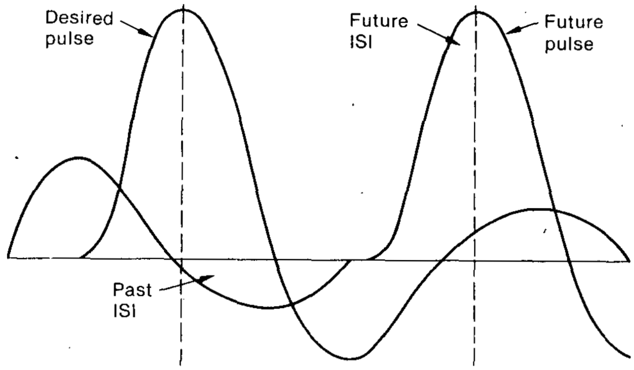

FADING CHANNEL COMMUNICATIONS

Adaptive processing can reduce the effects of fading on beyond-the-horizon digital radio links

Peter Monsen

INTRODUCTION

Two radio propagation channels for beyond-the-horizon communications, troposcatter, and HF are currently being reexamined. In the past, transmission over these radio channels had been considered unreliable due to fading effects. Recently, conversion from analog to digital transmission and the use of new adaptive signal processing techniques have offered promise of acceptable network communication quality. In addition, over-the-horizon radio provides economic and/or security advantages relative to satellite, cable, or line-of-sight terrestrial microwave links. There is renewed interest in the use of high frequency (HF) and troposcatter communications for networks carrying digital traffic.

HF radio uses frequencies in the range of 2 to 30 MHz. At these frequencies, communications beyond line-of-sight is achieved through refractive bending of the radio wave in the ionosphere from ionized layers at different elevations. In most cases more than one ionospheric "layer" causes the return of a refracted radio wave to the receiving antenna. The impulse response of such a channel exhibits a discrete multipath structure. The time between the first arrival and the last arrival is the multipath delay spread. Changes in ion density in individual layers due to solar heating cause fluctuations in each multipath return. This time varying multipath characteristic produces alternately destructive and constructive interference. The resulting fading can produce complete loss of signal as those who have heard replays of Winston Churchill's radio talks during World War II can attest.

Troposcatter radio transmission was discovered only after World War II when it was noted that microwave signals from beyond the horizon radars were much stronger than predicted from diffraction calculations over the earth's surface. A commonly held theory of this phenomenon, developed by Tatarski [1], holds that random fluctuations in the dielectric constant in the troposphere divert some small fraction of the impinging energy back to the receiver. The name tropospheric scatter or troposcatter derives from this concept of random redirection of the incident wave by the troposphere. Significant scatter returns occur from a "common volume" defined by the receiving and transmitting antenna beam patterns. Scatter returns from different points within the common volume have different path delays. Signals scattered from points separated by more than the decorrelation distance of the fluctuations in the dielectric constant are not correlated. Thus, as in the HF example, the impulse response characterizing the channel has a time-varying multipath structure with delay spread but without the discrete layers. Troposcatter systems have been widely used in military applications for beyond line-of-sight communications up to about 600 miles. The frequency range for this application extends from about 400 to 5000 MHz. Reliable communications require redundant transmission paths provided through the use of

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multiple frequencies, antennas separated in space, or scatter at two different beam angles. The several redundant paths are referred to as diversity paths, and the number of paths is termed the order of diversity.

The multipath delay spread limits the channel capacity which can be achieved in present analog systems. Only transmission bandwidths less than the reciprocal of this multipath delay can be achieved. Signals of larger bandwidths become distorted due to the multipath dispersion. In FM systems this dispersion causes intermodulation noise after detection.

With the introduction of satellite communication systems, which do not suffer from extensive multipath fading, the future of HF and troposcatter systems appeared to be limited. Economic and security factors have altered this assessment, particularly for digital transmission.

With digital signal formats, adaptive methods can exploit multipath structure to improve performance.

With digital signal formats, adaptive methods can be devised to measure the multipath structure and exploit it as an extra form of diversity to improve performance. Unlike the capacity of analog systems, the capacity of digital systems is not restricted by the multipath delay spread. From a network viewpoint, fades in tandem digital links do not have a cumulative effect because the signal can be regenerated at each node.

Adaptive troposcatter systems have been demonstrated which are efficiently able to detect digital signals perturbed by a fading channel medium while tracking the fading variations. If receiver adaptation requires significantly less signal-to-noise ratio per bit than receiver detection of the digital signal, the receiver decisions can be effectively used as the estimate of the transmitted signal to achieve what is referred to as *decision-directed* adaptation. Such systems can be operated without transmission of special pilot or reference signals for channel tracking.

In troposcatter communication, adaptive techniques have increased the digital rate capability by at least an order of magnitude.

In troposcatter communication, adaptive techniques have increased the digital rate capability by at least an order of magnitude. An adaptive equalizer modem [2],[3] developed for military troposcatter links has been successfully field tested at digital rates up to 12.6 Mbits/s in a 15 MHz channel allocation. In HF systems, adaptive signal processing techniques are now being considered with goals of digital rates on the order of 5 kbits/s in a 3 kHz channel. In both these examples, the digital symbol period is of the same order as the channel multipath delay spread.

Applicability of adaptive signal processing techniques is critically dependent on whether the rate of fading is slower than the rate of signaling. As discussed below, both HF and troposcatter radio links can be considered to be slow fading multipath channels.

