An Analytic Signal Approach for Transmultiplexers: Theory and Design

ENRICO DEL RE, MEMBER, IEEE, AND PIER LUIGI EMILIANI, MEMBER, IEEE

Abstract—**A new non-FFT approach to transmultiplexer implementation is presented, based on the theory of the baseband analytic signal and on its successive allocation in the FDM format by appropriate digital interpolation and filtering. The method allows wide transition bandwidths to the filters, avoids the use of any product modulator, and leads to a channel-by-channel structure.**

As a design example, the application to the 12-channel FDM primary group is considered in detail, using FIR digital filters realized by 16 bit arithmetic standard digital circuits. The analysis and the computer simulation of the system performance are finally reported, showing the actual feasibility of the method for the transmultiplexer implementation.

I. INTRODUCTION

THE rapid growth of communications has implied the need for multiplex systems able to transmit several signals simultaneously over the same communication channel. In present telecommunication networks, two standard techniques of digital multiplexing coexist: frequency-division multiplexing (FDM) for long distance interconnections, which assigns a specified frequency band to each signal to be transmitted, and time-division multiplexing (TDM) usually for shorter or local transmission trunks and switching centers, which appropriately interleaves in time the samples of all the signals [1]. The conversion between the two signal formats today is commonly achieved by a back-to-back connection of two FDM and TDM terminals, regenerating the analog baseband version of each multiplexed signal. In recent years, however, considerable attention has been devoted to the implementation of the conversion process by digital signal processing at the multiplex level (i.e., without recovering the baseband analog signal), which can be competitive or even more convenient with the decreasing cost and the increasing processing rate of digital integrated circuits [1]-[3].

At present all the approaches to an all-digital TDM-FDM translator (transmultiplexer) can be grouped in two general classes: FFT and non-FFT methods, depending, respectively, on the presence or absence of an FFT-type processor. Approaches of the first class are the system originally proposed by Bellanger *et al.* [3] and those described in [1], while an example of the second class is the system proposed by Freeny *et al.* [2], which is essentially based on a digital version of the Weaver SSB modulator.

This paper describes the design and the performance of a non-FFT transmultiplexer based on the method proposed in [4], which is basically concerned with the generation of a (complex) SSB baseband signal (analytic signal) and with its successive allocation in one of the multiplex format bands by digital interpolation and (complex) bandpass filtering. Section II describes the theory of the analytic signal approach to the transmultiplexer implementation, realized through real signal processing. A specific feature of the method is to allow relatively wide transition bandwidths to filters involved in the conversion process. Section III defines the filter characteristics and illustrates a design technique for the digital filters, which are assumed of the nonrecursive type. Finally Section IV shows, as an example, the application to the 12-channel FDM primary group. The system design is described and both theoretical and computer simulated system performances are reported.

Manuscript received June 26, 1981; revised October 21, 1981.

E. Del Re is with the Istituto di Elettronica, Università di Firenze, Florence, Italy.

P. L. Emiliani is with the Istituto di Ricerca sulle Onde Elettromagnetiche (IROE) of CNR, Florence, Italy.

II. ANALYTIC SIGNAL METHOD FOR TDM-FDM CONVERSION

To simplify we will consider for the moment the conversion procedure in the TDM-to-FDM direction. In this case L TDM signals (L even in all practical cases), each sampled at the frequency $f_s = 1/T$, have to be allocated in the SSB/FDM format in the frequency band 0 to $Lf_s/2$. The final analog FDM signal is generated after the D/A conversion of a digital signal in the FDM format obtained by an appropriate digital processing of the L input TDM signals. The digital FDM signal must be sampled at least at Lf_s. Of course, by appropriate bandpass D/A conversion the analog FDM signal can be allocated, if required, in any band $qLf_s/2$ to $(q+1)Lf_s/2$, q integer. For example [2], the 60–108 kHz FDM primary group can be obtained by extracting the required band through bandpass D/A conversion of a digital FDM signal sampled at 112 kHz.

