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Integrated DSP/RF Design for an MSAT Transmitter

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ABSTRACT

In mobile radio systems, the relatively inefficient use of the spectrum by existing FM modulation techniques is limiting the available channels. Linear modulation methods are thought to provide the necessary push into more efficient useage of the spectrum. The limitation, to date, of the linear modulation techniques is the fact that the nonlinear power amplifier tends to spread the spectrum and thus offset any spectrum efficiency advantage. Linearization techniques are presently being investigated, in the literature, in order to recover some of the lost spectrum, due to spreading. To this end, DSP techniques are thought to provide real time predistortion methods. Some of the complex modulation blocks, normally confined to analog circuitry, are being implemented using DSP circuitry. This paper will address the linearization techniques, for MSAT tranceivers, that use an integrated DSP/RF design.

INTRODUCTION

To date, mobile radio systems use FM modulation schemes like MSK,TFM, and others. Even though FM modulation is inefficient in spectrum useage, it is still widely used because of its constant envelope property. The constant envelope property provides the use of power efficient nonlinear amplifiers, at the same time as achieving low out-of-band radiation (-60 dB to -70 dB).

The next generation of digital mobile radio systems will use Quadrature Amplitude Modulation (QAM) patterns as a means of increasing the spectrum efficiency beyond that of FM modulation schemes. However, since QAM presents a variant envelope, it will be necessary to use more linear less-efficient power amplifiers. Linear amplifiers have the disadvantage of poor power efficiency to the extent that they cannot satisfy the required -60 to -70 dB adjacent channel interference power level. While, this severe constraint on adjacent channel interference may be relaxed for certain systems such as cellular mobile telephone systems, it remains as a requirement in general mobile radio systems.

The extremely high degrees of transmitter linearity required for operation with an acceptable overall system bit-error-rate (BER) performance is difficult to obtain. Because of the large number of possible states in the transmitted phase-plane, as well as the high peak-to-average signal power inherent in QAM modulations, any degradations due to transmitter nonlinearity can become quite severe. Transmitter intermodulation distortion (IMD), which can be observed on a QAM phase-plane as a combination of AM-to-AM and AM-to-PM, moves the transmitted states closer to decision thresholds, degrading the operating margins and hence attainable system BER. In addition to the increase in BER the distortion will cause spectrum spreading, which will place a limit on the maximum system data rate allowable to maintain within the FCC frequency mask limitations. In order to achieve the optimum tradeoff between spectrum spreading and efficiency, techniques for linearizing the power amplifier must be adopted.

Linearization techniques can be categorized into one of the following:
1) feed-forward, 2) feedback, 3) LINC and 4)

predistortion. These techniques have conventionally been implemented using analog circuitry. However, the advent of Digital Signal Processing (DSP) devices permits complicated operations to be handled in a simple manner. This paper demonstrates how DSP devices can be integrated into mobile radios.

DSP/RF INTEGRATED ARCHITECTURE

We have developed a digital signal processing (DSP) technique to apply compensating predistortion to the signal, at baseband, before modulation (Figure 1). The complex vector predistortion will track the inverse function of the MSAT power amplifier's AM-to-AM and AM-to-PM distortion. A recent paper[1] discussed this method and demonstrated significant improvement. Central to such baseband approaches is a quadrature modulator to translate the complex signals to an IF. The level of precision required in meeting emmission constraints for mobile applications makes the use of analog technology for quadrature modulation quite difficult. The problems, which include amplitude and phase imbalance, and component drift, can be circumvented by digital implementations.

The benefits to the designer of using the digital quadrature modulator are as follows:

1) It is a precision digital solution to a common design problem. Unlike the analog equivalents, amplitude matching and phase separation are essentially perfect, there is no residual carrier or DC offset, and there is no performance drift. The digital quadrature modulator requires one digital to analog converter (DAC), whereas an analog quadrature modulator driven by a DSP chip would require two DAC's.