SLOW-FADING MULTIPATH CHANNELS

For digital communication over beyond-the-horizon radio links, an attempt is made to maintain transmission linearity, i.e., the receiver output should be a linear superposition of the transmitter input plus channel noise. This is accomplished by operation of the power amplifier in a linear region or, with saturating power amplifiers, by using constant-envelope modulation techniques. For linear systems multipath fading can be characterized by a transfer function of the channel $H(f;\, t)$. This function is the two-dimensional random process in frequency f and time t that is observed as carrier modulation at the output of the channel when sine wave excitation at the carrier frequency is applied to the channel input. For any continuous random process, we can determine the minimum separation required to guarantee decorrelation with respect to each argument.

For the time varying transfer function $H(f;t)$, let t_d and f_d be the decorrelation separations in the time and frequency variables, respectively. If t_d is a measure of the time decorrelation in seconds, then

$$\sigma_t = \frac{1}{2\pi t_d}\ \text{Hz}$$

is a measure of the fading rate or bandwidth of the random channel. The quantity σ_t is often referred to as the *Doppler spread* because it is a measure of the width of the received spectrum when a single sine wave is transmitted through the channel. The dual relationship for the frequency decorrelation f_d in hertz suggests that a delay variable

$$\sigma_f = \frac{1}{2\pi f_d}\ \text{seconds}$$

defines the extent of the multipath delay. The quantity σ_f is often referred to as the *multipath delay spread* as it is a measure of the width of the received process in the time domain when a single impulse function is transmitted through the channel.

Typical values of these spread factors for HF and troposcatter communication are

HF	Troposcatter
$\sigma_t \sim 0.1$ Hz	$\sigma_t \sim 1$ Hz
$\sigma_f \sim 10^{-3}$ seconds	$\sigma_f \sim 10^{-7}$ seconds

where the symbol $\sim$ denotes "on the order of."

The spreads can be defined precisely as moments of spectra in a channel model [4] which assumes wide sense stationarity (WSS) in the time variable and uncorrelated scattering (US) in a multipath delay variable. This WSSUS model and the assumption of Gaussian statistics for $H(f;t)$ provide a statistical description in terms of a single two-dimensional correlation function of the random process $H(f;t)$.

This characterization has been quite useful and accurate for a variety of radio link applications. However the stationarity and Gaussian assumptions are not necessary for the utilization of adaptive signal processing techniques on these channels. What is necessary is first that sufficient time exists to "learn" the channel characteristics before they change, and second, that decorrelated portions of the frequency band be excited such that a diversity effect can be realized. These conditions are reflected in the following two relationships in terms of the previously defined channel factors, the data rate R, and the bandwidth B:

R(bits/s) $\gg \sigma_t$ (Hz) Learning Constraint

B(Hz) $\gtrsim f_d$ (Hz) Diversity Constraint

where the symbol $\gtrsim$ denotes "on the order of or greater than."

The learning constraint insures that sufficient signal-to-noise ratio (SNR) exists for reliable communication at rate R over the channel. Clearly if $R \sim \sigma_t$, the channel would change before significant energy for measurement purposes could be collected. When $R \gg \sigma_t$, the received data symbols can be viewed as the result of a channel sounding signal and appropriate processing can generate estimates of the channel character during that particular stationary epoch. The signal processing techniques in an adaptive receiver do not necessarily need to measure the channel directly in the optimization of the receiver, but the requirements on learning are approximately the same. If only information symbols are used in the sounding signal, the learning mode is referred to as *decision-directed.* When digital symbols known to both the transmitter and receiver are employed, the learning mode is called *reference-directed.* An important advantage of digital systems is that in many adaptive communications applications, adaptation of the receiver with no wasted power for sounding signals can be accomplished using the decision-directed mode. This is possible in digital systems because of the finite number of parameters or levels in the transmitted source symbols and the high likelihood that receiver decisions are correct.

Diversity in fading applications is used to provide redundant communications channels so that when some of the channels fade, communication will still be possible over the others that are not in a fade. Some of the forms of diversity employed are space using multiple antennas, angle of arrival using multiple feedhorns, polarization, frequency, and time. These diversity techniques are sometimes called *explicit diversity* because of their externally visible nature. An alternate form of diversity is termed *implicit diversity* because the channel itself provides redundancy. In order to capitalize on this implicit diversity for added protection, receiver techniques have to be employed to correctly assess and combine the redundant information. The potential for *implicit frequency diversity* arises because different parts of the frequency band fade independently. Thus, while one section of the band may be in a deep fade, the remainder can be used for reliable communications. However, if the transmitted bandwidth B is small compared to the frequency decorrelation interval f_d, the entire band will fade and no implicit diversity can result. Thus, the second constraint $B \gtrsim f_d$ must be met if an implicit diversity gain is to be realized. In diversity systems a little decorrelation between alternate signal paths can provide significant diversity gain. Thus it is not necessary for $B \gg f_d$ in order to realize implicit frequency diversity gain although the implicit diversity gain clearly increases with the ratio B/f_d. Note that the condition $R \ll B \gtrsim f_d$ does not preclude the use of implicit diversity because a bandwidth expansion technique can be used in the modulation process to spread the transmitted information over the available bandwidth B. We shall distinguish between these low-data rate and high-data rate conditions because the appropriate receiver structures take on somewhat different forms.

The learning constraint insures adequate time to measure the channel characteristics before they change. The diversity constraint insures adequate bandwidth to combat deep, frequency-selective fades.

