It is pointed out that, under very noisy conditions, a better performance can be achieved through the AFM algorithm.

The hybrid-series structure has shown a better performance than conventional IIR and FIR structures, except in a few cases where the echo path has complex poles. However, in this case, it is possible to improve the performance by increasing the order of the structure without an heavy increase in computational effort, since the signal processing required by the new structure is very simple.

We have studied the new structure as an echo canceller for data transmission, but the results may easily be extended to other system identification situations.

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A Digital FM Demodulator for FM, TV, and Wireless

Bang-Sup Song and In Seop Lee

Abstract— **An FM demodulator is implemented digitally in software using a quadri-correlator algorithm so that the FM function can be made compatible with future digital wireless and FM receiver systems. The proposed digital FM demodulator uses a sinc^3 decimation filter with its first zero either on the alternate channels or on the adjacent channels for high channel selectivity, a digital differentiator using a three-point approximation for frequency discrimination, and a digital division for AM rejection. A bitstream FM signal from a fourth-order bandpass delta-sigma modulator is FM-demodulated to exhibit a SNDR of 71 dB, a THD of 0.01%, and an AM rejection of 77 dB in simulations using a signal bandlimited to $\frac{1}{200}$ of the sampling frequency and amplitude-modulated with a modulation index of 0.9 (90% AM).**

I. INTRODUCTION

In superheterodyne FM receivers, channel filtering is done at IF stages using high-Q surface acoustic wave or crystal filters. An effort to integrate superheterodyne receivers using IC technologies has been facing a major bottleneck of IF filtering [1], and alternative zero-IF architectures based on analog baseband processing have been limited by the performance of analog circuits [2], [3]. However, in an RF receiver architecture recently proposed, the IF stage is replaced by an anti-aliasing bandpass filter followed by a bandpass delta-sigma modulator [4], whose output can be digitally quadrature-demodulated and lowpass-filtered. The bitstream output of a bandpass modulator at a sampling rate of 4 times the passband center frequency is ideally suited for digital quadrature demodulation [5], [6]. In this work, the inphase and quadrature (I–Q) components are digitally low-pass-filtered and FM-demodulated at baseband so that the FM function can be implemented in software to be compatible with future digital systems.

Existing digital FM demodulation systems use either a phase-locked loop or a tangent method [7]–[9]. Recently, a delta-sigma modulation technique was applied to quantize instantaneous frequencies of an FM signal [10], [11], but the performance of such an FM discriminator is limited by a front-end analog phase/frequency detector. The digital tangent method, though promising, requires a jump detector for broadbanding to avoid numerical division by zero and a $\tan^{-1}$ function or a ROM look-up table. This brief is based on a quadri-correlator algorithm [12], which doesn't need numerical division for frequency discrimination. In the proposed work, numerical division is still used, but for a different purpose to reject AM components. The quadri-correlator has been used to aid frequency acquisition in phase-locked loops [13], but has suffered from inaccuracy when implemented using analog circuits. Since this work is implemented digitally, it does not exhibit errors resulting from inaccurate analog quadrature demodulation.

In Section II, the principle of baseband FM demodulation is introduced, and in Section III, its digital implementation is considered for practical use. Numerically simulated results are summarized in Section IV.

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The authors are with the Department of Electrical and Computer Engineering, Coordinated Science Laboratory, University of Illinois at Urbana-Champaign, Urbana, IL 61801 USA.

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II. Principle of Digital FM Demodulation

In conventional FM broadcasting receivers, overall receiver performance, measured in terms of THD and adjacent or alternate channel rejection, relies on the quality of IF filters. Even IF filters, which are usually made of crystal and have an order of 14–20, can not reject adjacent channels by more than 10 dB without sacrificing distortion performance. Since linear phase in IF filtering is the most critical factor in achieving low distortion, only linear-phase filters, such as Bessel, can meet the requirement of hi-fi FM receivers. Thus, alternate channel rejection of more than 60 dB is rare in commercial products. However, one advantage of the existing architecture is the simplicity in rejecting AM components. The AM component in an FM signal, unless suppressed, directly contributes to the phase error of the demodulated output. A simple voltage limiter, usually a high-gain high-speed amplifier, has been used to remove AM components. After limited, the FM signal is demodulated using a bandpass frequency discriminator centered around IF. The discriminator is implemented either by two resonators slightly off-centered with opposite polarity or by phase-locked loops [14], and the linearity of the bandpass frequency discriminator limits FM demodulator performance.

