same role as that of the deleted data block in our method, and this probabilistic mapping model was called an expanded HMM [4].

For the estimation of \( m_f(r) \), although the direct use of HMM parameter training such as the forward–backward algorithm or the segmental approach is considered to be an ideal method, the computational cost is high. To reduce computation we suggest an approximated algorithm in which we use the information available during the training of original HMM parameters and no direct training is needed for the estimation of expanded HMM parameters. We describe the proposed algorithm in detail as follows:

(i) Step 1: When training the original HMM parameters, we keep separate counts for each block during the final iteration. Let \( y_i(k) \) denote the number of times that the output symbol \( s_0 \) occurred in state \( s \) on the training data block, \( B_i \).

(ii) Step 2: While the convergence condition has not been met, we iteratively update the smoothing matrix as follows. Let the elements of the smoothing matrix at iterations \( r \) be \( m_f(r) \). For each output symbol pair, we calculate the DS counts, \( c_f(s) \), which is given as

\[
c_f(s) = \sum_{i=1}^{M} \sum_{k=1}^{K} y_i(k) \cdot \frac{m_f(r) p_f(y_i | s)}{\sum_{j=1}^{N} m_f(r) p_f(y_i | j)}
\]

where \( \mathcal{S} \) denotes the set of all states. After the calculation of DS counts, by normalising the counts, we update the elements of smoothing matrix as follows:

\[
m_f(r+1) = \frac{c_f(s)}{\sum_{k=1}^{N} c_f(s)}
\]

(iii) Step 3: At every iteration, we calculate the log likelihood, \( L \), which is given by

\[
L = \sum_{i=1}^{M} \sum_{k=1}^{K} y_i(k) \log \left( \sum_{j=1}^{N} m_f(r) p_f(y_i | j) \right)
\]

and if the value of \( L \) converges, we stop the iteration.

This simplified algorithm is based on the assumption that the alignments of speech data against the HMM states are fixed and only the smoothing matrix should be estimated during the DS procedure. Further, the proposed algorithm is of estimate maximise (EM) type. Thus, any kind of smoothing matrix can be used as an initial estimate and the convergence of the iteration to local optima is guaranteed.

Simulation results: The performance of the DS method was tested in a speaker-independent isolated word recognition system. The vocabulary consists of 75 phonetically balanced Korean words which are mutually confusable. Five male speakers uttered the 75 words once to form a training set, and three male speakers who were not in trained speakers uttered the words once for testing. We used 32 Korean context independent phoneme models for the units of recognition. Each HMM has three states and represents a simple left-to-right structure. Utterances were lowpass filtered with a cutoff frequency of 4.5 kHz and digitised with a sampling rate of 10 kHz. We used 12th-order linear predictive coding (LPC) cepstral coefficients and differenced LPC cepstral coefficients as the feature vectors, and extracted them for every frame of 10 ms. Two separate codebooks were constructed such that the number of codewords was 256 for each codebook.

For each smoothing method, we interpolated the smoothed PDFs using the original PDFs using the technique of deleted interpolation. The performances of the DS methods with varying initial estimates were compared with those of the floor smoothing, distance-based smoothing and co-occurrence smoothing methods. The results are shown in Table 1, where TOP1 represents the conventional recognition rate and TOP2 the recognition rate including the words of the second highest score. The floor smoothing can be considered as a method for smoothing a PDF with the flat matrix where all the elements are equal to \( 1/N \). For the distance-based smoothing method, we used the algorithm similar to that for constant fuzzy (CF) smoothing [3]. To apply the DS method, we divided the training set into two blocks and then the iterations were run until \( L \) converged. The floor smoothing and CF smoothing methods yield recognition rates of 80.1% and 83.8%, respectively. For the co-occurrence smoothing, although the rate for TOP1 is high as that of the DS method which used the flat matrix as an initial estimate, the rate for TOP2 is much lower than those of the DS methods. According to the results, it is seen that the DS methods with various initial estimates yield better performances than those of the smoothing matrices initially given. Furthermore, the DS method with no priori information (i.e., using the flat matrix as an initial estimate) yields an even higher recognition rate than those of the CF smoothing or the co-occurrence smoothing methods.

Table 1 PERFORMANCE COMPARISON OF VARIOUS SMOOTHING METHODS IN SPEAKER-INDEPENDENT WORD RECOGNITION

<table>
<thead>
<tr>
<th>Method</th>
<th>a</th>
<th>b</th>
<th>c</th>
<th>d</th>
<th>e</th>
<th>f</th>
</tr>
</thead>
<tbody>
<tr>
<td>TOP1(%)</td>
<td>80.1</td>
<td>83.8</td>
<td>84.3</td>
<td>84.3</td>
<td>84.3</td>
<td>84.7</td>
</tr>
<tr>
<td>TOP2(%)</td>
<td>90.0</td>
<td>91.2</td>
<td>90.7</td>
<td>92.6</td>
<td>92.1</td>
<td>92.1</td>
</tr>
</tbody>
</table>

\( a \) floor smoothing, \( b \) CF smoothing, \( c \) co-occurrence smoothing, \( d \) DS initially with flat smoothing matrix, \( e \) DS initially with CF smoothing matrix, \( f \) DS initially with co-occurrence smoothing matrix.