The implicit diversity effect described here results from decorrelation in the frequency domain in a slow-fading ($R \gg \sigma_t$) application. This implicit frequency diversity can in some circumstances be supplemented by an *implicit time diversity* effect which results from decorrelation in the time domain. In fast fading applications ($R \gtrsim \sigma_t$) redundant symbols in a coding scheme can be used to provide time diversity provided the code word spans more than one fade epoch. In our slow-fading application this condition of spanning the fade epoch can be realized by interleaving the code words to provide large time gaps between successive symbols in a particular code word. The interleaving process requires the introduction of signal delay longer than the time decorrelation separation t_d. In many practical applications which require transmission of digitized speech, the required time delay is unsatisfactorily long for two-way speech communication. For these reasons there is more emphasis on implicit frequency diversity techniques in practical systems. The receiver structures to be discussed next are applicable to situations where the implicit frequency diversity applies.

ADAPTIVE RECEIVER STRUCTURES

We consider a pulse amplitude modulation system wherein the sample set $\{a_k\}$ is to be communicated over the channel using a modulation technique which forms a one-to-one correspondence between the sample a_k and the amplitude of a transmitted pulse. Independent modulation of quadrature carrier signals (i.e., $\sin 2\pi f_o t$ and $\cos 2\pi f_o t$, f_o = carrier frequency) is included in this class. An important example with optimum detection properties is quadrature phase shift keying (QPSK) which transmits the sample set $\{a_k = +1 + j\}$ by changing the sign of quadrature carrier pulses in accordance with the sign of the real and imaginary parts of the source sequence $\{a_k\}$.

A. Receivers for Channels with Negligible Intersymbol Interference

If each sample a_k can be one of M possible amplitudes ($M=4$ for QPSK), the transmitted data rate is

$$R = (\log_2 (M))/T$$

where $1/T$ is the transmitted symbol rate.

Most terrestrial over-the-horizon channel applications utilize media which are signal-to-noise ratio limited rather than bandwidth limited. In order to maximize signal detectability, only a few amplitudes are usually employed in these applications. When the symbol period T is much greater than the total width of the multipath dispersion of the channel, only a small portion of adjacent symbols interfere with the detection of a particular symbol. For the slow fading application, the diversity constraint requires that the signal bandwidth B be on the order of or larger than the frequency decorrelation interval f_d. Conditions for negligible intersymbol interference (ISI) and adaptive processing to obtain implicit diversity are then

$$T \gg 2\pi\sigma_F = \frac{1}{f_d}$$

$$B \gtrsim f_d.$$

When the number of amplitudes M is small, these conditions imply a low-data rate system relative to the available bandwidth, i.e., a bandwidth expansion system. The low-data rate condition for implicit frequency diversity can be expressed in terms of the data rate and bandwidth as

$$R \ll B \log_2 (M).$$

In the absence of intersymbol interference, it is well known [5] that the optimum detection scheme contains a noise filter and a filter matched to the received pulse shape. The optimum noise filter has a transfer function equal to the reciprocal of the noise power spectrum $K(f)$. When the additive noise at the receiver input is white, i.e., its spectrum is flat over the frequency band of interest, the noise filter can be omitted in the optimum receiver. The optimum receiver for a fixed channel transfer function $H(f)$ then contains a cascade filter with component transfer functions

$$R(f) = \underbrace{\frac{1}{K(f)}}_{\text{Noise Filter}} \underbrace{H^*(f)}_{\text{Matched Filter}}$$

where the * denotes complex conjugation. In practical applications, signal delay must be introduced in order to make these filters realizable.

In general K and H change with time and the adaptive receiver must track these variations. The tapped-delay-line (TDL) filter is an important filter structure for such channel tracking applications. The TDL filter shown in Fig. 1 consists of a tapped delay line with signal multiplications by the tap weight w_i for each tap. For a bandpass system of bandwidth B, the sampling theorem states that any linear filter can be represented by parallel TDL filters operating on each quadrature carrier component with a tap spacing of $1/B$ or less. The optimum receiver can then be realized by a cascade of two such parallel TDL quadrature filters: one with tap weights adjusted to form the noise filter, the second with tap weights adjusted to form the matched filter. Since the cascade of two bandlimited linear filters is another bandlimited filter, in some applications it is more convenient to employ one TDL to realize $R(f)$ directly. In practice, signals cannot be both time and frequency limited so that these TDL filters can only approximate the ideal solution. One advantage of the TDL filter is the convenience in adjusting the tap weight control voltage as a means of tracking the channel and noise spectrum variations.

The optimum receiver requires knowledge of the noise power spectrum $K(f)$ and the channel transfer function $H(f)$. When $K(f)$ is not flat over the band of interest, the input noise process contains correlation which is to be

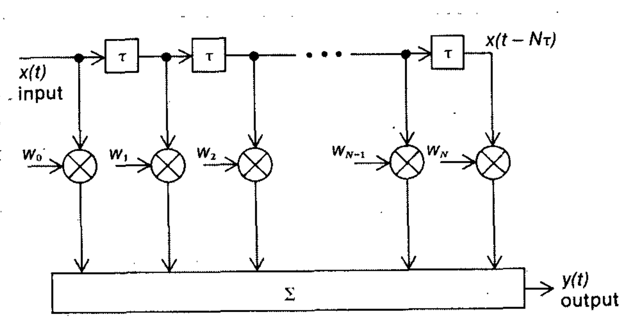

Fig. 1. Tapped-delay line (TDL) filter.

removed by the noise filter. Techniques to reduce *correlated noise effects* include: 1) prediction of future noise values and cancellation of the correlated component; 2) mean square error filtering techniques using an appropriate error criterion; and 3) noise excision techniques where a fast Fourier transform (FFT) is used to identify and excise noise peaks in the frequency domain.