In the baseband digital FM demodulator to be described, the signal band is centered at dc, and can not be clipped using a limiter as in the existing architecture. The amplitude component of the signal needs to be normalized numerically, giving the same effect as a limiter. The baseband FM signal can be lowpass-filtered digitally before baseband demodulation is performed. Therefore, equal group delay in channel filtering is easier to achieve digitally in lowpass filters than in bandpass filters. Even adjacent channels can be rejected digitally with a greater degree than possible by analog bandpass filters.

The baseband FM demodulation algorithm is illustrated in Fig. 1. The FM-modulated signal is

$$x_i(t) = e^{a(t)+j\int m(t)+j\omega_c t}, \tag{1}$$

where the time functions $a(t)$ and $\int m(t)$ represent the magnitude and phase of the signal and ω_c is the carrier frequency. The real part of $x_i(t)$ represents the real signal. The demodulation process begins with removing the carrier by modulation equivalent to a frequency shift of $-\omega_c$. In the time domain, the received signal is multiplied by $e^{-j\omega_c t}$ to move the signal to baseband. Therefore, the baseband FM signal becomes

$$\begin{aligned} x_b(t) &= x_i(t)e^{-j\omega_c t} \\ &= e^{a(t)+j\int m(t)}. \end{aligned} \tag{2}$$

For FM demodulation, this baseband FM signal needs to be frequency-discriminated by differentiating it. After differentiation, the baseband signal becomes

$$x_b'(t) = \frac{dx_b(t)}{dt} = \{a'(t) + jm(t)\}e^{a(t)+j\int m(t)}. \tag{3}$$

Now the original signal should be removed from (3) to recover the demodulated output $m(t)$ as in the following.

$$\begin{aligned} x_o(t) &= \frac{x_b'(t)\cdot\{-jx_b^*(t)\}}{x_b(t)\cdot x_b^*(t)} \\ &= m(t) - ja'(t), \end{aligned} \tag{4}$$

where the superscript $*$ represents the complex conjugate. The FM-demodulated output has both real and imaginary parts, but the real part $m(t)$ represents the FM-demodulated output.

This algorithm, based on a mathematical model, can be implemented by maintaining the I–Q branches, representing the real and imaginary parts of the signal, as shown in Fig. 2. The mathematical frequency shift operation of multiplying by $e^{-j\omega_c t}$, in (2), is in effect identical to multiplying the digital FM input by two real carriers $\cos\,\omega_c t$ and $-\sin\,\omega_c t$ and lowpass-filtering them. In the implementation of Fig. 2, the output polarity is inverted so that $\sin\,\omega_c t$ may be used in place of $-\sin\,\omega_c t$. That is, the baseband I and Q components are obtained by multiplying the input by periodic bit sequences of $\{1,\ 0,\ -1,\ 0\}$ and $\{0,\ 1,\ 0,\ -1\}$ representing cosine and sine carriers, respectively. The I and Q branches are equivalent to the real and imaginary parts of the complex baseband FM signal. To separate the baseband signal at dc from the neighboring ones, the I–Q branches are lowpass-filtered first. The product $x_b(t)\cdot x_b^*(t)$ of (4) is obtained by squaring the I–Q signals and summing them. By cross-multiplying the I and Q signals by their derivatives, taking the difference of them, and dividing it by the squared sum of I and Q signals, the final demodulated output $m(t)$ is obtained.

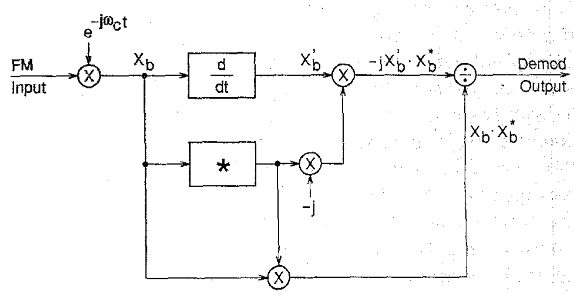

Fig. 1. Principle of the baseband FM demodulation.