2) It simplifies the RF analog design. Upconversion from a 480 KHz IF, for example, simply requires the second IF stage to remove images 960 KHz away, which is a relatively easy filter implementation.

3) The digital quadrature modulator, together with the DSP chip driving it, can act as an universal modulator, for a multimode tranceiver. FM, AM, SSB, ACSB, TTIB, MSK, QPSK, 16QAM, etc, can all be resident in the DSP chip's ROM. and selected as appropriate during operation.

Some of the potential problems with using digital quadrature modulators in mobile transmitter architectures are: the significant DC power consumption; the delay introduced by internal filtering operations on both the digital quadrature modulator and DSP chips; and the reduced spectral purity.

DIGITAL QUADRATURE MODULATOR

General Description

Figure 2 is a functional view of an ideal quadrature modulator (quad mod). The complex baseband signal $v_2(t)$, represented by real and imaginary components $v_{2I}(t)$ and $v_{2Q}(t)$, is modulated to form the real bandpass output $\tilde{v}_3(t)$ with complex envelope $v_3(t)$, at a center frequency f_{c3} :

$$\begin{aligned}\tilde{v}_3(t) &= v_{3I}(t)\cdot\cos(2\pi f_{c3}t) - v_{3Q}(t)\cdot\sin(2\pi f_{c3}t)\\ &= Re[v_3(t)\cdot\exp(j2\pi f_{c3}t)] \qquad (1)\end{aligned}$$

Ideally, the complex envelope $v_3(t)$ equals the input $v_2(t)$. The frequency domain sketches show the input $V_2(f)$ as asymmetric to emphasize the fact that the input is complex. Its bandwidth is W_2 and the corresponding RF bandwidth is $2W_2$. Figure 3 illustrates the operation of the digital quad mod. The inputs are digital, and enter at a relatively low sampling rate f_{s2} . The output is analog, from a DAC driven at a sampling rate f_{s3} , which is precisely 4 times the center frequency f_{c3} . The cos/sin carriers in (1) degenerate to:

cseq: 1,0,-1,0,1,0,-1,0,1,0,-1,0,1 ...
sseq: 0,1,0,-1,0,1,0,-1,0,1,0,-1,0 ... (2)

which eliminates both trigonometric generation and multiplication. Since f_{s3} is significantly greater than f_{s2} , we need more samples of $v_2(t)$ than are available from the input. The intermediate samples are created by interpolation (i.e. lowpass filtering). The periodic zero in (2) means that the I and Q channels supply the output alternately, which cuts computation in half. In a further simplification, we can combine the cseq and sseq sequences with the interpolation coefficients.

One consequence of interpolation is delay, which increases with the length of the interpolator. The delay becomes important when the quad mod is embedded in a feedback loop, as in the case of adaptive amplifier linearization.

Interpolation and Computation

Performing interpolations at the rate f_{s3} is less demanding than it appears, as an example will make clear. Suppose that a 16 point interpolating lowpass filter is sufficient. Figure 4 shows the impulse response and a typical output sequence generated by convolution if the step up ratio $f_{s3}/f_{s2} = 8$. Only two input samples affect each output sample. Since I and Q inputs alternate, only 2 multiply/adds are required per output sample.

More generally, for an N-point lowpass filter, we require:

Nf_{s2}/f_{s3} multiply-adds per output sample,
and
Nf_{s2} multiply-adds per second. (3)

For a fixed output sampling rate, therefore, reducing f_{s2} (provided it is above the Nyquist rate $2W_2$) allows a proportional increase in the number of points.

