users. An active implementation of the filter must itself be *low noise* so as not to degrade the overall receiver noise figure, and it must have *wide dynamic range* so that out-of-band interferers, prior to filtration, do not create intermodulation distortion which falls in the passband. Passive filters admirably meet these requirements and are therefore widely used in wireless receivers today. However, they take up precious volume on the circuit board. The receiver IC must also drive small signals at a low characteristic impedance into the filter, often at very high frequencies. The off-chip buffers required to do so dissipate additional power, and the off-chip interconnections may also pickup noise.

For at least three reasons, active filters have made little impact in superheterodyne receivers, where a *bandpass* filter with high- Q poles selects the desired channel at some intermediate frequency (IF). First, the power dissipation of an active

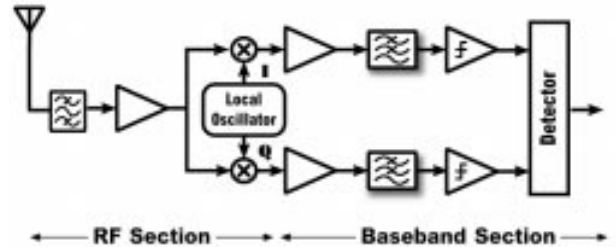


Fig. 1. Direct-conversion frequency shift keying (FSK) receiver. Low-pass channel-select filter is highlighted.

bandpass filter rises proportionally to the center frequency and the required dynamic range [10]. Second, it is increasingly difficult because of parasitic effects to accurately realize high- Q poles at high center frequencies. Third, it is difficult to meet the specifications on distortion because the filter must handle signals and interferers which are *amplified* by the front-end. For these reasons, ceramic and surface acoustic wave (SAW) filters tuned to frequencies from 100's of kHz to 100's of MHz are used in vast numbers in production broadcast radio receivers and conventional cellular telephone handsets. In the case of the popular 10.7 MHz IF used in FM receivers, it is difficult to meet receiver specifications with even the best reported continuous-time filters [11], although bandpass delta-sigma modulators may prove a viable alternative [12].

However, it has always been one of the attractions of a direct-conversion, or zero-IF, receiver that a low-pass filter with poles of a modest Q selects the desired channel, and that this filter may be integrated with the rest of the receiver [13] (Fig. 1). Some active channel-select filters have been developed for this purpose [14]–[17]. These bipolar continuous-time filters usually trade off large-signal handling against noise. Recently, attention has turned to CMOS switched-capacitor (SC) filters [18]–[21], which are better suited to handle large signals. It is now possible with the appropriate filter architecture and good circuit design to preserve the dynamic range of the receiver frontend. A CMOS continuous-time baseband filter has also been published, with a noise and dynamic range comparable to the best SC filter results [22]. This shows that aside from the additional overhead of self-tuning, the performance of a continuous-time filter is on a par with an SC implementation.

II. FILTER SPECIFICATIONS

The SC channel-select filter described in this paper is designed for use in a frequency-hopped spread-spectrum direct-conversion receiver operating in the 902–928 MHz ISM band

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[23]. Data at 160 kb/s is transmitted by quaternary frequency shift keying (FSK) modulation, where each pair of data bits offsets the carrier frequency by one of the amounts -160 , -80 , $+80$, or $+160$ kHz. The modulated spectrum, therefore, lies within ± 200 kHz of the carrier, which hops pseudorandomly once every four symbols across 50 slots in the 26 MHz-wide ISM band. The RF preselect filter at the front of the receiver passes the entire ISM band, and an agile local oscillator synchronized to the hopping pattern of a particular transmitter (the desired user) downconverts the ISM band to center the sought channel at dc.

The specifications of the required channel-select filter are illustrated in Fig. 2. The filter passband encompasses the spectrum of four FSK tones at ± 80 and ± 160 kHz, modulated in continuous-phase by pseudorandom data. The filter must suppress all the other signals populating the downconverted ISM band, particularly the adjacent user whose carrier frequency is 480 kHz above and below the desired user. The sidelobes of each quaternary tone at a baud rate of 80 kHz occupy a 6-dB signal bandwidth of ± 40 kHz around the tone frequency, which means that the spectrum of the tones at 160 kHz extends to ± 200 kHz. The filter passband is 230 kHz wide. System level simulations show that a passband ripple of 1 dB or so does not adversely affect the detector's ability to discriminate between the four tones. Ultimately, the correlating detector [24] sets the overall receiver noise bandwidth to 160 kHz. The stopband edge is placed at the closest tone generated by the adjacent user, 320 kHz, where the stopband loss is 50 dB. One sidelobe of this tone, however, lies in the filter transition band, and is therefore not attenuated as much. The worst case occurs at the -6 dB down frequency of the sidelobe, where the transition band loss is only 24 dB. There, the filter attenuates the adjacent user's spectrum by only 30 dB, compared to 50 dB at 320 kHz and beyond. The partial passage of the adjacent channel energy does not measurably degrade receiver sensitivity.

