Channel Equalization for the IS-54 Digital Cellular System With the TMS320C5x

Application Report

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Introduction

Transmitting digital information on a radio frequency carrier is not a new concept, but it continues to attract attention because of the need to utilize the radio spectrum more efficiently through multiple access techniques that are available only with digital links. Digital signal processors (DSPs) are required by today's communications equipment to perform complicated algorithms in a limited amount of time. One such algorithm in the digital receiver is an equalizer, which is a filter that removes the distortion caused by the communications link between the transmitting antenna and the receiving antenna.

This paper's two sections discuss ways to successfully implement an equalizer for the IS-54 standard on the fixed-point TMS320C5x. The first section gives background on the digital modulation used in the IS-54 and in the radio environment that should be taken into consideration when designing an IS-54 receiver. The second section describes the design of an equalizer for the IS-54.

Design Considerations

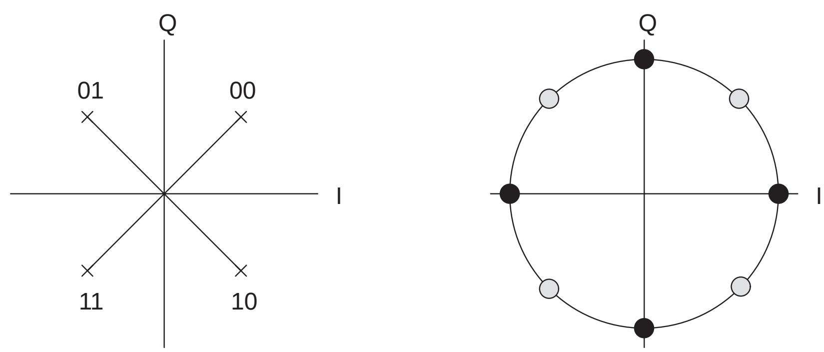
Many conditions affect a system's design. The type of digital modulation and the types of distortion and their limits influence how the receiver is structured.

Maximum-Effect Points

The IS-54 standard uses $\pi/4$ differential quaternary phase-shift keying (DQPSK) to encode a pair of bits into a phase change between two points in the complex plane. The resulting phase change is called a symbol. The points between which the change is made are known as maximum-effect points (MEPs), which are recovered by the receiver. The changes between them are investigated to decode the digital information.

Figure 1 shows the corresponding phase changes for the four dibits. The encoding process produces an eight-point constellation around the unit circle in the complex plane. Notice that these eight points can be divided into two subsets of four points. One subset is composed of the four points that are located on the axes of the complex plane. The other subset consists of the points that are in each of the four quadrants. For a given allowable phase change, if the starting point is an axis point, the end point will be a quadrant point. Similarly, if the starting point is a quadrant point, the end point will be an axis point. When a sequence of bits is encoded into a sequence of phase changes, the result is a sequence of points in the complex plane, which alternates between the subset of axis points and the subset of quadrant points.