Interpolator Design

This section develops frequency domain design guidelines for the quad mod interpolator. Although the quad mod algorithm accommodates a wide range of sampling and center frequencies, we will assume specific values, in order to make the discussion and illustration simpler:

$W_2 = 17.5$ kHz (5 kHz MSAT channels, 7th order predistortion)
$f_{s2} = 480$ kHz (to simplify the illustration only)
$f_{s3} = 1920$ kHz, $f_{c3} = 480$ kHz (4)

Figure 5a shows that the input spectrum consists of images with a repetition interval of 480 kHz. The interpolator suppresses images other than the one centered at 0 Hz (Figure 5b), thereby changing the effective repetition interval to 1920 kHz. The greatest attenuation is at the folding frequency of 960 kHz. The complex carrier (1) is also sampled at 1920 kHz, and therefore has the spectrum shown in Figure 5c. Figure 5d is the result of convolving the spectra of the sampled input and the sampled complex carrier, to represent multiplication of the corresponding time domain signals. Finally, the operation of taking the real part of the time domain signal in (1) has a spectral equivalent of superimposing the conjugate reflection, as shown in Figure 5e.

The desired components in Figure 5e are at +/- 480 kHz. The rest may require further suppression by an analog bandpass filter, depending on the application. However, the frequency reversed image superimposed on the desired one at 480 kHz cannot be suppressed by filtering. It derives originally from the input image at 960 kHz.

The most important design objective for the lowpass filter, therefore, is attenuation of at least 60 dB in a 35 kHz stop band centered at 960 kHz. A secondary objective is to maximize the attenuation of other input images, so that the design of any following analog filters is simplified. The attenuation at other frequencies can take any convenient value, since there is no signal component there to require attenuation.

To continue the example, assume that the DSP hardware is capable of 11.52 million multiply-adds per second (6 multiply/adds per output sample at an output sampling rate $f_{s3} = 1920$ kHz). Since the step-up ratio is $1920/480 = 4$, we can have $N = 24$ points in the lowpass filter. Figure 6 shows a multiband design using the Parks- McClellan algorithm with only 16 points, which places 35 kHz stopbands at 480 kHz and 960 kHz. Clearly, the interpolation more than meets the requirements. An analog bandpass is necessary only to suppress harmonics at multiples of 960 kHz from channel center, which is not difficult.

Lower Input Sampling Rates

The 480 kHz input sampling rate used in the example above is unnecessarily high for MSAT applications. However, it can be reduced with no change to the output center frequency or computational load. Suppose the input sampling rate is $f_{s2} = 60$ kHz, sufficient for 7th order predistorted MSAT channels. The images to be removed are 35 kHz wide, centered at multiples of 60 kHz. If we use the same DSP hardware as above, then from (3) the filter length can be N = 192, without change to the computation rate. Figure 7 shows a Kaiser window design which suppresses even the closest images by almost 50 dB.

Number of Bits

The DAC adds quantization noise. Its level relative to the desired signal is easily computed. Assume the peak DAC voltages are +/- 1. Then the quantization noise power per sample is:

$$\sigma_q^2 = (2 \cdot 2^{-n})^2/12 \qquad (5)$$

where n is the number of DAC bits. Assume a typical data signal with a peak to average power ratio of 6 dB. Then the signal power per sample is

$$\sigma_v^2 = 1/4 \qquad (6)$$

The ratio of signal and quantization noise power spectral densities (PSDs) is more important than per-sample variances. Assume for simplicity that the signal spectrum is approximately rectangular with bandwidth W_2, and that successive quantization noise samples are independent (i.e. white). Then the ratio of PSDs of quantization noise and signal is given by:

$$\frac{PSD_q}{PSD_v} = \frac{\sigma_q^2}{f_{s3}} \cdot \frac{W_2}{\sigma_v^2} = \frac{W_2}{f_{s3}} \cdot \frac{4}{3} \cdot 2^{-2n} \qquad (7)$$

For W_2 = 5 kHz, and f_{s3} = 1920 kHz, we can achieve a noise floor of -60 dBc with only 6 DAC bits. However, this figure is valid only for random data. Short period repetitive sequences produce quantization noise line spectra. In this case, 10 to 12 bit DACs are advisable.

PREDISTORTION

The general transmitter configuration is shown in Figure 8. The modem, predistorter, quad mod. and quad demod. are implemented digitally at complex baseband. The RF power amplifier, and up and down converters, which translate between IF and RF, are analog. Since the power amp exhibits both AM-to-AM and AM-to-PM distortion, any modulation with nonconstant envelope will generate adjacent channel interference because of intermodulation products. The function of the predistorter is to apply an inverse nonlinearity, so that the power amplifier output is free of intermodulation products.