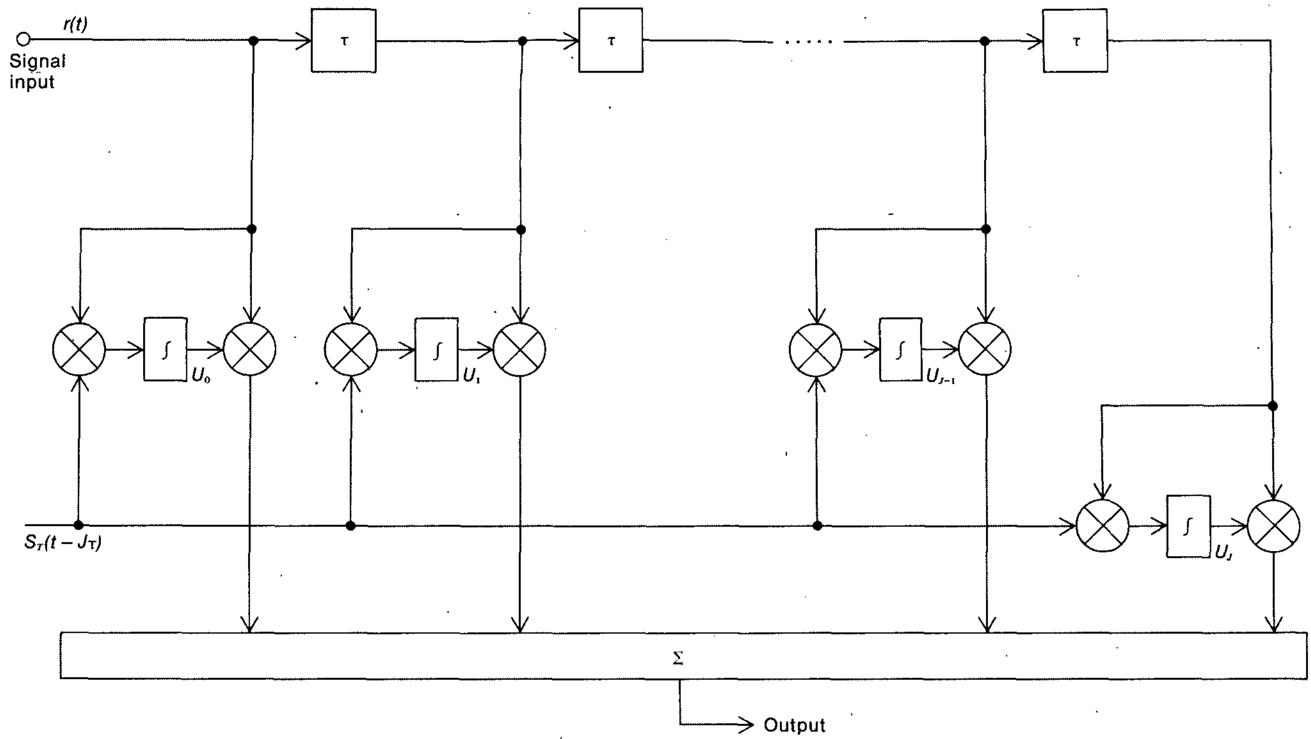

Fig. 2. RAKE filter

The problem of noise filtering is usually important in bandwidth expansion systems because of interference from other users as well as hostile jamming threats. For realization of the matched filter, a RAKE TDL filter using the concepts developed by Price and Green [6] can be used to adaptively derive an approximation to $H^*(f)$. A RAKE filter is so named because it acts to "rake" all the multipath contributions together. This can be accomplished using the TDL filter shown in Fig. 2, where the TDL weights are derived from a correlation of the tap voltages with a *common test sequence*, i.e., $S(t)$. This correlation results in estimates of the equivalent TDL channel tap values. By proper time alignment of the test sequence, the RAKE filter weights become estimates of the channel tap values but in inverse time order as required in a matched filter design. For adaptation of the RAKE filter, the test sequence may be either a known sequence multiplexed with the modulated information or it may be receiver decisions used in a decision-directed adaptation.

An alternate structure for realizing the matched filter is a recirculating delay line which forms an average of the received pulses. This structure was proposed as a means of reducing complexity in a RAKE filter design [7] for a frequency-shift-keying (FSK) system. For a pulse amplitude modulation (PAM) system, the structure would take the form shown in Fig. 3. An inverse modulation operation between the input signal and a local replica of the signal modulation is used to strip the signal modulation from the arriving signal. The recirculating delay line can then form an average of the received pulse which is used in a correlator to produce the matched filter output. This correlation filter is considerably simpler than the RAKE TDL filter shown in Fig. 2.

In both the RAKE TDL and correlation filter, an averaging process is used to generate estimates of the received signal pulse. Because this signal pulse is imbedded in receiver noise it is necessary that the measurement process realize sufficient signal-to-noise ratio. This fundamental requirement is the basis for the learning constraint

$$R(\text{bits/s}) \gg \sigma_t\ (\text{Hz})$$

introduced earlier. If the signal rate R from which adapation is being accomplished is not much greater than the channel rate of change σ_t then the channel will change before the averaging process can build up sufficient signal-to-noise ratio for an accurate measurement. This requirement limits the application of adaptive receiver techniques with implicit frequency diversity gain to slow fading applications relative to the data rate. Fortunately, many channels have fade rates on the order of a few Hertz and data requirements thousands of times larger.