Without loss of generality, we can suppose that the ith TDM signal $s_i(nT)$, n integer, should be allocated in the ith band of the FDM format, i.e., in the band $if_s/2$ to $(i+1)f_s/2$, $i = 1, 2, \cdots, L-2$. The index i runs from 1 to $L-2$, because in all practical cases there are at least two empty guardbands at the limits of the frequency intervals [1] (essentially for implementation convenience), so that the actual number of signals to be allocated is $L-2$.

Let $S_i(e^{j\omega T})$, $\omega = 2\pi f$, be the spectrum of the signal $s_i(nT)$, schematically shown in Fig. 1(a). For any signal $s_i(nT)$ consider the "alternate" sampled analytic signal (ASAS) defined as

$$\bar{s}_i(nT) = (-1)^{in}[s_i(nT) + j\hat{s}_i(nT)] \quad (1)$$

where $j = \sqrt{-1}$ and $\hat{}$ denotes the Hilbert transform operator.

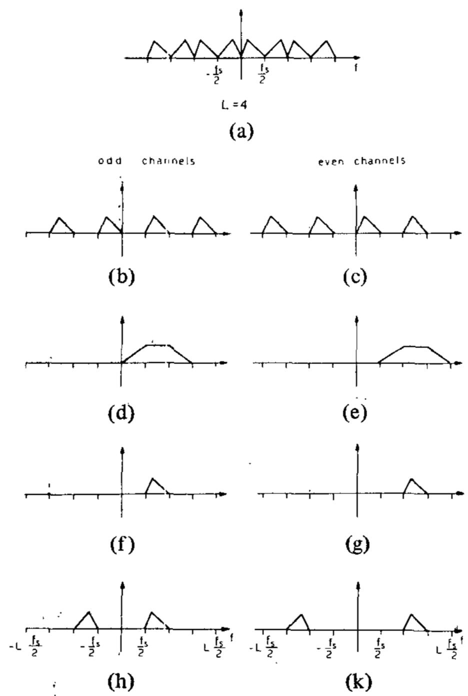

Fig. 1. Frequency plots of the TDM-FDM conversion procedure. (a) Spectrum of a TDM signal $s_i(nT)$. (b) and (c) Spectra of the sampled analytic signals obtained from $s_i(nT)$. (d) and (e) Frequency responses of the channel filters. (f) and (g) Spectra of the complex signals allocated in the FDM bands. (h) and (k) Spectra of the real SSB signals allocated in the FDM bands.

The spectrum $\bar{S}_i(e^{j\omega T})$ of $\bar{s}_i(nT)$ is shown in Fig. 1(b) for i an odd integer (odd channels) and in Fig. 1(c) for i an even integer (even channels). The signal $\bar{s}_i(nT)$ represents the usual sampled analytic signal associated with the real signal $s_i(nT)$ for i even [5], whereas for i odd its spectrum is an $f_s/2$-shifted version of the spectrum of the usual sampled analytic signal.

Formally $\bar{S}_i(e^{j\omega T})$ can be written as

$$\bar{S}_i(e^{j\omega T}) = S_i(e^{j(\omega T + i\pi)})\bar{G}_i(e^{j\omega T}) \tag{2}$$

where $\bar{G}_i(e^{j\omega T})$ is an appropriate complex filter. Ideally, its frequency response should be given by the expression

$$\bar{G}_i(e^{j\omega T}) \triangleq 1 + (-1)^i \frac{|\sin \omega T|}{\sin \omega T}. \tag{3a}$$