III. FILTER ARCHITECTURE

A. Main Channel-Select Filter

The channel-select filter must implement a sharp transition from 230 to 320 kHz. Since the order of the filter determines the number of opamps, a low-order realization is sought to reduce size and power consumption. A *sixth-order elliptic* filter fulfills the magnitude specification, given the sharp transition band. The group delay, however, changes somewhat across the elliptic filter's passband. This happens to be not very important for FSK detection, where the detector is sensitive only to frequency and not to phase. The group delay of the elliptic filter varies by $3 \mu\text{s}$ across the range of the signal frequency spectrum (Fig. 3), which is $\pm 10\%$ of the $12.5 \mu\text{s}$ symbol period. This degrades the received SNR by at most 0.5 dB. In this application, the main impairment arises at a carrier frequency hop, when the sudden change in phase upon a new frequency propagating through the wireless channel from the transmitter to receiver stimulates a time-domain transient in the receiver's channel-select filter.

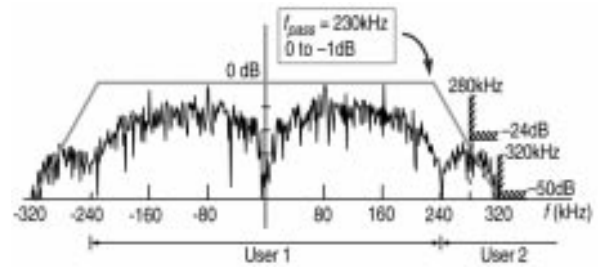


Fig. 2. Channel-select filter specification for reception of quaternary FSK. Frequency spectrum of 4-FSK modulated pseudorandom shown in passband.

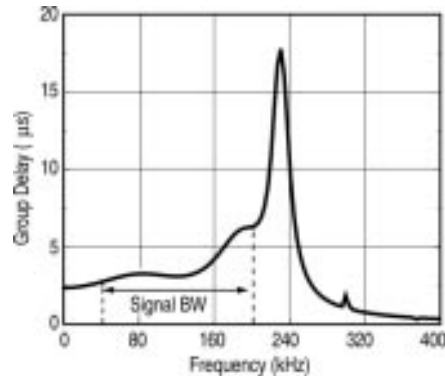


Fig. 3. Group delay variation of elliptic filter.

The sixth-order elliptic channel-select filter is implemented with a cascade of three switched-capacitor biquadratic sections. Although the filter characteristics of a ladder are less sensitive to element-value variations, a cascade of biquads with distributed gain gives the circuit designer more flexibility to optimize dynamic range. The nonlinearity of the filter, which is covered more in detail in a later section, is measured by how well it handles overload caused by large off-channel interferers. By interspersing amplification with filtering, the desired signal progressively grows through the filter while the interferers diminish, so that at the later stages of the filter where the signal swing is largest, the interferers are substantially attenuated.

Nonlinearity defines the upper end of the dynamic range, and filter noise the lower end. The input-referred noise spectrum rises as the frequency approaches the passband edge. This phenomenon is well understood to arise from the noise contributions of the high- Q filter sections. Similarly, at low frequencies the noise spectrum rises because of flicker noise. In case the spectral density of the input-referred noise is unequal at the frequencies of the two baseband FSK tones, then the higher level of the two determines the filter dynamic range.

The three biquad sections are cascaded in an order set by their Q -factors [25]. The low- Q section appears first, to attenuate high frequency out-of-band signals and input noise, and to ease the requirements on any antialiasing filter prior to the SCF. The high- Q section follows, so the gain of the previous stage lessens the noise this section contributes to the filter input, particularly near the passband edge where the input-referred noise spectrum rises. The medium- Q section is placed last.

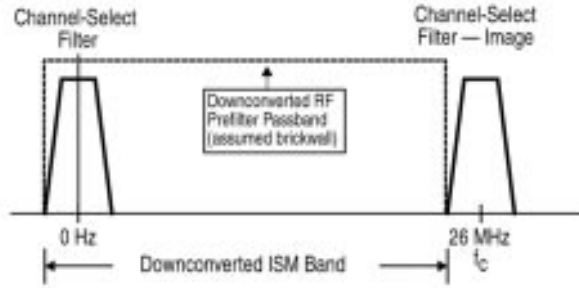


Fig. 4. Minimum clock rate for SC channel-select filter. Assumes brickwall RF preselect filter.