Figure 1. π/4 DQPSK



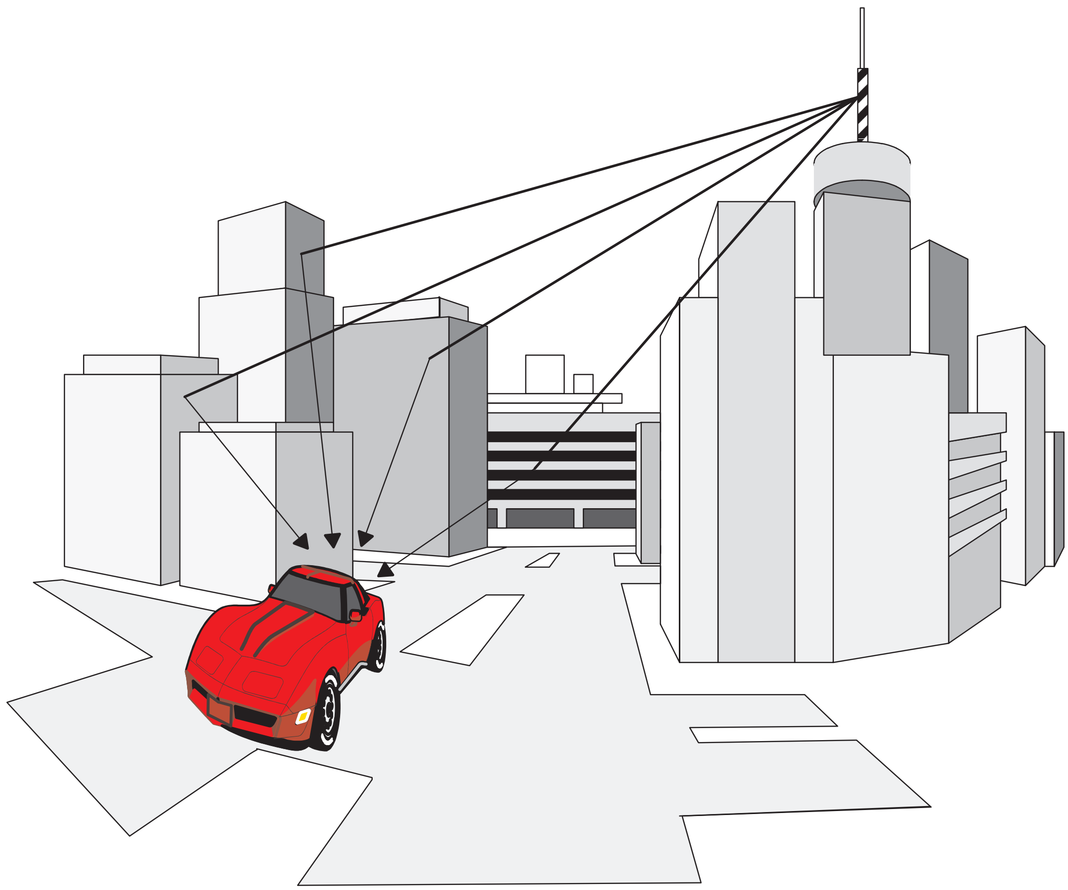
Relationship Between
Phase Changes and Dibits

Constellation of
Maximum-Effect Points

Multipath Interference

When a radio signal is transmitted, it can propagate along many paths to reach the receiver. At the receiver antenna, the received signal can be viewed as a complex sum of vectors with independent gains and phases. Multipath interference is the effect of multiple versions of a transmitted signal arriving at a radio receiver and combining in a way that produces distortion of the original signal. Figure 2 shows how multipath interference is produced by reflections from buildings or other objects.

Figure 2. Multipath Interference

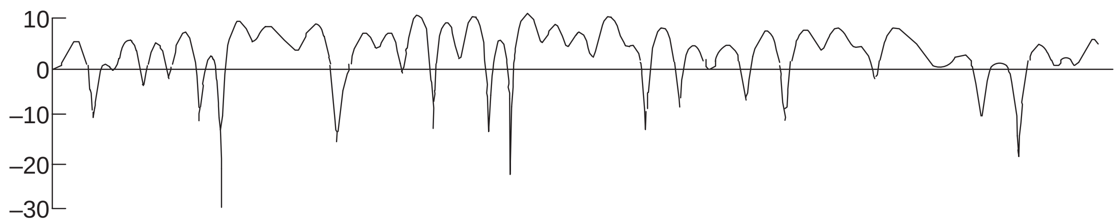


When multiple versions of the same transmitted signal arrive simultaneously, an interference pattern is formed with the various signals combining according to their amplitudes and phases. As the mobile receiver moves, the relationship between the amplitudes and phases of the signals from the various paths changes, causing undulations in both the composite amplitude and the composite phase. This effect is known as Rayleigh fading because the magnitude envelope has a Rayleigh probability distribution. Figure 3 shows magnitude and phase plots of fading for 0.64 seconds of a signal received by a mobile unit traveling at 25 MPH.

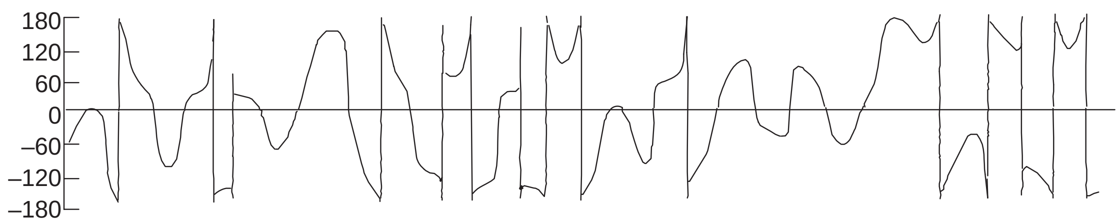
The accepted limits [2] on the amount of gain and attenuation provided by fading are +10 dB and –30 dB, respectively. Statistically, there is a 0.01% probability that the faded signal will be above 10 dB and a 99.9% probability that it will be above –30 dB.