The receiver for this small-ISI example has, in general, a noise filter to accentuate frequencies where noise power is weakest and a matched filter structure which coherently recombines the received signal elements to provide the implicit diversity gain. The implicit diversity can be viewed as a frequency diversity because of the decorrelation of received frequencies. The matched filter in this view is a frequency diversity combiner which combines each frequency coherently according to its received strength. Without the matched filter, incoherent combining of the received frequencies would occur and no implicit diversity effect would be realized.

An important application of this low-data rate system is found in jamming environments where excess bandwidth is used to decrease jamming vulnerability. In more benign environments, however, most communication requirements do not allow for a large bandwidth relative to the data rate and if implicit diversity is to be realized in these applications, the effect of intersymbol interference must be considered.

B. High-Data Rate Receivers

When the transmitted symbol rate is on the order of the frequency decorrelation interval of the channel, the frequencies in the transmitted pulse will undergo different gain and phase variations resulting in reception of a distorted pulse.

Although there may have been no intersymbol interference (ISI) at the transmitter, the pulse distortion from the channel medium will cause interference between adjacent samples of the received signal. In the time domain, ISI can be viewed as a smearing of the transmitted pulse by the multipath thus causing overlap between successive pulses. The condition for ISI can be expressed in the frequency domain as

$$T^{-1}(\mathrm{Hz}) \gtrsim f_d\,(\mathrm{Hz})$$

or in terms of the multipath spread

$$T(\mathrm{seconds}) \lesssim 2\pi\sigma_f\,(\mathrm{seconds}).$$

Since the bandwidth of a PAM signal is at least on the order of the symbol rate T^{-1} Hz, there is no need for bandwidth expansion under ISI conditions in order to provide signal occupancy of decorrelated portions of the frequency band for implicit diversity. However it is not

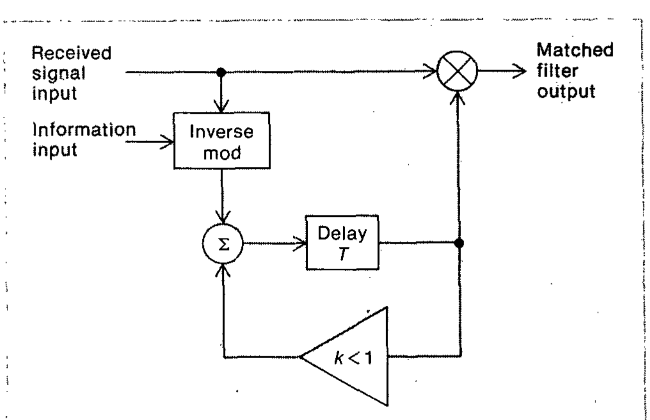

Fig. 3. Correlation filter for PAM system.

obvious whether the presence of the intersymbol interference can wipe out the available implicit diversity gain. Within the last decade it has been established that adaptive receivers can be used which cope with the intersymbol interference and in most practical cases wind up with a net implicit diversity gain. These receiver structures fall into three general classes: correlation filters with time gating, equalizers, and maximum likelihood detectors.

Adaptive receivers can cope with intersymbol interference and wind up with a net diversity gain.

1) *Correlation Filters:* These filters approximate the matched filter portion of the optimum no-ISI receiver. The correlation filter shown in Fig. 3 would fail to operate correctly when there is intersymbol interference between received pulses because the averaging process would add overlapped pulses incoherently. When the multipath spread is less than the symbol interval, this condition can be alleviated by transmitting a time gated pulse whose "off" time is approximately equal to the width of the channel multipath. The multipath causes the gated transmitted pulse to be smeared out over the entire symbol duration but with little or no intersymbol interference. The correlation filter can then be used to match the received pulse and provide implicit diversity [8]. In a configuration with both explicit and implicit diversity, moderate intersymbol interference can be tolerated because the diversity combining adds signal components coherently and ISI components incoherently. Because the off-time of the pulse can not exceed 100 percent, this approach is clearly data rate limited for fixed multipath conditions. In addition, the time gating at the transmitter results in an increased bandwidth which may be undesirable in a bandwidth-limited application. The power loss in peak power limited transmitters due to time gating can be partially offset by using two carrier frequencies with independent data modulation [9].

2) *Adaptive Equalizers:* Adaptive equalizers are linear filter systems with electronically adjustable parameters which are controlled in an attempt to compensate for intersymbol interference. Tapped delay line filters are a common choice for the equalizer structure as the tap weights provide a convenient adjustable parameter set. Adaptive equalizers have been widely employed in telephone channel applications [10] to reduce ISI effects due to channel filtering. In a fading multipath channel application, the equalizer can provide three functions simultaneously: noise filtering, matched filtering for explicit and implicit diversity, and removal of ISI. These functions are accomplished by adapting a tapped delay line equalizer (TDLE) to force error measure to a minimum. By designing the error measure to include the degrada-

tion due to correlated noise, ISI, filtering, and improper diversion combining, the TDLE will minimize their combined effects.