For convenience, the sampling frequency of the elliptic SC filter is derived by dividing down the receiver's fixed LO frequency, f_{LO} , by a power-of-two. It follows from fundamental considerations that for a fixed filter cutoff, the higher the sampling frequency, the larger the spread in filter capacitance [25]. Assume, initially, a perfect RF preselect filter with brickwall band edges that passes only the ISM band. Now, if the desired channel hops to one edge of the ISM band, the channel-select filter must attenuate all ISM users populating the 26 MHz-wide spectrum to the other edge of the band. The RF preselect filter removes all signals outside the ISM band. Therefore, a sample rate of at least 26 MHz forces all the images of the channel-select filter's transition band and passband to lie outside the ISM band, where the RF preselect filter removes all out-of-band signals (Fig. 4).

However, a practical RF preselect filter is not perfect. For instance, the transition band of a sixth-order 902–928 MHz dielectric resonator filter spans 50 MHz, and its stopband loss is only about 40 dB. Therefore, a non-ISM band interferer in the transition band of the RF preselect filter will, after only mild attenuation, fall in the image passband of the channel-select filter. It is impractical to push out the image into the preselect filter's stopband by raising the channel filter's clock rate, because the capacitor spread will then grow excessively. A multirate filter cascade works much better, where a *low-order prefilter* clocked at a high rate suppresses signals lying at the images of the higher-order channel-select filter's passband [25]. The channel-select filter subsamples, or decimates, the prefilter's output. The following section discusses the prefilter design in greater detail.

B. Prefilter

A prefilter is designed whose passband extends beyond that of the main filter, so that it does not contribute a measurable droop or ripple in the main filter passband. However, its stopband edge is placed at a considerably higher frequency, $f_C - f_{STOP}$, where f_C is the main (channel) filter's clock frequency, and f_{STOP} is its stopband frequency. The prefilter is clocked at some integer multiple, N , of f_C , preferably a power-of-two for convenient frequency division. Thus, the first $N - 1$ images of the main filter passband lie in the prefilter's stopband, and when N is sufficiently large, the first passband image of the prefilter itself, at Nf_C , lies well in the stopband of the RF preselect filter.

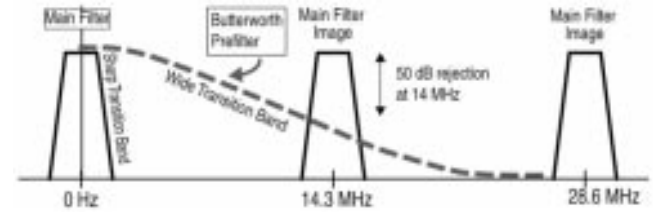


Fig. 5. Cascade of low-order prefilter and high-order main filter. Prefilter suppresses images of main filter while controlling capacitor spread and power dissipation.

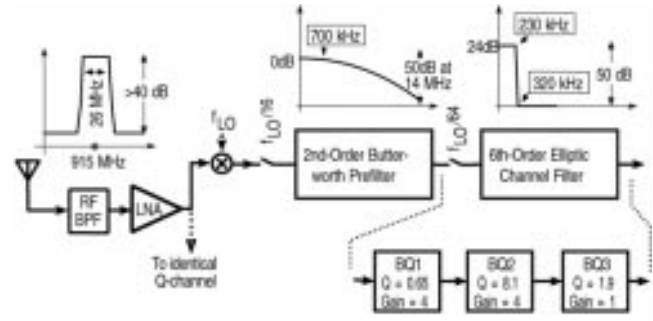


Fig. 6. Final filter, in the context of entire receiver.

Here we use $f_C = f_{LO}/64 = 14.3$ MHz, and $N = 4$, so the prefilter samples its continuous-time input at 57.2 MHz. The first passband image of the prefilter at 57.2 MHz lies in the stopband of the RF preselect filter, and well beyond the output cutoff frequency at a few MHz of the conventional downconversion mixer which precedes it in the receiver. A second-order Butterworth low-pass section realizes the prefilter, with a -3 -dB frequency of 700 kHz, and more than 50 dB rejection at 14.3 MHz (Fig. 5). The maximally flat passband of the Butterworth filter adds only 0.1 dB of droop within the 240 kHz-wide passband of the channel-select filter.

The final filter comprises a cascade of four biquad blocks (Fig. 6). The first block implements the second-order Butterworth prefilter, and the following three blocks make up the sixth-order elliptic channel-select filter.

IV. CIRCUIT DESIGN

The circuit diagram of the SC biquads used in the filter is shown in Fig. 7. Appropriate biquad architectures [25] implement the low- Q ($= 0.65$) and the high- Q ($= 1.9, 8.1$) pairs of poles in the main filter. The capacitors are fabricated as polysilicon plates over gate-oxide, with a heavily-doped N-type layer underneath. The smallest capacitor in the circuit, 215 fF, consists of two unit capacitors in antiparallel.