Figure 3. Rayleigh Fading

Magnitude in dB



Phase Angle in Degrees



Intersymbol Interference

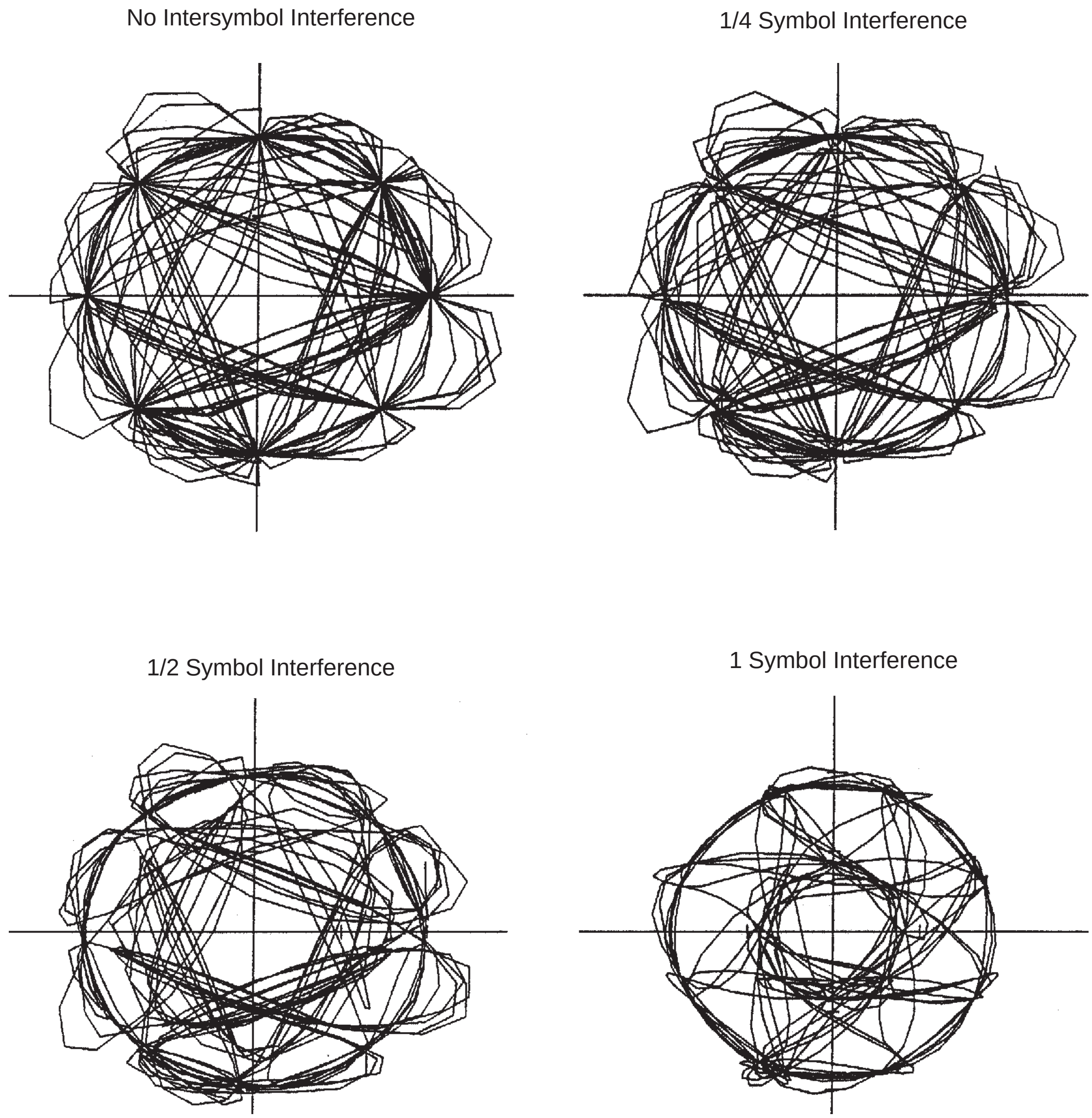
Occasionally, the arrival of some signals can be significantly delayed. This situation can result in intersymbol interference (ISI). In other words, the received signal components of previous symbols smear into later ones, thus producing distortion. In small cells, the propagation times of the different paths are nearly equal, so intersymbol interference due to multiple paths is minimal. However, in large cells, the intersymbol interference due to multiple paths can be significant. The difference in time (or symbols) between two rays' arrivals is called the *delay spread* of the channel.