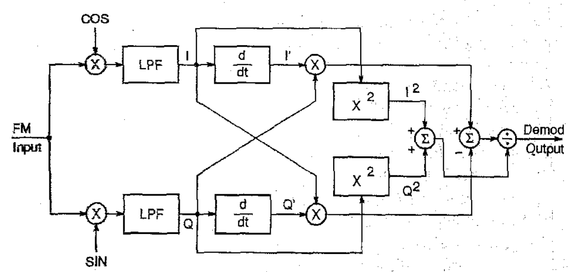

Fig. 2. Simplified block diagram of the baseband FM demodulator.

III. Design of Baseband Digital FM Demodulator

The performance of a digitally-implemented baseband FM demodulator depends on the digital filter and differentiator accuracy as well as on input FM signal quality. A detailed block diagram of the proposed digital FM demodulator is illustrated in Fig. 3. The input FM signal at IF, which is $\frac{1}{4}$ of the sampling frequency f_s, is converted to a digital bitstream by a 4th-order delta-sigma bandpass modulator [15]. The I–Q signals are then low-pass-filtered and decimated by the sinc^3 filter [16]. Overall demodulator performance is mainly limited by the gain droop of this lowpass filter. Since it is difficult to implement a wideband linear digital differentiator, the overall system is configured so that a high degree of linearity may be obtained only inside the passband. The magnitude information is used to suppress AM components as well as to feed an AGC servo loop to keep the input level to the demodulator constant. Therefore, low-frequency AM components are rejected by the AGC loop, but high-frequency AM components are rejected by numerical division. Numerical division by zero or by small numbers causes an overflow, and it should be avoided by detecting the condition before it occurs.

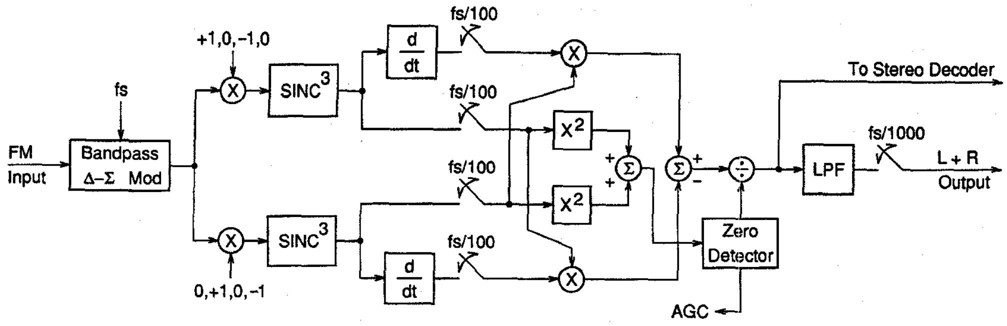

Fig. 3. Detailed block diagram of the digital FM demodulator.

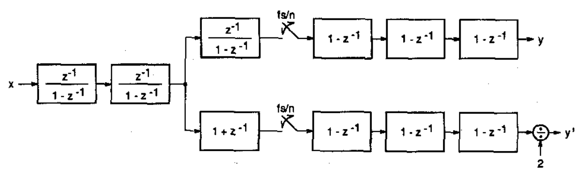

Fig. 4. Schematic of the sinc^3 filter and the differentiator.

A. Decimator

The out-of-band noise of a bitstream from a bandpass delta-sigma modulator should be digitally filtered. To get a high SNR in FM demodulation, the low-pass filter should have a linear phase characteristic with a sharp cutoff and a flat passband. A general FIR filter for this requires at least two separate multipliers for I and Q branches since the sampling rate is high. Therefore, the sampling rate needs to be lowered prior to digital filtering. Although the magnitude response is not flat in the passband, the sinc^3 filter is the simplest form of a low-pass filter that can be implemented using only accumulators. In Fig. 4, the top row is a sinc^3 filter using modulo arithmetic [17], and the bottom row is a differentiator using a three-point formula.