However, the practically limited extension (300-3400 Hz) of the spectrum of the sampled signal $s_i(nT)$ allows the filter $\bar{G}_i(e^{j\omega T})$ to have nonzero left and right transition bands. This possibility will be exploited in Section III for the actual design of $\bar{G}_i(e^{j\omega T})$. For subsequent developments it is convenient to express $\bar{G}_i(e^{j\omega T})$ as [5]

$$\bar{G}_i(e^{j\omega T}) = G_i(e^{j\omega T}) + jG_i'(e^{j\omega T}) \tag{3b}$$

through its conjugate symmetric part $G_i(e^{j\omega T})$ and its conjugate antisymmetric part $jG_i'(e^{j\omega T})$. In other words, $G_i(e^{j\omega T})$ and $G_i'(e^{j\omega T})$ are the frequency responses of the real part and the imaginary part, respectively, of the complex impulse response of the filter $\bar{G}_i(e^{j\omega T})$. It is clear that the number of different filters $\bar{G}_i(e^{j\omega T})$ is actually two: one for the odd channels and the other for the even channels. However, for notation convenience we will continue using an explicit dependence on the channel index i.

The complex signal

$$\bar{u}_i(nT/L) \triangleq u_i(nT/L) + ju_i'(nT/L) = \begin{cases} \bar{s}_i(nT/L), & n = 0, \pm L, \pm 2L, \cdots \\ 0, & \text{otherwise,} \end{cases} \tag{4}$$

obtained by inserting $L-1$ zeros between two consecutive samples of $\bar{s}_i(nT)$, is a signal sampled at Lf_s, whose spectrum $\bar{U}_i(e^{j\omega T/L})$ is [6]

$$\bar{U}_i(e^{j\omega T/L}) = \bar{S}_i(e^{j\omega T}). \tag{5}$$

Hence, it is still represented by Fig. 1(b) and (c) for the odd and even channels, respectively, but now its baseband (i.e., the range of frequencies in magnitude not greater than half the sampling frequency) extends to $Lf_s/2$. The signal $\bar{u}_i(nT/L)$ contains replicas of the baseband spectrum of the TDM signal $s_i(nT)$ correctly allocated in the frequency bands of the FDM format. The filtering of $\bar{u}_i(nT/L)$ with the complex bandpass filter $\bar{H}_i(e^{j\omega T/L})$, ideally defined as

$$\bar{H}_i(e^{j\omega T/L}) = \begin{cases} 1, & if_s/2 \leqslant f \leqslant (i+1)f_s/2 \\ \text{undefined}, & (i-1)f_s/2 \leqslant f \leqslant if_s/2 \\ \text{undefined}, & (i+1)f_s/2 \leqslant f \leqslant (i+2)f_s/2 \\ 0, & \text{elsewhere} \end{cases} \tag{6}$$

$\omega = 2\pi f$

in the frequency band 0 to Lf_s and periodic with a frequency period Lf_s, whose frequency response is sketched in Fig. 1(d) and (e) for the odd and even channels, respectively, produces a new sampled analytic signal $\bar{v}_i(nT/L)$ at the sampling rate Lf_s, which can be expressed as

$$\bar{v}_i(nT/L) \triangleq v_i(nT/L) + j\hat{v}_i(nT/L) \tag{7}$$

and whose spectrum is given by [Fig. 1(f) and (g)]

$$\bar{V}_i(e^{j\omega T/L}) = \bar{S}_i(e^{j\omega T})\bar{H}_i(e^{j\omega T/L}). \tag{8}$$

The real part $v_i(nT/L)$ gives an SSB signal correctly allocated in one of the assigned bands of the FDM format [Fig. 1(h) and (k)]. Its spectrum $V_i(e^{j\omega T/L})$ is given by [5]

$$V_i(e^{j\omega T/L}) = \tfrac{1}{2}[\bar{V}_i(e^{j\omega T/L}) + \bar{V}_i^*(e^{-j\omega T/L})] \tag{9}$$

where * respresents the complex conjugate operation. Recalling the expressions (8), (2), and (3b) and writing the complex filter (6) too in the form

$$\bar{H}_i(e^{j\omega T/L}) = H_i(e^{j\omega T/L}) + jH_i'(e^{j\omega T/L}) \tag{10}$$

through its conjugate symmetric part $H_i(e^{j\omega T/L})$ and its conjugate antisymmetric part $jH_i'(e^{j\omega T/L})$, we arrive for