A. Filter as Subsampling Mixer

The received signal may be downconverted to baseband by a continuous-time mixer, or alternatively by subsampling. The well-known continuous-time mixer is implemented with a Gilbert analog multiplier or some variant. Downconversion by subsampling, on the other hand, makes use of the generalized sampling theorem, which states that no information in the

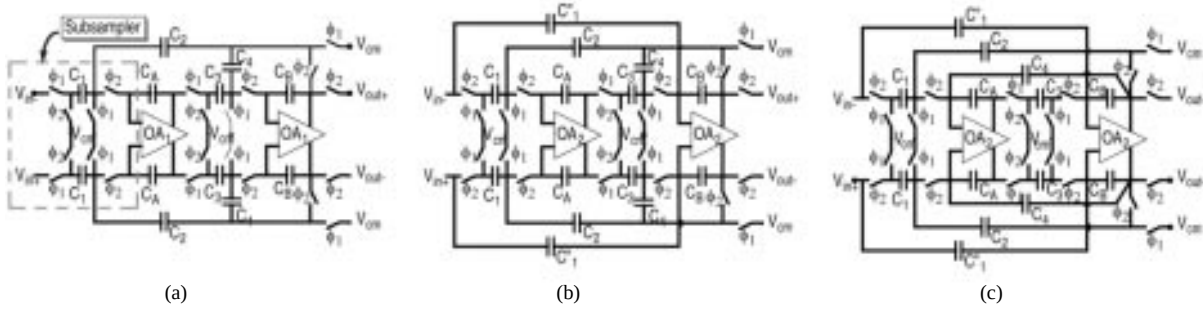


Fig. 7. SC biquads used in prefilter, low- Q section, and high- Q sections, respectively.

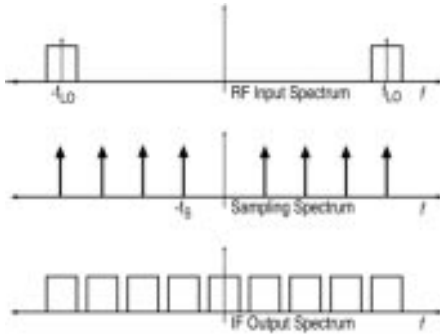


Fig. 8. Downconversion by subsampling, represented in the frequency domain.

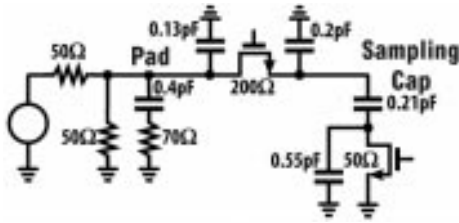


Fig. 9. Equivalent circuit of prefilter input designed to track 900 MHz RF. The various parasitic capacitors and resistors are shown.

original signal is lost if the signal is sampled at a rate greater than twice the *signal bandwidth*. The narrowband RF channel is sampled at a rate much lower than the carrier frequency, but at more than twice the channel bandwidth. When the sample rate equals the carrier frequency divided by an integer, the modulation is translated to dc (Fig. 8) [2]. A track-and-hold (T/H) consisting of MOS switches, two sampling capacitors, and a balanced differential operational amplifier is well-suited to subsampling. The time constant of the switch resistance and sampling capacitor is chosen for a wide acquisition bandwidth, but the opamp clocks at the low sample rate.

As an experiment, the input switched-capacitor and the first opamp in the Butterworth biquad were designed to downconvert the input by subsampling. A minimum-size capacitor of 215 fF samples the input, and the switch size is chosen to minimize the ON-resistance without overloading the low-noise amplifier (LNA) preceding the filter. The equivalent circuit is carefully designed to track the RF input on the sampling capacitor (Fig. 9).

We emphasize that this is not how the filter is finally used in the receiver. Downconversion by subsampling suffers from the fundamental disadvantage of a high noise figure, because in sampling a narrowband signal at RF, the mixer also acquires *wideband noise* extending from dc to RF and aliases it into the much lower Nyquist bandwidth. For white noise, this raises the noise spectral density by the subsampling factor, as has been verified in previous reports on downconversion by subsampling [2], [19]. Therefore, downconversion by subsampling is not to be recommended at the RF frontend, although it may have its uses in the IF section of a receiver, where it no longer aliases noise which has been bandpass-filtered to less than half the sample rate in preceding stages. Here we examine the feasibility of a merged mixer and filter as an object of scientific curiosity.

B. Gain Assignment

For two good reasons, both to do with lower noise, this filter must also amplify the signal. First, with amplification in the first stage of the filter cascade, later stages contribute less to the overall noise. Second, for a given feedback capacitor, gain in the first biquad requires use of a larger sampling capacitor, which therefore also contributes a lower kT/C noise.