The technology under development is a digitally implemented predistorter which operates at complex baseband[2]. Currently, the predistorter can operate at a sampling rate of 240 KHz, which is sufficient for 7th order nonlinearities generated from a standard 25KHz mobile channel.

In a nonadaptive configuration, there is no feedback path. The predistorter coefficients are determined for a given amplifier at the factory and are fixed. Aging and temperature changes will cause the amplifier characteristics to drift, of course, so the predistorter and amplifier are no longer perfectly matched. It is possible, though, that the combination will still meet the relatively lenient MSAT specifications on out-of-band power, and it is much cheaper than an adaptive configuration.

Adaptivity is achieved by feeding back the demodulated amplifier output and comparing it to the desired output, then updating the predistorter coefficients. Note that adaptation can be very slow if all it has to track is drift in the amplifier characteristics; however, a channel switch can introduce larger changes, and readaptation must be rapid to avoid dropouts. The requirement of a receiver operating in parallel with the transmitter means that it may be an expensive technology.

CONCLUSIONS

This paper has described the integration of DSP hardware into a mobile communications transmitter. We have addressed the significance of implementing the quadrature modulator digitally. The design considerations for the digital quadrature modulator have been discussed. In order to improve the spectral efficiency of mobile radio systems, QAM patterns will be required. However, the variant envelope present in QAM will necessitate the use of predistortion techniques. DSP chips can be used for complex baseband predistortion. Whether the system is dynamic or static will depend on the system FCC limitations.

ACKNOWLEDGEMENTS

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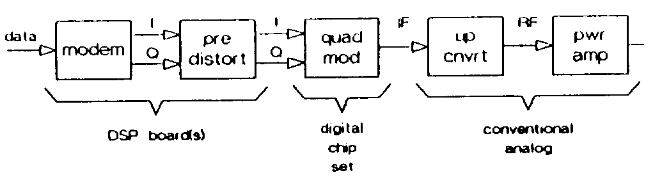