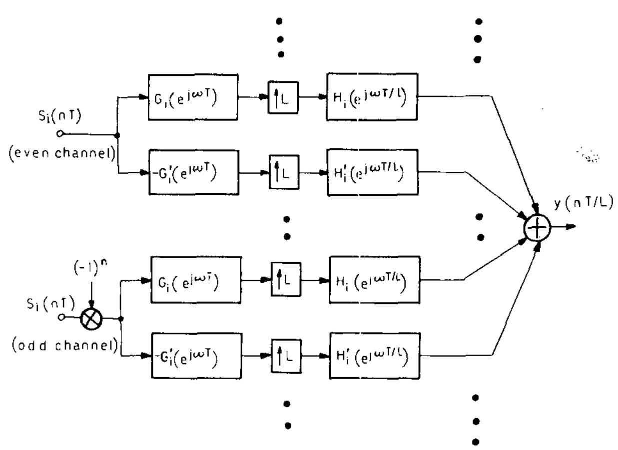

Fig. 2. Block diagram of the TDM-FDM conversion system.

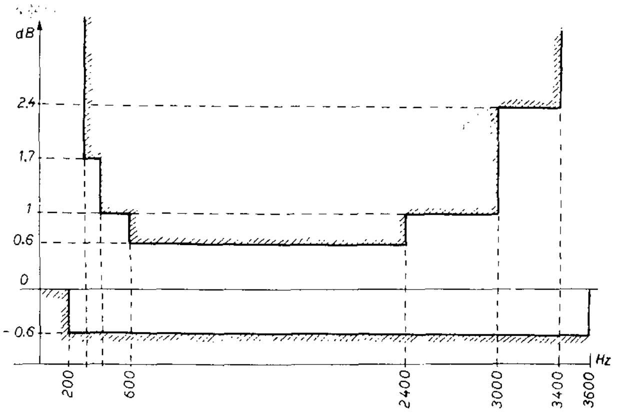

Fig. 3. Passband specifications for the tandem connection of transmultiplexer equipments.

$V_i(e^{j\omega T/L})$ at the expression

$$V_i(e^{j\omega T/L}) = S_i(e^{j(\omega T + i\pi)})[G_i(e^{j\omega T})H_i(e^{j\omega T/L}) - G_i'(e^{j\omega T})H_i'(e^{j\omega T/L})]. \tag{11}$$

Finally, the real digital FDM signal $y(nT/L)$ is obtained by summing all the $v_i(nT/L)$

$$y(nT/L) = \sum_{i=1}^{L-2} v_i(nT/L). \tag{12}$$

According to (11) and (12), the block diagram of the system configuration for the TDM-to-FDM conversion results to be that shown in Fig. 2, involving only processing of real signals by means of real filters. In particular, the conversion system does not require any product modulator. The two signal paths for each channel, according to (11), can be interpreted as performing the required SSB signal allocation by means of an in-band signal phase addition and a phase cancellation of the signals in the two adjacent bands.

III. FILTER SPECIFICATIONS AND DESIGN TECHNIQUE

The applicable specifications for transmultiplexers are to be issued by the CCITT. The specifications agreed upon by the CCITT Study Group XVIII require for the tandem connection of a TDM-FDM conversion and an FDM-TDM conversion to have for each channel the overall frequency response shown in Fig. 3, referred to the signal baseband. In the same conditions the minimum required stopband attenuation is 65 dB.

The conversion system (11) requires for each channel the design of two complex filters, $\bar{G}_i(e^{j\omega T})$ and $\bar{H}_i(e^{j\omega T/L})$: the first one to obtain the signal (1) at the low sampling rate $1/T$ and the second one to extract the replica of the baseband spectrum in the desired band. Hence, the two complex filters, which are used in cascade, must be designed so that the sum of their in-band ripples be not greater than 0.3 dB, while the stopband attenuation for each filter must be not less than 65 dB.