One advantage of the sinc filter is that it is possible to reject neighboring channels by placing its zero at the frequency to reject. In the proposed system, the zero's of the sinc^3 filter are placed at the center of the alternate channels. Because the FM spectrum has a significant portion of its power around the carrier frequency, this rejection by zero is very effective. However, for extremely narrowband applications requiring adjacent channel rejection, an additional sinc comb filter can be inserted into the signal path so that its zero can be placed at the center of the adjacent channels. The passband droop due to this extra sinc filter degrades distortion performance by about 10 dB in simulations. After rejecting the noise and channel interference by the sinc^3 filter, the sampling rate is lowered to a frequency of 4 times the FM bandwidth for demodulation.

B. Digital Differentiator

A wideband digital differentiator is inadequate for practical use because the signal bandwidth is very narrow compared to the sampling frequency. In this work, a differentiator approximated by a three-point formula is used before the decimation stage so that its error can be reduced by operating it at a high oversampling rate. The three-point formula for digital differentiation is

$$f'(t) \approx \frac{f(t+\Delta t) - f(t-\Delta t)}{2\Delta t}, \tag{5}$$

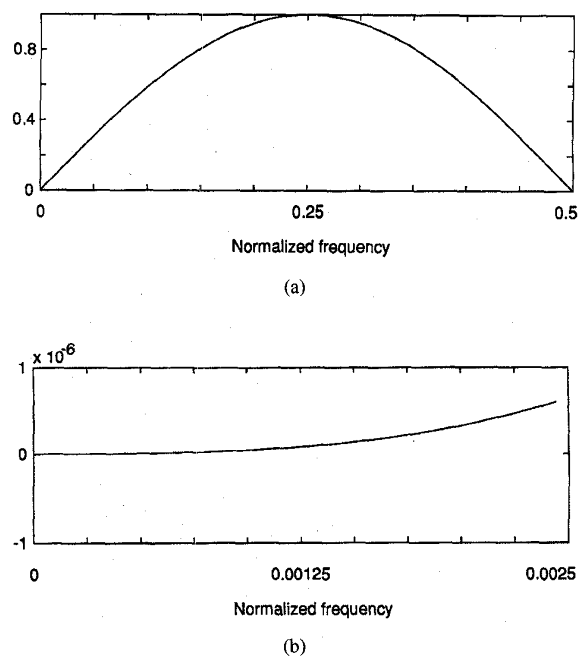

Fig. 5. (a) Frequency response of the differentiator and (b) error in the three-point approximation.

and it has, by omitting higher order terms, an error of

$$\varepsilon \approx -\frac{(\omega\Delta t)^2}{3!} \times 100(\%). \tag{6}$$

This error is proportional to the sampling interval Δt and the number of points in approximation. Therefore, for accurate digital differentiation, the oversampling ratio should be kept high. Otherwise, given the low sampling rate, the accuracy of digital differentiation can only be improved by increasing the number of points in the approximation.

The digital differentiator used in this work is implemented as shown in the bottom row of Fig. 4, and its frequency response and detailed passband are shown in Fig. 5 for a 100 times oversampling case. The passband flatness can be improved, but the maximum approximation error in the passband of this digital differentiator is less than $-$ 90 dB. Considering the fact that AM rejection is limited to about 77 dB due to the sinc^3 filter used as a channel filter, the digital differentiator using the three-point formula will suffice.

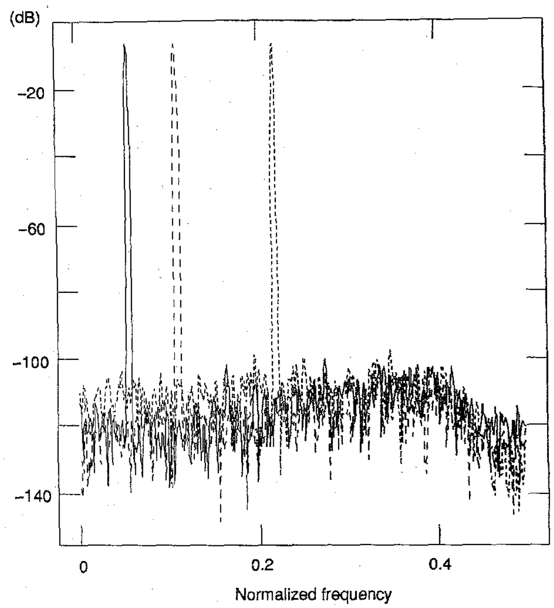

Fig. 6. Spectrum of the FM-demodulated output of different frequencies.