Gain is assigned to each block in the cascade to balance signal amplification and lower noise on the one hand, and on the other hand to preserve good large signal handling. That is, with too high a gain in the blocks, the filter input noise becomes so small that it is overwhelmed by the noise of the preceding stages, but now a small input signal to the filter clips the output and compromises the overall receiver dynamic range.

Various gain assignments were considered to optimize the dynamic range of the filter, and finally a gain of $4\times$ is assigned to both the low- Q and high- Q biquad sections, to obtain a total filter gain of $16\times$, or 24 dB. The capacitors C_1 and C'_1 set the gain of the biquads in the main filter (Fig. 7). In addition to improving dynamic range, this distribution of gains flattens the input noise spectrum across the passband. The prefilter biquad on this test chip is designed to subsample a 900 MHz input unity gain, so its input capacitor is minimum size, resulting in unity gain in the biquad.

The capacitance spread, that is, the ratio of the largest to the smallest capacitors in the circuit, is approximately set by the ratio between the sampling rate and the corner frequency

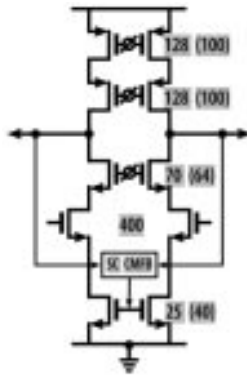


Fig. 10. Opamp circuit. FET widths in μm shown for low clock rate opamp OA 2 (widths for high clock rate opamp OA 1 in parentheses).

of the filter [25]. This predicts a spread of roughly 60 in this elliptic filter. However, larger internal capacitors lower noise, so the final spread for the entire filter is 110.

C. Transconductance Amplifier Design

The opamp determines filter linearity and power consumption. Of the eight opamps in the filter, the first two clock at 57.2 MHz and the rest at 14.3 MHz. The current drain of the two versions is 0.59 mA and 0.47 mA, respectively, and the transconductance is 4.6 mS and 3.9 mS, respectively. These are optimum for the intended operating frequency.

Due to finite dc gain in the opamp, the realized Q of the complex-conjugate poles will deviate from the designed value, leading to either ripple or droop in the filter passband. However, the detector following the filter is insensitive to the resulting slightly unequal amplitudes of the FSK tones. A simple differential single-cascode opamp with SC common-mode feedback (Fig. 10) provides adequate dc gain to satisfy this requirement. As is described below, incomplete settling at the opamp output is the larger source of error.

The opamp is designed for high-frequency operation. Generally, low-power design results in low g_m , while high-frequency operation requires high g_m . Increasing the input transistor size at a given power consumption raises g_m —doubling the transistor size raises g_m by about $\sqrt{2}$. However, the junction capacitance at the drain of the input FET also gets larger with the FET size and lowers the frequency of the second pole, eventually causing instability. Therefore, the input transistor is sized up to the point that the opamp retains a sufficient phase margin.

The output of the opamp clips at a differential voltage swing of ± 2.2 V.

V. EXPERIMENTAL RESULTS

The test chip of the channel-select filter is implemented as a fully balanced circuit in a single-poly, double-metal $1\text{-}\mu\text{m}$ CMOS process and occupies an active area of $2.1 \times 2.4 \text{ mm}^2$ (Fig. 11). A total of 200 pF on-chip capacitance takes up about 60% of the area. The input leads of the IC package are terminated in 50Ω . The balanced outputs, taken off-chip through source follower buffers, are converted to single-ended in a passive power combiner. A balanced input is applied to the

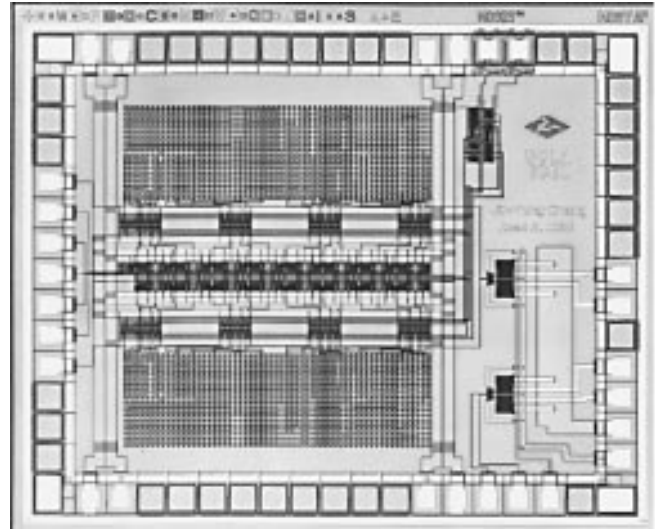


Fig. 11. Chip photograph.