Finite-impulse-response (FIR) linear phase filters are a convenient choice for the channel filters $\bar{H}_i(e^{j\omega T/L})$, because they can fully exploit the presence of the padded zeros in their input signal. The functions $\bar{G}_i(e^{j\omega T})$ could be implemented by infinite-impulse-response (IIR) filters as well. However, the FIR approach was preferred because of the higher noise immunity and the absence of limit cycles and of stability problems. A related approach is that proposed in [3], where the cascade of a recursive bandpass filter and a recursive 90° phase-splitting network produces the required low-rate analytic signal.

Because of the inherent in-band signal phase addition and adjacent-band signal phase cancellation mechanism of the conversion system, the required complex filters cannot be obtained through two separate designs of their real and imaginary parts, respectively, as this design approach does not assure to a sufficient degree similar filter responses in the passband and, mainly, the transition bands.

Thus, the design of the filters $\bar{G}_i(e^{j\omega T})$ has been carried out by means of a modified version of the Parks-McClellan program [7], able to deal with complex conjugate impulse responses, according to the technique proposed in [8]. Actually, the design of only one filter is required, for example, the filter for the even channels. Indeed, the other filter, that for the odd channels, has a frequency response exactly shifted by $f_s/2$. Hence, its complex impulse response is simply obtained by multiplying the complex impulse response of the first filter by the sequence $(-1)^n$. Therefore a unique complex filter at the low sampling rate $1/T$ has been designed by the modified Parks-McClellan program.

Also, the complex channel filters $\bar{H}_i(e^{j\omega T/L})$ of Fig. 1(d) and (e) can be considered as frequency-translated versions of the equivalent low-pass prototype $\bar{H}_0(e^{j\omega T/L})$ schematically shown in Fig. 4. Ideally, the prototype $\bar{H}_0(e^{j\omega T/L})$ could be a real low-pass filter. However, the limited extension (300-3400 Hz) of the spectrum of $s_i(nT)$ allows for wider and nonequal transition bands. Thus, a more efficient filter design can be achieved by starting from a complex prototype $\bar{H}_0(e^{j\omega T/L})$. This was therefore designed using the same modified program as before and from its complex impulse response $\bar{h}_0(nT/L)$ the complex impulse responses $\bar{h}_i(nT/L)$ of the channel filters $\bar{H}_i(e^{j\omega T/L})$ have been obtained through the relation

$$\bar{h}_i(nT/L) \triangleq h_i(nT/L) + jh_i'(nT/L) = \bar{h}_0(nT/L)e^{j\pi n(i+\frac{1}{2})/L} \qquad i = 1, 2, \cdots, L-2. \tag{13}$$

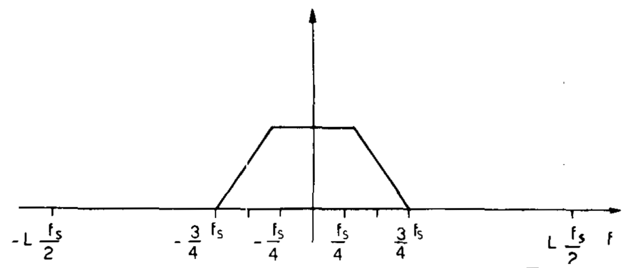

Fig. 4. Frequency response of the low-pass prototype $\bar{H}_0(e^{j\omega T/L})$.

This design procedure, in particular, guarantees identical performances for the low-rate filters $\bar{G}_i(e^{j\omega T})$ and for all the channel filters $\bar{H}_i(e^{j\omega T/L})$ through the appropriate design of a unique prototype $\bar{H}_0(e^{j\omega T/L})$.