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References

NEW FORM OF DELAYED N-PATH RECURSIVE DIGITAL FILTERS

H. K. Kwan

A new form of delay N-path recursive digital filter with transfer function expressed as a summation of N delayed parallel recursive subfilters (each with a distinct denominator) is presented. This new filter is suitable for efficient systolic realisation and does not contain uncontrollable don’t care bands. Use of the proposed filter for system modelling is demonstrated.

Introduction: There are two methods for deriving delay N-path recursive digital filters. The first method is to decompose separately the numerator and denominator transfer functions of a recursive digital filter into their respective delay N-path structures as described in References 1-3. The second

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method is to transform, mathematically, a recursive digital filter into a summation of \( L \) delayed parallel recursive digital filters (of decimation factor \( N \) of \( L \) identical denominator transfer functions as described in Reference 4. Owing to the presence of \( L \) identical denominator transfer functions, the filtering capability of such a delayed \( L \)-path recursive digital filter has been fully exploited as compared to that with \( L \) distinct denominator transfer functions. In this letter, a new form of delayed \( N \)-path recursive digital filter with the following filter format is presented:

\[
H(z^{-1}) = \sum_{i=0}^{N-1} z^{-i} H_i(z^{-N})
\]

(1)

where each of \( H_i(z^{-N}) \) for \( i = 0 \) to \( N - 1 \) is a general \( M \)-th order recursive transfer function of the following format:

\[
H_i(z^{-N}) = \frac{a_{i,0} + a_{i,1}z^{-1} + \ldots + a_{i,M}z^{-M}}{1 + b_{i,1}z^{-1} + \ldots + b_{i,M}z^{-M}}
\]

(2)

The denominators of \( H_i(z^{-N}) \) for \( i = 0 \) to \( N - 1 \) are distinct polynomials in \( z^{-N} \), as are the corresponding numerators of \( H_i(z^{-N}) \). A distinct denominator for each \( H_i(z^{-N}) \) of eqns. 1 and 2 is useful in providing a more efficient overall transfer function \( H(z^{-1}) \) for approximating arbitrary magnitude and phase specifications as compared to that with identical denominators. Uncontrollable don't care bands [5-6] exist in recursive decimating digital filters (for \( N > 2 \)) of the general form given by eqn. 1 where \( H_i(z^{-N}) \) for \( i = 0 \) to \( N - 1 \) are distinct allpass digital filters. Using the system modelling technique [7], we shall illustrate how a given digital filter can be modelled into the new form of delayed \( N \)-path recursive digital filter which does not contain any don't care bands.

\[
y(k) = \sum_{j=0}^{N-1} y_j(k)
\]

(3)

\[
y(k) = \sum_{j=0}^{N-1} x_j(k - jN) - \sum_{j=0}^{N-1} b_{j,j} y(k - jN)
\]

(4)

Given a digital filter, the coefficients of a corresponding delayed \( N \)-path recursive digital filter can be obtained by the system modelling technique as shown in Fig. 1. Define the output squared error as

\[
e^2(k) = (\hat{y}(k) - y(k))^2
\]

(5)

The changes in coefficients are defined and can be obtained as follows:

\[
\Delta a_{ij} = -\mu \frac{\partial e^2(k)}{\partial a_{ij}}
\]

for \( j = 0, 1, 2 \) and \( i = 0 \) to \( N - 1 \)

(6)

\[
\Delta b_{ij} = -\mu \frac{\partial e^2(k)}{\partial b_{ij}}
\]

for \( j = 1, 2 \) and \( i = 0 \) to \( N - 1 \)

(7)

where \( \mu \) is the coefficient learning rate parameter.

Simulation results: The new form of delayed \( N \)-path recursive digital filter has been used to model a number of recursive and nonrecursive digital filters. In the adaptive algorithm, numerator coefficients and stable denominator coefficients were updated at the end of each iteration; momentum terms were used to speed up convergence. With a sufficiently high \( N \) in the proposed filter, perfect modelling of both magnitude and phase responses of a given digital filter can be obtained. For an illustration, we describe the results obtained when the proposed filter was used to model the 34th-order linear phase FIR lowpass digital filter of Reference 8. The resultant magnitude and phase responses of both the proposed filter for \( N = 12 \) and the 34th-order FIR digital filter after 1778000 iterations with an output squared error of \( 4.39 \times 10^{-14} \) are shown in Fig. 2. The execution time was 1.54 CPU hours in which the simulations were coded in Fortran and were carried out using a 486, 50 MHz compatible PC. In Fig. 2, the sawtooth wave at the top represents the phase response. The magnitude and phase responses of both filters were close to perfect mapping in the passband and stopband even when they were enlarged. Consequently, the magnitude responses as well as the phase responses of both filters are overlapping and cannot be distinguished from each other.