Fig. 1. MSAT Terminal with Static Predistortion

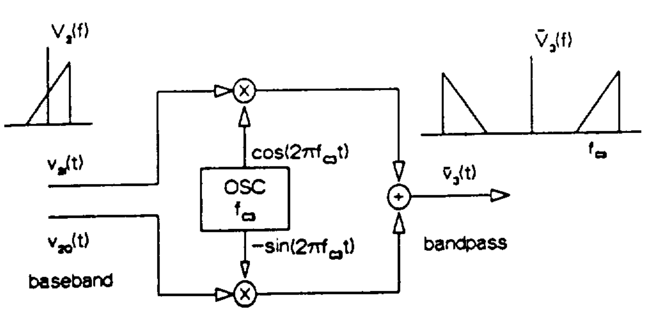

Fig. 2. Analog Quadrature Modulator

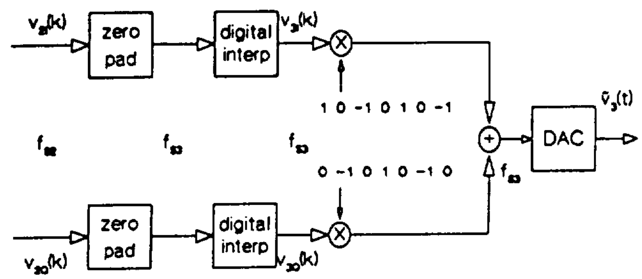

Fig. 3. Digital Quadrature Modulator

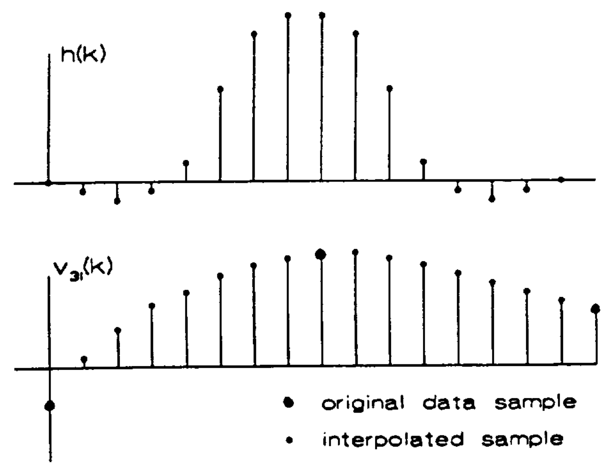

Fig. 4. Impulse Response for a 16 Point Interpolating Lowpass Filter

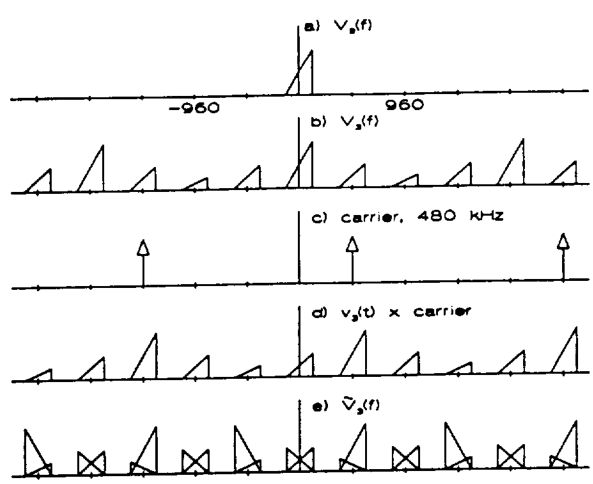

Fig. 5. Frequency Domain Spectrum for the Digital Quad. Mod.

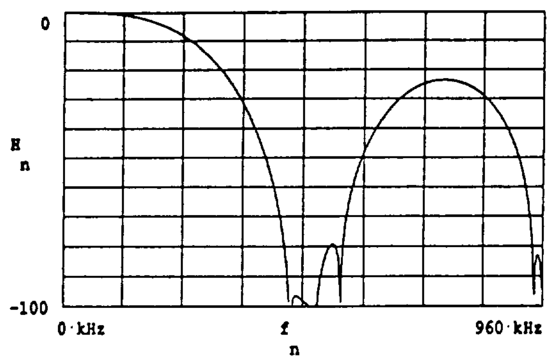

Fig. 6. 16-Point Filter for 480 KHz Samples

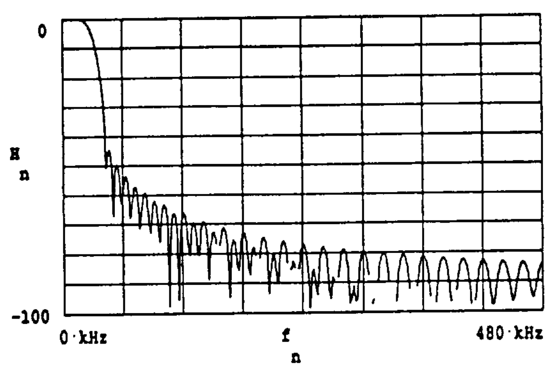

Fig. 7. 192-Point Filter for 60 KHz Samples

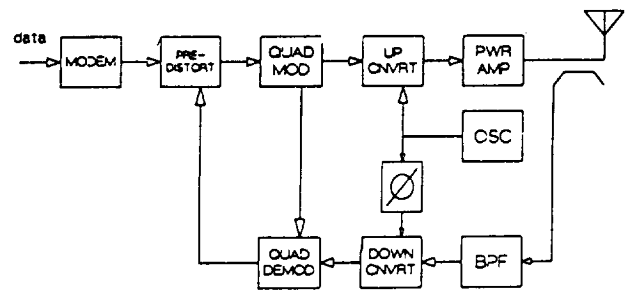

Fig. 8. Transmitter Functional Block Diagram