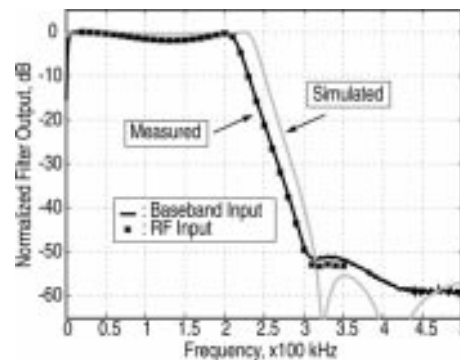


Fig. 12. Measured frequency response. Passband details with baseband and RF inputs.

filter IC through an off-chip passive power splitter. The filter is tested with a baseband input, and then with an RF input which the filter downconverts by subsampling at the first stage. The results of both these tests are now presented.

A. Filter Frequency Response

The measured filter passband response corresponds very well to the expected response as shown in Fig. 12, except for an additional 1 dB droop in the passband attributed to incomplete settling in the opamps of the high-gain biquads. The corner frequency is at 215 kHz, and the stopband starts at 300 kHz. The frequency response remains unchanged when the prefilter subsamples a frequency offset from 915 MHz, accomplishing downconversion and filtering.

Fig. 13 shows the stopband of the filter measured on a large frequency scale. As expected, the Butterworth prefilter suppresses the first three images of the elliptic filter passband, in effect extending the stopband to 57 MHz. Interferers beyond this frequency lie in the stopband of the RF preselect filter, which suppresses, for instance, the passband image at 57 MHz by an additional 35 dB. The measured stopband attenuation is at least 50 dB up to 57 MHz, and limited to at most 65 dB by parasitic effects such as on-chip signal feedthrough.

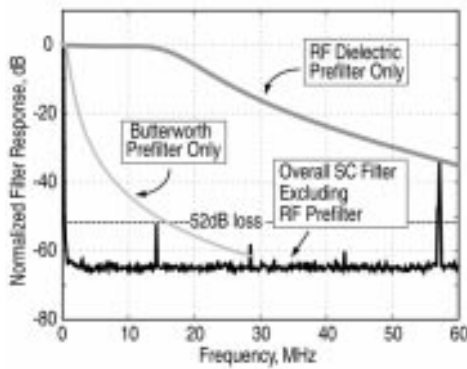


Fig. 13. Measured frequency response. Stopband measured on large frequency scale.

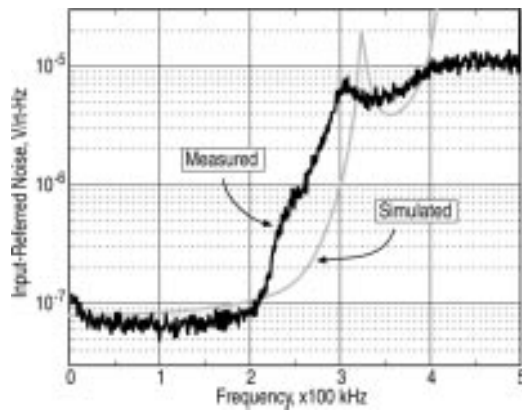


Fig. 14. Measured input-referred noise spectrum.

The overall filter response measurement shows that when used in a quadrature receiver (Fig. 1), this filter selects a 480 kHz-wide channel from a band hundreds of MHz wide, while rejecting out-of-band signals by at least 50 dB. The RF test proves that this filter can also downconvert by subsampling without any degradation in its frequency response.

B. Filter Noise and Receiver Performance

The input-referred noise plot is obtained by dividing the noise spectral density at the output by the measured frequency response of the filter (Fig. 14). A flat input-referred noise level of about $70 \text{ nV}/\sqrt{\text{Hz}}$ is obtained in the passband, as expected from the simulation, which shows that the gain distribution indeed suppresses high-frequency noise peaking in the high- Q biquad. The flicker noise corner lies below 20 kHz. The slight discrepancy between the measured and simulated noise spectra in the transition band is the same as the difference between the measured and simulated filter response. Noise is simulated on SPICE by replacing every switched capacitor, C , in the circuit with the equivalent continuous-time resistor $1/(f_c C)$, where f_c is the clock frequency. In addition, the prefilter noise level is multiplied by the decimation factor of $4\times$ to capture the effect of noise aliasing in the main filter.