IV. APPLICATION TO THE 12-CHANNEL PRIMARY GROUP–SYSTEM DESIGN AND PERFORMANCE

As a particular application, the case of the 12-channel FDM primary group (60–108 kHz) was considered, with an output sampling rate of the digital FDM signal equal to 112 kHz ($L = 14$). This choice was mainly motivated by the simulation convenience of the global conversion system, which will be reported in the following, and for comparison with other non-FFT approaches developed for the 12-channel primary group.

According to the specifications of Section III, the low rate filters $\bar{G}_i(e^{j\omega T})$ and the channel filters $\bar{H}_i(e^{j\omega T/L})$ must have each at least 65 dB stopband attenuation and their cascade must satisfy half the passband deviations (referred to the signal baseband) of Fig. 3. The channel filters, operating at the system highest rate, are the most critical units from a computational complexity point of view. Because of the presence of $L - 1$ padded zeros in their input signals, from an implementation point of view it is convenient to constrain their length to be of the form $qL - 1$, q integer [6]. Therefore, the minimum number for the coefficients of the low-pass complex prototype $\bar{H}_0(e^{j\omega T/L})$ satisfying this length constraint was determined. Afterwards the low-rate filter $\bar{G}_i(e^{j\omega T})$ for the even channels was designed in order to meet the overall system specifications. This filter design procedure with inifinite-precision coefficients was followed by the determination of the minimum wordlengths for the coefficients of the low-rate filters $\bar{G}_i(e^{j\omega T})$ and of all the channel filters $\bar{H}_i(e^{j\omega T/L})$ that still satisfied the overall system requirements. Table I summarizes the actual specifications and the obtained results for the designed filters. Figs. 5 and 6 show the magnitude response of $\bar{G}_i(e^{j\omega T})$ (even channels) and of $\bar{H}_1(e^{j\omega T/L})$, respectively, with coefficients rounded to the bit number of Table I.

The (linear-phase) filters $G_i(e^{j\omega T})$ and $G_i'(e^{j\omega T})$ (Fig. 2) have impulse responses with even and odd symmetry, respectively. The channel filters $H_i(e^{j\omega T/L})$ and $H_i'(e^{j\omega T/L})$ each require five multiplications per output sample. Therefore, the overall multiplication rate amounts to

$$R_M = (2 \times 5 \times 112 + 17 \times 8 + 16 \times 8) \times 10^3$$
$$= 1384 \times 10^3 \quad \text{mults/s/channel} \qquad (14)$$

which appears to be well within the computational speed of available hardware array multipliers (e.g., TRW) and could also be satisfied more cheaply by using a simple serial/parallel multiplier (e.g., realized by the AM25LS14 chips) for each channel path.

TABLE I
CHARACTERISTICS OF THE FILTERS FOR THE TDM-FDM CONVERSION SYSTEM FOR THE 60-108 kHz PRIMARY GROUP

	Passband (Hz)	Stopband (Hz)	Order	Coeff. bit number
Low rate (8 kHz) complex filter (even channels)	300 to 3600	< - 300 > 4400	32 (33 taps)	16 (even and odd channels)
High rate (112 kHz) lowpass complex prototype for the channel filters	- 1600 to 700	< - 6600 > 6300	68 (69 taps)	12 (all the channel filters)

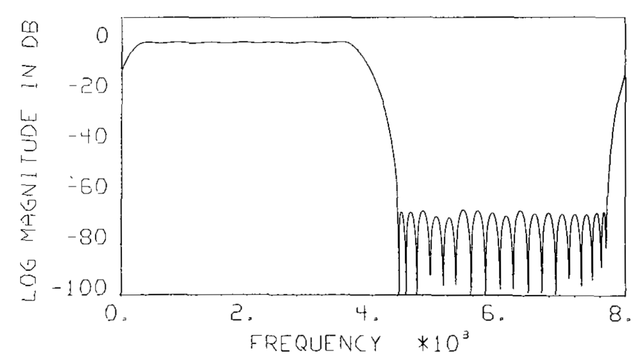

Fig. 5. Frequency response of $\bar{G}_i(e^{j\omega T})$ for the even channels with 16 bit coefficients.