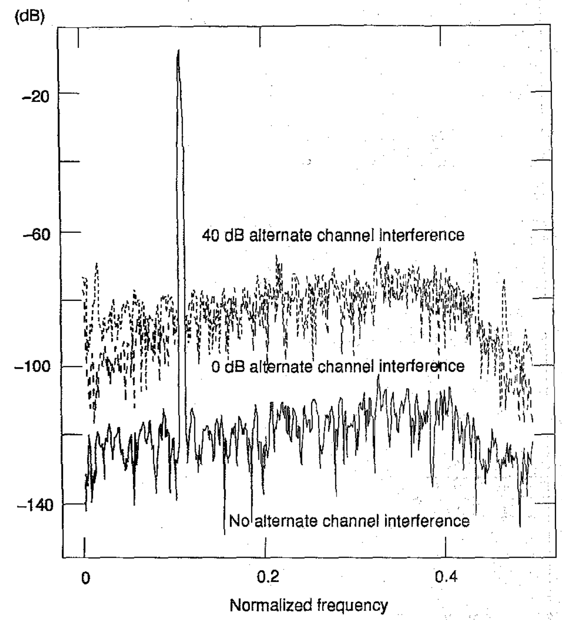

Fig. 7. Spectrum of the FM-demodulated output with interferences.

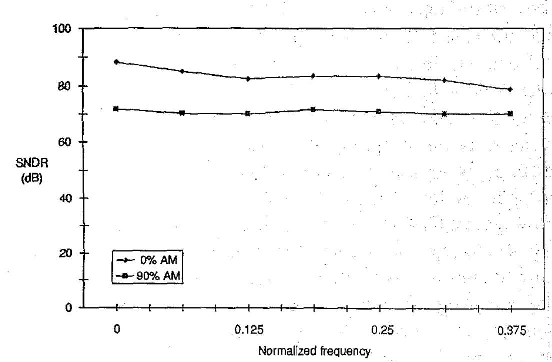

Fig. 8. SNDR versus input frequency.

C. Output Low-Pass Filter

The output of the FM demodulator does not need extra filtering, but in commercial FM broadcasting, a 19 kHz pilot signal which is a reference for stereo decoding needs to be separated, and the left and the right channels should be recovered. In this work, the stereo decoder is not covered, but the output lowpass filter for the $L + R$ sum signal is included in simulations to see the performance degradation resulting from this extra filtering. The output lowpass filter added to the system is a 11th-order Chebychev type II filter. Since the bandwidth of the passband is narrow compared to the sampling frequency, the output is decimated again to a sampling frequency which is 10 times the signal bandwidth using a sinc^2 decimation filter. This oversampling ratio is high enough to keep the attenuation of the decimation filter at the passband edge within 1 dB. After the output lowpass filtering, the final output is resampled at a Nyquist rate.

IV. Simulated Results

The system to be described is based on an assumption that a 10-MHz input can be sampled at 40 MHz to achieve a 12-bit dynamic range within a 300-kHz bandwidth. Signal bandwidth as well as FM spectrum bandwidth may vary from system to system. Different ratios can be implemented by adjusting oversampling ratio, but the example shown is typical of a wideband FM system such as commercial FM broadcasting. The message bandwidth of the example shown in Fig. 3 is $\frac{1}{2000}$ of the sampling frequency or $\frac{1}{500}$ times the IF frequency while the channel spacing is 10 times wider than the message bandwidth. The maximum frequency deviation is approximately $\frac{1}{3}$ of the channel spacing. In reality, the performance of this system will be limited by input signal purity, but simulations results presented are valid only within the limits of digital processing and dynamic range of a fourth-order bandpass delta-sigma modulator. To reduce digital errors resulting from truncation, the number of bits internally processed is increased to 24 bits.