As the received signal in a direct-conversion receiver is only amplified by 20 or 30 dB in the RF front-end before it enters the baseband section, noise in the baseband circuits can

easily further degrade the signal-to-noise ratio. This reflects as a lower receiver sensitivity, or, equivalently, a higher input-referred noise figure. The total filter input-referred noise of $70 \text{ nV}/\sqrt{\text{Hz}}$, or $-130 \text{ dBm}/\text{Hz}$ referenced to a hypothetical 50Ω resistor, implies that with 30 dB of conversion gain in the RF frontend, the receiver noise figure due to the *filter noise* alone is 13.6 dB. This is not acceptable. The filter noise must be lowered. In a redesign, the input-referred noise is halved to $35 \text{ nV}/\sqrt{\text{Hz}}$ by merely redistributing the gain in the various filter blocks. As the filter in its final version is not to be used as a downconversion mixer, a larger sampling capacitor is used in the prefilter biquad, which results in a gain of $4\times$ and suppresses the noise contributions of all the later filter blocks. Then, a gain of $2\times$ is assigned to both the first and second biquad of the elliptic filter to suppress the noise contributions, as well as peaking in the input-referred noise spectrum at the filter band-edge. The final filter gain remains $16\times$, or 24 dB. This redistribution of gains does not change the total current drain of the filter.

C. Filter Nonlinearity

An active filter's ability to handle large signals with little intermodulation distortion is of special interest, as its inevitable inferiority to a passive filter in this one respect may disqualify its use in a wireless receiver. There are no standards on how to evaluate this feature of a filter. The sources of distortion which matter in a particular receiver must be identified and then characterized in a way linked to the receiver's intended performance.

Nonlinearity in a balanced baseband circuit is measured by the third-harmonic distortion, or in the case of a bandpass circuit, by the third-order intermodulation created by two in-band tones. Harmonics of a single FSK-tone at 160 kHz in this filter lie in the stopband and are further rejected in the correlating detector [24]. Therefore, harmonic distortion is unimportant in the intended use. Similarly, in the unlikely event that an extraneous interferer appears in the filter passband, the receiver will grossly malfunction because the detector cannot discriminate the interferer from the legitimate FSK tone. Therefore, intermodulation distortion between the two is inconsequential.

On the other hand, it is more likely that third-order intermodulation between certain combinations of two tones in the *stopband* produces spurious products in the passband at the frequency of one of the expected FSK tones (Fig. 15). This nonlinearity is characterized by plotting the third-order intermodulation between 320 and 480 kHz at the filter output on the same scale as its response to a single 160 kHz tone, and extrapolating the intercept point from this data. The measured input IP3 is 30 dBm (Fig. 16). This plot of third-order intermodulation is unlike an amplifier in that it includes filtering of the input tones. The two tones at 320 kHz and 480 kHz are significantly attenuated by the filter while the 160 kHz tone is not. The experiment is repeated by translating all these frequencies by 915 MHz, which the prefilter biquad downconverts to baseband. The IP3 is now lowered to 21 dBm. The measured input IP3 of a previously reported unity-gain subsampling mixer [2] is also about 22 dBm, which suggests

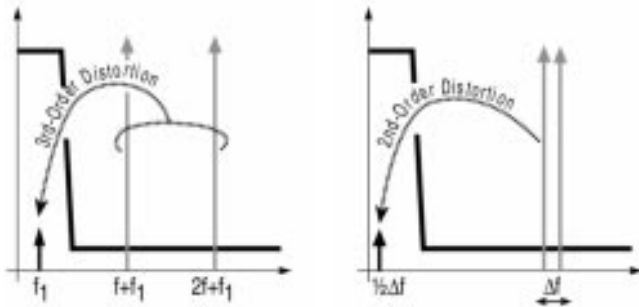


Fig. 15. Two likely sources of distortion in filter. Intermodulation between out-of-band signals (left) and AM detection of out-of-band signal (right).

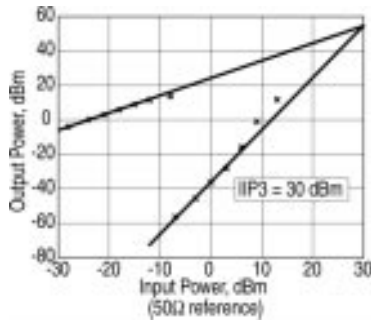


Fig. 16. Measured third-order intermodulation due to out-of-band signals, and intercept point.

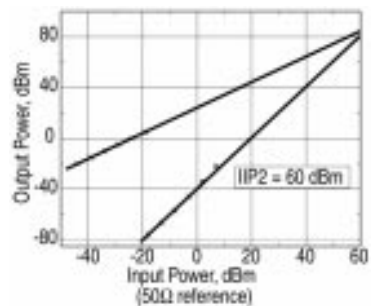


Fig. 17. Measured second-order envelope detection of out-of-band signal and intercept point.

that this may be the limit to attainable linearity in a 3.3 V subsampling front-end.