A linear equalizer (LE) is defined as an equalizer which linearly filters each of the N explicit diversity inputs. An improvement to the LE is realized when an additional filtering is performed upon the detected data decisions. Because it uses decisions in a feedback scheme, this equalizer is known as a decision-feedback equalizer (DFE).

The operation of a matched filter receiver, an LE, and a DFE can be compared from examination of the received pulse train example of Fig. 4. The binary modulated pulses have been smeared by the channel medium producing pulse distortion and interference from adjacent pulses. Conventional detection without multipath protection would integrate the process over a symbol period and decide a $+1$ was transmitted if the integrated voltage is positive and -1 if the voltage is negative. The pulse distortion reduces the margin again in that integration process. A matched filter correlates the received waveform with the received pulse replica thus increasing the noise margin. The intersymbol interference arises from both future and past pulses in these radio systems since the multipath contributors near the mean path delay normally have the greatest strength. This ISI can be compensated for in a linear equalizer by using properly weighted time shifted versions of the received signal to cancel future and past interferers. The DFE uses time shifted versions of the received signal only to reduce the future ISI. The past ISI is cancelled by filtering past detected symbols to produce the correct ISI voltage from these interferers. The matched filtering property in both the LE and DFE is realized by spacing the taps on the TDLE at intervals smaller than the symbol period.

The DFE is shown in Fig. 5 for an Nth order explicit diversity system. A forward filter (FF) TDLE is used for each diversity branch to reduce correlated noise effects, provide matched filtering and proper weighting for explicit diversity combining, and reduce ISI effects. After diversity combining, demodulation, and detection, the data decisions are filtered by a backward filter TDLE to eliminate intersymbol interference from previous pulses. Because the backward filter compensates for this "past" ISI, the forward filter need only compensate for "future" ISI.

An automatic gain control (AGC) amplifier is shown for each diversity branch in order to bring the fading signal into the dynamic range of the TDLE. A decision-directed error signal for adaptation of the DFE is shown as the difference between the detector input and output. Qualitatively one can see that if the DFE is well adapted this error signal should be small. Reference-directed adaptation can be accomplished by multiplexing a known bit pattern into the message stream for periodic adaptation.

When error propagation due to detector errors is ignored, the DFE has the same or smaller mean-square error than the LE for all channels [11]. The error propagation mechanism has been examined by a Markov chain analysis [12] and shown to be negligible in practical fading channel applications. Also in an Nth order diversity application, the total number of TDLE taps is generally less for the DFE than for the LE. This follows because the former uses only one backward filter after combining of the diversity channels in the forward filter.

The performance of a DFE on a fading channel can be predicted [16]-[18] using a transformation technique which converts implicit diversity into explicit diversity and which treats the ISI effects as a Gaussian interferer. As an example, the average probability of error versus the total received bit energy (E_b) relative to the noise spectral density (N_0) is shown in Fig. 6 for a quadruple diversity system. The dashed line represents the zero

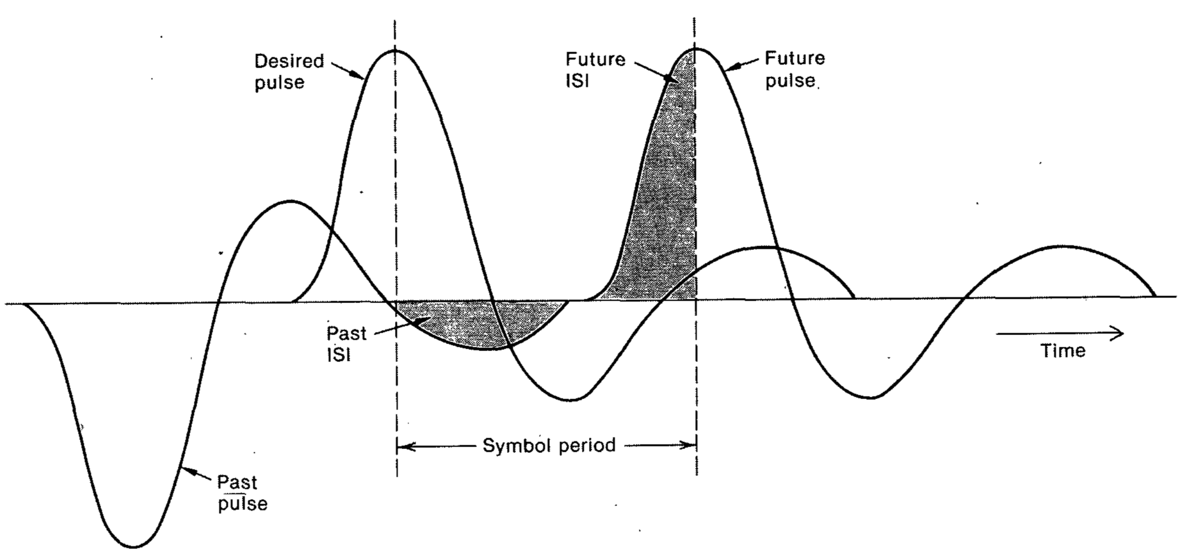

Fig. 4. Received pulse sequence after channel filtering.