In simulations, 2^{20} samples are calculated to estimate FM demodulator performance. Shown in Fig. 6 are three FM-demodulated tones of different frequencies. Note that the distortion level is consistently low. However, the noise floor rises if the interference from the alternate channels are considered as shown in Fig. 7, where up to 40-dB stronger alternate channels are added to the signal. The signal-to-noise-and-distortion ratios (SNDR) for different input tones are shown in Fig. 8 with and without 90% AM modulation. The simulated AM rejection and THD are 77 dB and 0.01%, respectively.

V. Conclusion

The feasibility of digitally implementing a high-quality FM demodulator with AM rejection is demonstrated by processing the bitstream output of a fourth-order bandpass delta-sigma modulator. The noise floor is limited by the order of a bandpass modulator and the number of digital bits as well as by a baseband lowpass filter. The proposed demodulation system can greatly simplify the task of realizing FM receivers for dual-mode wireless applications as well as for existing FM broadcast receiver or TV audio applications. To date, all digital FM demodulators have not been implemented probably due to high cost, complexity, or limited digital processing speed, but considering the rapid advance of CMOS technologies, integrating the proposed all-digital FM demodulator for commercial FM receivers is considered feasible with a 40 MHz clock using 0.8 μ CMOS and well within the capacity of current VLSI.

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Optimum Analog Preprocessing for Folding ADC's

P. E. Pace, J. L. Schafer, and D. Styer

***Abstract*—Folding analog-to-digital converters (ADC's) that use symmetrical number system (SNS) preprocessing, require fewer comparators than those that use conventional binary encoding. This paper considers an alternate SNS definition that considerably extends the dynamic range of the SNS ADC. The efficiency of this definition is compared to previous definitions and is shown to be *optimum*. As an example, a unipolar 7-b SNS ADC using pairwise relatively prime moduli $m_1 = 4$, $m_2 = 5$ and $m_3 = 7$ is evaluated numerically. Transfer functions are shown which detail encoding errors that occur when the folded input samples lie at one of the code transition points. To discard the encoding errors that occur, a *decimation band* is constructed at each transition point. The effectiveness of the decimation band in eliminating the encoding errors is also quantified.**

I. Introduction

Analog-to-digital converters (ADC's) are critical building blocks in a wide range of hardware, from radar and electronic warfare systems to multimedia based personal computers and work stations. The need constantly exists for converters with higher resolution, faster conversion speeds, and lower power dissipation. A straight $n = 10$-b flash ADC for example, requires $2^n - 1 = 1023$ comparators loaded in parallel. To reduce the number of power consuming components, high performance ADC's employ a parallel configuration of analog folding circuits to symmetrically fold the input signal prior to quantization by high speed comparators (analog preprocessing) [1]. For a 10-b ADC, two parallel channels can be used to amplitude analyze the signal; one for the five least significant bits (LSB's) and one for the five most significant bits (MSB's). The MSB's are determined by direct (coarse) quantization with a 5-b flash converter. The second channel uses analog preprocessing to symmetrically fold the input signal into a repetitive triangular shaped output. The LSB's are obtained by (fine) quantizing the triangular output also with a 5-b flash converter. A delay adjustment is required between channels since the MSB's are digitized directly. A total of 62 comparators are required for this approach with 31 loaded in parallel in each channel.

To reduce the number of required comparators in the folding ADC, a preprocessing scheme that incorporates a symmetrical number system (SNS) encoding has been described [2]. The SNS preprocessing is used to decompose the analog amplitude analyzer operation into a number of parallel suboperations (moduli) which are of smaller computational complexity. Each suboperation symmetrically folds the analog signal with folding period equal to the modulus. A small comparator ladder midlevel quantizes each folded output. Each suboperation only requires a precision in accordance with that modulus. A much higher resolution is achieved after the N different SNS moduli are used and the results of these low precision sub-operations are recombined. The parallel use of folding circuits increases the ADC resolution without increasing the folding rate

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P. E. Pace and J. L. Schafer are with the Department of Electrical and Computer Engineering, Naval Postgraduate School, Monterey, CA 93943 USA.

D. Styer is with the Department of Mathematical Sciences, University of Cincinnati, Cincinnati, OH 45221 USA.

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