Another cause of ISI is simply the bandlimited nature of the communications channel. A bandlimited channel disperses pulses going through it. This is a result of the nonideal amplitude and phase characteristics of the communications channel.

Severe ISI from one or multiple sources can render the received signal unrecoverable. For situations in which ISI is a problem, you can use an adaptive filter called an equalizer to compensate. The channel characteristics change considerably over the slot length of the IS-54 system; thus, adaptation of the equalizer coefficients is required, and the adaptive equalizer's taps must change while it is filtering the data sequence.

IS-54 defines the limit for the amount of ISI that must be compensated for by an equalizer. The IS-54 channel model was chosen to consist of a faded main ray and an independently faded delayed ray. The limit on the amount of delay is one symbol time (41.17 μs). The delayed ray can also be of equal nominal magnitude to the main ray. Figure 4 shows the effect of ISI on the π/4 DQPSK constellation, with the delayed ray three dB below the main ray.

Figure 4. Intersymbol Interference: Interferer Level –3 dBc



Equalizer Design

Set Data 12 dB Below Full Scale

Figure 3 shows that Rayleigh fading can cause the magnitude of the data to be amplified by 10 dB. To prevent the sampled representations of the I and Q baseband signals from clipping, the nominal point (0 dB) should be kept at least 10 dB below full scale. For sampled signals, the amount of dynamic range represented by each bit is 6 dB. A convenient figure to work with for a fade margin is 12 dB, which corresponds to 2 bits in sampled form.

Software AGC

To maintain maximum resolution in the presence of fading, some mechanism is required to keep the nominal value of the data at 12 dB below full scale. An RF section may contain an automatic gain control (AGC) circuit that prevents strong signals from overloading the receiver and boosts weak signals for better recovery. However, building an AGC with a large amount of dynamic range (–30 to –115 dBm) can be an issue. An alternative to a full-dynamic range hardware AGC is a combination of hardware to attenuate the large signals and software to boost signals below the desired nominal value.

An algorithm for the software AGC can be based on signal strength measurements of the incoming data. Often, the software must perform measurements for reporting the received signal strength indication (RSSI) of the cell to the cellular system. The software AGC could use the RSSI directly or in some derivative form, but a single measurement would look like:

$$m(n) = \frac{1}{N}\sum_{i=0}^{N}\left[\left(real\{r(i)\}\right)^2 + \left(imag\{r(i)\}\right)^2\right]$$

and the RSSI value would be the weighted average of individual measurements:

$$RSSI_{sw}(n) = \lambda m(n) + (1-\lambda)\, RSSI_{sw}(n-1)$$

where λ is an exponential weighting factor that is less than 1. The final RSSI report would include the amount of attenuation provided by the hardware AGC portion, which could simply be a resistive pad switched in to prevent overloading the A/Ds. $RSSI_{sw}(n)$ represents the software scale factor. Ideally

$$RSSI_{sw}(n) = 0.0625 = -12dB$$

where 12 dB is the fade margin discussed above. So the scale factor will be

$$sc\_fact(n) = 0.0625 / RSSI_{sw}(n)$$

An assembly language example is included in the source code. See the *Code Availability* section on page 187.

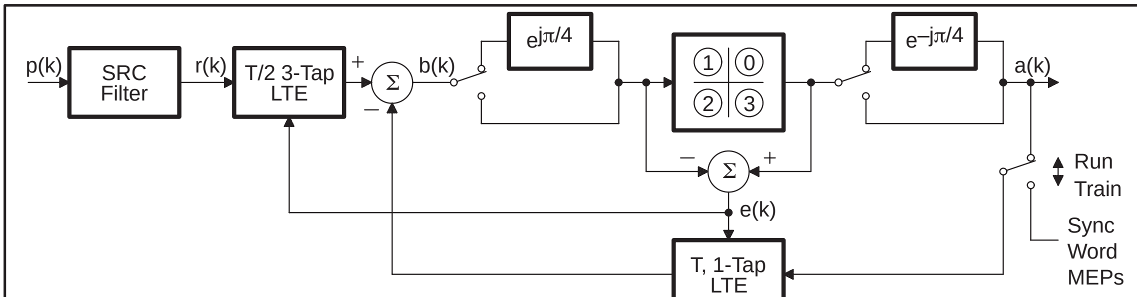
Equalization and Estimating Maximum-Effect Points