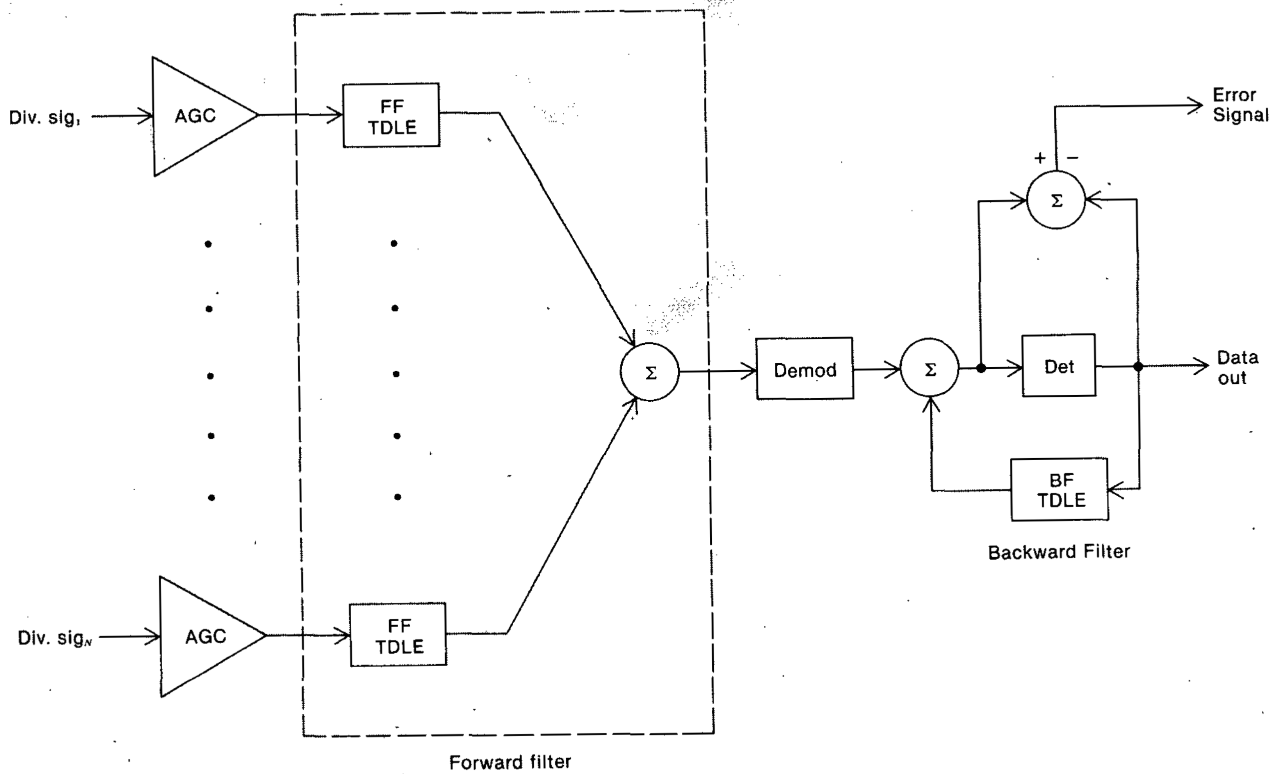

Fig. 5. Decision-feedback equalizer, *Nth*-order diversity.

multipath spread ($\sigma_f = 0$) performance and the solid lines show performance for different DFE configurations (N = number of forward filter taps and Δ = normalized tap spacing) and ISI conditions when the ratio of multipath spread to symbol period T is 0.25. The no-ISI conditions are performance bounds determining by setting the ISI components to zero. When $\sigma/T = 0.25$, performance would be to the right of the dashed line if adaptive signal processing were not employed. The equalizer is seen to remove this degradation and also provide an implicit diversity gain which is measured by the difference between the solid line $N = 3$, $\Delta = 0.5$ curve and the dashed line. The difference between the $N = 3$, $\Delta = 0.5$ curve and the next curve labeled "No ISI" is the intersymbol interference penalty. With the filter parameters $N = 3$, $\Delta = 0.5$, no technique for removing ISI can do better than this curve. The small-ISI penalty in this typical example is a strong argument for the use of the DFE versus more powerful ISI techniques. Finally the leftmost solid line approximates the very best that can be done as results show negligible improvement as the number of

A decision feedback equalizer with a modest number of taps performs almost as well as one with an infinite number.

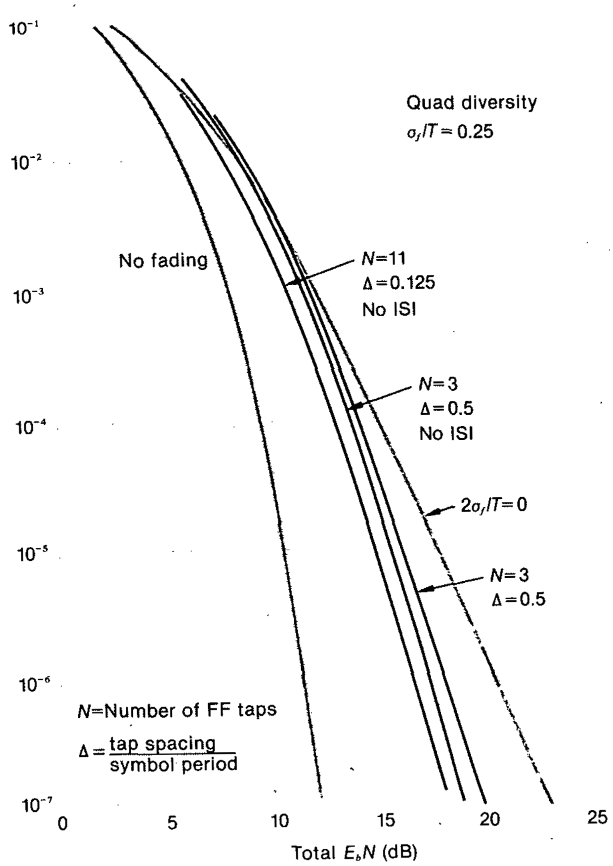

Fig. 6. DFE performance, quad diversity.

taps is increased further. The small difference exhibited shows that a DFE with only a modest number of forward filter taps performs to an ideal DFE with an infinite number of taps.