Second-order distortion in the filter frontend affects receiver performance in a different way, by detecting the time-varying envelope of an AM signal in the stopband and translating it to baseband (Fig. 15). This is shown by the following trigonometric relation, where a carrier frequency, f_0 , is amplitude-modulated at index ϵ by the tone f_m , and only the baseband term is retained on the right-hand side

$$[(1 + \epsilon \sin \omega_m t) \sin \omega_0 t]^2 \rightarrow \epsilon \cos \omega_m t.$$

If f_m lies at one of the expected FSK tones, it triggers a spurious response in the detector. The second-order AM detection is measured with a 1 MHz carrier amplitude-modulated by 50 kHz at a 100% index applied to the filter input. The input IP2 is about 60 dBm (Fig. 17). This shows a very small second-order nonlinearity, chiefly due to the strong rejection of second harmonic by a balanced circuit.

The filter's input intercept point due to both the third-order intermodulation and the second-order AM detection is higher than the 28 dBm output intercept point of a previously published RF front-end [26], which means that when the two are cascaded in a receiver, the filter will preserve the dynamic range of the front-end. The high linearity of the active filter results from careful gain distribution among the various stages.

D. Current Drain

The circuit operates at a 3.3 V power supply. The filter drains 4.6 mA of current as expected from the simulation, which corresponds on average to 0.58 mA/pole, including the two prefilter poles. The clock divider logic drains another 1.3 mA. The filter current is lower than the extrapolated current drain of previously published continuous-time bipolar channel-select filters when they are scaled to the same noise level. For instance, a 7.5 kHz sixth-order elliptic low-pass filter for a paging receiver [14] will dissipate 8 mA after scaling, while a 172 kHz third-order Butterworth filter will require 30 mA [16]. Neither filter's stopband has been evaluated to 57 MHz, nor are they capable of handling large signals as well as this CMOS circuit. A BiCMOS RF frontend switched capacitor filter using biquads [19] dissipates a total power of 112 mW from a 3.3 V supply including the clock generator, and offers a minimum stopband attenuation of 38 dB with input IP3 of -23 dBm.

A more germane comparison is with the aforementioned 600 kHz 0.6- μ m CMOS continuous-time channel-select low-pass filter [22]. That filter operates at 3 V, its IP3 of 34 dBm is comparable to ours, its input-referred noise is 42 nV/ $\sqrt{\text{Hz}}$ slightly higher than 35 nV/ $\sqrt{\text{Hz}}$ in our redesign, and it drains 0.5 mA/pole, whereas our redesigned filter, which includes a prefilter with two poles of its own, drains 0.58 mA/pole. It is noteworthy that for this building block, two different design approaches lead to a remarkably similar total performance.

E. DC Offset

The nonautozeroed integrators in the filter contribute dc offset, which is amplified and may saturate the subsequent limiting amplifier. The measured output offset ranges from 80–200 mV. Owing to a dc biasing feedback in the limiting amplifier which is decoupled by an off-chip capacitor, an input offset of ± 100 mV may be absorbed with little loss of dynamic range [5]. A higher offset will require on-chip ac coupling of the signal. Therefore, filter offset is not a limiting factor in this receiver.

VI. CONCLUSION AND DISCUSSION

The portable wireless transceiver of the future requires a high level of integration and low-power dissipation. The filter block, however, has traditionally not been amenable to integration.

This work successfully demonstrates a highly selective and highly linear channel-select filter using a switched-capacitor circuit in 1- μ m CMOS technology. In concert with standard

RF preselection, this filter is capable of selecting a 480 kHz-wide baseband channel out of a bandwidth of over hundreds of MHz. Its low noise and exceptional linearity make it a suitable choice for this application. A zero-IF receiver makes it easier to attain these very useful attributes in the active filter.

As an exercise to enable comparison with filters for well-known wireless standards, this filter has been modified in simulation to satisfy the GSM channel-selection specifications in a direct-conversion receiver, such as the one from Alcatel [27]. The key requirements are for a 300 kHz-wide passband whose gain varies by less than 1 dB and group delay by less than 1 μ s, a transition band of 250 kHz, and a stopband rejection of 35 dB. Although a fifth-order Butterworth filter satisfies all requirements, a sixth-order filter is designed here to avail the existing design tools. With a passband gain of 24 dB, the input-referred filter noise spectral density is 20 nV/ $\sqrt{\text{Hz}}$, and there are no high- Q stages to raise this noise near the passband edge. Extrapolating now to a fifth-order implementation preceded by a second-order prefilter, this GSM channel-select filter will drain only 4.1 mA. Assuming a 34 dB conversion gain in the front-end, it also meets the GSM specifications on receiver noise figure and IIP3.

This filter is the final linear block in the receive chain to realize a single-chip 900 MHz direct-conversion wireless receiver in 1- μ m CMOS.

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