To recover the maximum-effect points in the presence of ISI, an adaptive equalizer is used in mobile stations that are compatible with IS-54. The equalizer inverts the distortion of the communications channel. Amplitude and phase distortion from fading is compensated, and components of previous symbols are removed. Figure 5 shows a block diagram of a decision-feedback equalizer (DFE). The received data is square-root raised cosine (SRC) filtered and fed into a feed-forward filter section. In this example, the feed-forward filter is a linear transversal equalizer (LTE) composed of three taps that are spaced at 1/2 symbol, or T/2. It is well-known [5] that equalizers with taps spaced in fractions of a symbol perform better than those with taps spaced at the symbol interval. Three taps spaced at T/2 provide the capability to compensate for up to one symbol of ISI, which is the upper limit specified in IS-54. The feedback filter contains a single adaptive tap. Previous decisions are filtered by the feedback filter and subtracted from the output of the feed-forward filter. This result is the estimated maximum-effect point at time k. The estimate is then fed into a data slicer, which decides which maximum-effect point is being estimated on the basis of its phase. The error vector is the difference between the estimate and the decision and drives the filter tap adaptation.

The slicer makes its decision on the basis of the phase of the estimate. Recall that there are two subsets of maximum-effect points that the encoded sequence alternates between in $\pi/4$ DQPSK. In both subsets, the points are offset by 90 degrees. In the case of the subset of quadrant points, the decision regions are trivial: determine which quadrant the estimate is in from the signs of the real and imaginary components of the estimate. In the case of the subset of axis points, the decision region boundaries are at odd multiples of π/4 in phase. However, if the estimates for the subset of axis points were rotated by π/4 in phase, the decision regions would be the quadrants of the complex plane. In Figure 5, there are two paths into and out of the slicer. The upper path is for the estimates of axis points, and the lower path is for the estimates of quadrant points.

For the equalizer to be able to track changes in the communications channel, it must first be trained to the channel's characteristics. At the beginning of every TDMA slot received by the mobile unit is a 14-symbol, 15-maximum-effect point synchronization word. The mobile unit uses this known sequence of phase changes to synchronize its receiver to the base station's transmitter. When an arbitrary starting point on the unit circle is chosen, this known sequence of phase changes can be encoded into a sequence of maximum-effect points and stored in the memory of the DSP. The stored MEPs drive the error, and the equalizer taps converge to a state in which the error is minimized; thus, the equalizer adapts to the channel's characteristics. The equalizer compensates for the phase difference, which can be completely arbitrary, between the stored MEP sequence and the received one. The taps take on whatever values are required to produce estimates of the *stored* MEPs.