A DFE modem has been developed [3] with data rates up to 12.5 Mbits/s for application on troposcatter channels with up to four orders of diversity. This DFE modem uses only a three-tap forward filter TDLE and a three-tap backward filter TDLE. Extensive simulator and field tests [3],[16] have shown that implicit diversity gain is realized over a wide range of actual conditions while ISI effects are mostly eliminated. Thus operation at data rates near the frequency decorrelation distance is possible with no large intersymbol interference penalty. Measured results agree well with the predicted performance for which Fig. 6 is a typical example.

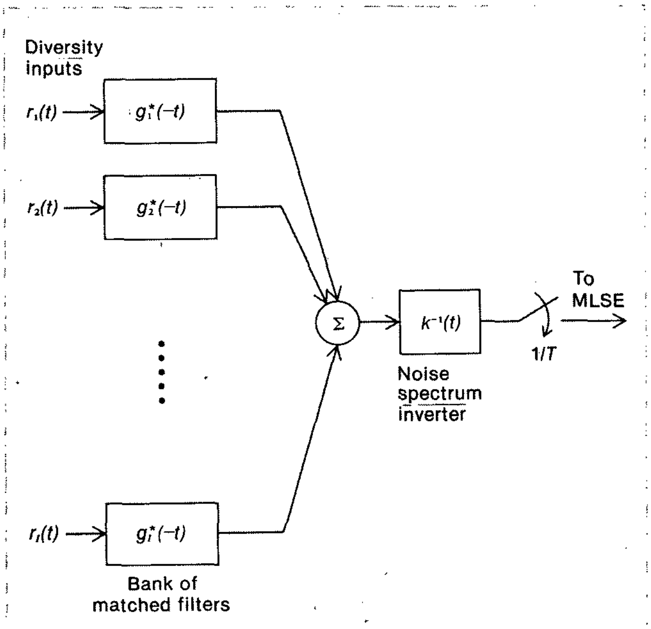

Fig. 7. Diversity combiner for MLSE receiver.

3) *Maximum Likelihood Detectors:* Since the DFE minimizes an analog detector voltage, it is unlikely that it is optimum for all channels with respect to bit error probability. By considering intersymbol interference as a conventional code defined on the real line (or complex line for bandpass channels), maximum likelihood sequence estimation algorithms have been derived [13],[14] for the PAM channel. These algorithms provide a decoding procedure for receiver decisions which minimize the probability of sequence error. A maximum likelihood sequence estimator (MLSE) receiver still requires a noise filter and matched filters for each diversity channel. After these filtering and combining operations, a trellis decoding technique is used to find the most likely transmitted sequence. Fig. 7 illustrates the filtering, combining, and sampling functions which precede the MLSE.

The MLSE algorithm works by assigning a state for each intersymbol interference combination. Because of the one-to-one correspondence between the states and the ISI, the maximum likelihood source sequence can be found by determining the trajectory of states.

If some intermediate state is known to be on the optimum path, then the maximum likelihood path originating from that state and ending in the final state will be identical to the optimal path. If at time n, each of the states has associated with it a maximum likelihood path ending in that state, it follows that sufficiently far in the past the path history will not depend on the specific final state to which it belongs. The common path history is the maximum likelihood state trajectory [13].

Since the number of ISI combinations and thus the number of states is an exponential function of the multipath spread, the MLSE algorithm has complexity which grows exponentially with multipath spread. The equalizer structure exhibits a linear growth with multipath spread. Also, the requirement for diversity combining and matched filtering in the MLSE receiver requires about the same circuit and adaptation implementation complexity as an equalizer for this requirement alone. By comparing Figs. 5 and 7 for the DFE and MLSE receiver, the systems are seen to be similar except for the replacement of the backward filter in the DFE by the decoding algorithm in the MLSE receiver. The backward filter is an L tap TDL filter whereas the MLSE decoding algorithm has computational complexity with exponential growth as a function of multipath spread. In return for this additional complexity, the MLSE receiver results in a smaller (sometimes zero) intersymbol interference penalty for channels with isolated and deep frequency selective fades. However in many applications where high orders of diversity are employed, these deep selective frequency fades do not occur frequently enough to significantly affect the average error probability. This result is illustrated in the performance curve given in Fig. 6 which showed only a small-ISI penalty for the DFE with just three taps as compared to the DFE with as many taps when there is no ISI.

NEW AREAS OF RESEARCH

Present adaptive equalizers for fading channel applications use an estimated gradient algorithm for channel tracking which can be quite slow for channels with deep selective frequency fades or in the presence of hostile electronic interference. Algorithms derived from the Kalman estimation equations have been suggested [19] as a means of realizing the full potential of adaptive tracking capability. Faster tracking would provide an impetus for HF equalization where digital data rates are not always many orders of magnitude greater than the channel rate of change. Adaptive receivers using multiple antennas in a fading multipath channel environment can provide antijamming protection from antenna nulling in addition to implicit diversity gain. Faster tracking algorithms would increase system flexibility to a wider range of jamming threats.

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