Concluding remarks: A new form of delayed \( N \)-path recursive digital filter has been presented. Using the system modelling technique, results indicate that it is feasible to use such a new form of filter to model any given digital filter without uncontrollable don’t care bands. This new form of delayed \( N \)-path digital filter is efficient and is suitable for very high speed filtering applications with a sampling rate of \( (T_s + 2T_m)N(N - 1) \) (where \( N \geq 2 \); \( T_m \) and \( T_s \), respectively, represent the times for two-input real multiplication and two-input real addition) [4].

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References

FIRST LOW POLARISATION DEPENDENCE OPTICAL FSK RECEIVER BASED ON A FABRY-PEROT LASER WITH A NEAR SQUARE ACTIVE WAVEGUIDE STRUCTURE

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Indexing terms: Optical receivers, Semiconductor lasers

A low polarisation dependent optical FSK receiver based on an FP laser with one antireflection coated facet and a near square active waveguide structure is reported. The FP laser receiver presents a polarisation dependence of less than 2dB.

Introduction: Direct detection optical receivers based on multifunction semiconductor lasers have been investigated over the several past years. A number of authors have considered different semiconductor laser structures as a direct detection optical receiver. First, we mention the DMR [1] and DFB [2] lasers that can be used as an FSK receiver. Using an identical DBR laser as both transmitter and receiver an FSK transmission system up to 250MHz/s has been demonstrated. In the receiving end, the DMR laser acts as a tunable filter, FSK discriminator and photodetector. Recently, using a two electrode DFB laser, a 1.5GHz/s high bit rate transmission with 20dBm sensitivity has been reported. More recently, a highly sensitive FSK receiver has been achieved using a Fabry-Perot laser with one antireflection coated facet [3]; FSK transmission up to 1.5GHz/s with 30dBm sensitivity for 10^-7 BER was obtained.

These new active optical receivers have the advantage of combining several functions in the same device but their behaviour depends on the polarisation. However, the polarisation dependence was very strong, owing to the rectangular active waveguide structure (0.2 x 1.5 μm² cross-section), responsible for both internal gain and effective refractive index differences between TE and TM polarisation eigenstates.

In this Letter, the for the first time, a low polarisation dependence direct detection optical receiver is reported based on a Fabry-Perot laser amplifier with a near-square active waveguide structure.

Experiment: The experimental setup for the different measurements is presented in Fig. 1. At the transmitting end, a 1.56μm two-electrode DFB laser was used as a tunable FSK source. This laser presents a continuous tuning range of over 2nm with a minimum spectral linewidth of 5MHz. At the receiving end two single-electrode 350μm long BH FP lasers were used as an FSK receiver. The first Fabry-Perot laser (FP1) has a rectangular active layer with 0.2 x 1.5 μm² cross-section. The second laser (FP2) has a near square active layer* with 0.32μm thickness and a width of 0.6 μm [4]. This near square active layer has been designed to provide equal gain for TE and TM modes. A broadband multimode (four layers) antireflection coating was deposited only on the input facet of each FP laser, yielding a reflectivity of 8 x 10^-5 at 1563μm. Both FP lasers are biased above their initial threshold currents and the electrode of each laser was matched with a 47Ω resistance and a bias current connection. The DFB and FP lasers were connected by a variable optical attenuator (VA), a polariser and a 1/2 plate. An optical spectrum analyser was used to observe the spectral output of the lasers.

Device performance and system application: As mentioned in Reference 3, the FP laser with one antireflection coated facet presents multiple gain resonances or ripple. By using the linear slope regions of one of these resonances, the FP laser acts simultaneously as an optical frequency discriminator and a photodetector. The photodetection process involves detecting the forward voltage changes across the laser induced by the modulated input light.

In the near square active waveguide structure laser FP2, the gain resonances of both TE and TM polarisations overlap over a large wavelength range (>20nm).

First, the polarisation dependences of the rectangular (FP1) and near square (FP2) structures are measured. This can be carried out by measuring the variation of electrical photodetected signal, extracted from the FP laser receiver, with respect to the polarisation state of the injected light. Fig. 2 shows this polarisation dependence for a constant input power of -20dBm and at frequency f_s = 500MHz. From Fig. 2, the photodetected signal depends on the polarisation state of the injected light and when it changes from the TE to TM state, the photodetected signal decreases by 28dB for the

Fig. 1 Experimental setup

Fig. 2 Polarisation dependences of rectangular and near square FP laser structures

P_s = -20dBm, f_s = 500MHz
(i) Rectangular
(ii) Near square

* The laser was developed under France Telecom contract (DR1 90 35 129)