Figure 5. Block Diagram of a Decision-Feedback Equalizer

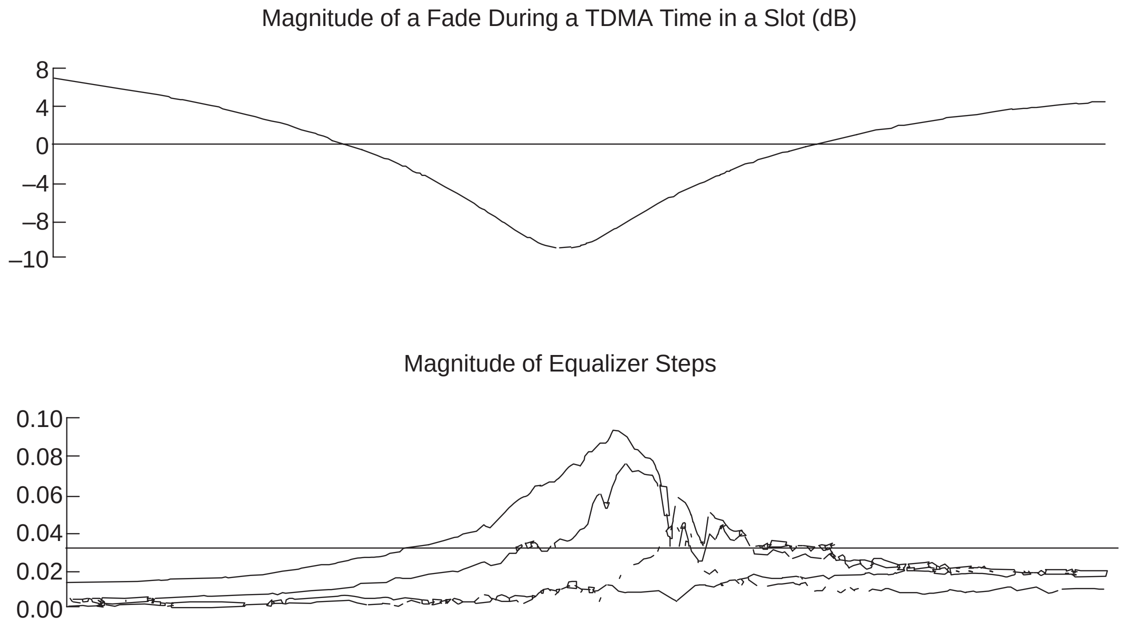


$$b(k) = \sum_{n=0}^{2} g_n(k)\ r(k-n) - f(k)\ a(k-1)$$

LTE = Linear Transverse Equalizer

The 'C5x is a fixed-point machine; to prevent the equalizer taps from exceeding 1.0, it is necessary to scale the decision points. Since the equalizer taps adapt to the inverse of the channel, an amplification by the equalizer tap compensates for attenuation of the signal. This is illustrated in Figure 6. Since fading can attenuate the received signal by 30 dB from its nominal value, the same amount of amplification could be applied by the equalizer. The desired magnitude for I and Q is 0.25, which is 12 dB below full scale. This is also the value of the decision used in the feedback path in Figure 5. The decision points must be scaled by another 30 dB (42 dB altogether). It was verified by simulation that an attenuation of the signal by 30 dB (corresponding to a constant 30-dB fade) produced a main tap magnitude equal to 1.0 when the decision points were scaled by 42 dB.

Figure 6. Equalizer Taps Responding to a Fade



Choosing an Update Algorithm

To track the changing communications channel, the adaptive equalizer uses an algorithm that updates the taps according to the error signal. Because of the requirement for tracking a fast-fading channel in a fixed-point implementation, the update algorithm should be chosen carefully.

Table 1 compares the best possible candidates. Using Table 6.8.5 in [1], a complexity comparison can be made for an equalizer with N1 = 3 feed-forward taps and N2 = 1 feedback tap. Assuming 4 DSP operations for a complex multiply and 40 DSP operations for a complex divide provides a comparative figure for the number of DSP operations required for each algorithm. These DSP operations are in the parentheses in the middle two columns of the table.

Table 1. Complexity Comparison of Update Algorithms

	Number of Complex Operations	Number of Complex Divisions	Number of Complex Multiplications	Number of DSP Operations
LMS	9	0 (0)	9 (36)	36
Fast Kalman	85	3 (120)	82 (328)	448
Conventional Kalman	58	2 (80)	56 (224)	304
Square-Root Kalman	50	4 (160)	46 (184)	344
Gradient Lattice	30	6 (240)	24 (96)	336
RLS Lattice	54	6 (240)	48 (192)	432

Of the six choices, one *must* be disqualified, and one *will* be disqualified. The LMS algorithm, although overwhelmingly simpler than the others, has insufficient convergence properties (tracking ability) for the types of channels that must be dealt with. It was included to show that the price to be paid for enough convergence is an order-of-magnitude increase in complexity. The conventional Kalman is known to have stability issues [1] and therefore should be used with caution — especially in a fixed-point implementation. For this discussion it is disqualified, as well. Of the remaining four candidates, two are clearly more complex for the desired number of taps, so the final choice is between the square-root Kalman and the gradient lattice. According to [1], the gradient lattice is a suboptimum derivative of the RLS lattice with reduced complexity and processing requirements. The square-root Kalman, however, maintains the optimal convergence properties of the conventional Kalman but uses a more stable method for updating the Kalman gain vector. It seems worthwhile to choose a slightly more complex algorithm that has significantly better convergence properties.

The list of algorithms in Table 1 is by no means comprehensive. There is a multitude of algorithms to choose from. This discussion considers only a few well-known and proven options.

Code Availability

The associated program files are available from the Texas Instruments TMS320 Bulletin Board System (BBS) at (713) 274-2323. Internet users can access the BBS via anonymous ftp at *ti.com.*

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5. Qureshi, S. U. H., "Adaptive Equalization," *Proceedings of the IEEE*, Vol. 53, September 1985, pp. 1349–1387.