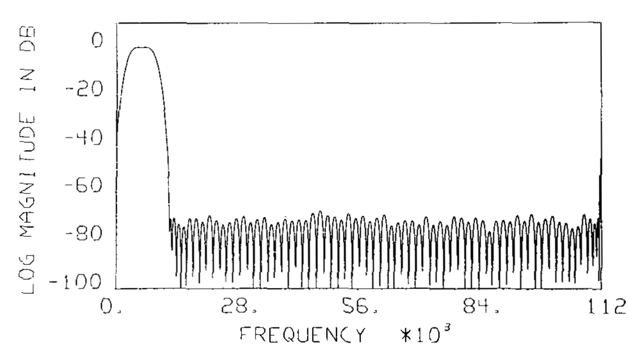

Fig. 6. Frequency response of $\bar{H}_1(e^{j\omega T/L})$ with 12 bit coefficients.

The use of FIR filters, in addition to the advantages pointed out in Section III, when implemented in direct form, allows simpler hardware solutions to improve their performances with respect to the finite-precision arithmetic noise.

The conversion system in the TDM-to-FDM direction accepts an input 8 bit A-law or μ-law coded TDM signal, which corresponds to a 13 bit linearly quantized signal. The filter coefficients are quantized to 16 bits and 12 bits for the $G_i(e^{j\omega T})$, $G_i'(e^{j\omega T})$ and the $H_i(e^{j\omega T/L})$, $H_i'(e^{j\omega T/L})$, respectively. The remaining parameter to be determined is the bit accuracy of the arithmetic operations for the FIR filterings. In the allocation of the transmultiplexer noise budget among the different noise sources, a suitable design objective is to require an output noise variance due to the rounding operations inside the filters of the same order as the variance of the quantization noise associated with the 13 bit input signal.

An estimation of the necessary wordlength for the accumulator register can be obtained from the general results for the quantization noise model applied to the chosen filter topology with the suitable overflow constraints [9]. It is known that the maximum output of a filter is equal to the sum of the absolute values of its impulse response samples times the maximum value of the input signal. However, less stringent overflow constraints can be stated if the class of input signals is restricted. For example, for a sinusoidal input signal the overflow constraints only require that the gain at the output of each filter be not greater than 1. This case will be first considered in the following noise performance evaluations.

For the assumed noise model each multiplication rounding to B bits (including sign) produces an additive noise with a variance

$$\sigma^2 = 2^{-2B}/3. \tag{15}$$

The noise variance at the output of the filters $G_i(e^{j\omega T})$ and $G_i'(e^{j\omega T})$, if realized in direct form, is therefore

$$\sigma_g^2 = 33 \times 2^{-2B}/3. \tag{16}$$

The output signal from $G_i(e^{j\omega T})$ and $G_i'(e^{j\omega T})$ is filtered by $H_i(e^{j\omega T/L})$ and $H_i'(e^{j\omega T/L})$, respectively, which require five multiplications per output sample. Hence, at the output of the filter $H_i(e^{j\omega T/L})$ the maximum noise variance amounts to

$$\sigma_1^2 = \sigma_g^2 \left\{ \max_{\substack{0 \leq k \leq 13 \\ 1 \leq i \leq 12}} \sum_{m=0}^{4} h_i^2[(14m + k)T/L] \right\} + 5 \times 2^{-2B}/3 \tag{17}$$

where the first term is the contribution of the noise from the filter $G_i(e^{j\omega T})$ and the second term is due to the multiplication rounding to B bits inside the filter $H_i(e^{j\omega T/L})$. Likewise, the maximum noise variance at the output of the filter $H_i'(e^{j\omega T/L})$ amounts to

$$\sigma_2^2 = \sigma_g^2 \left\{ \max_{\substack{0 \leq k \leq 13 \\ 1 \leq i \leq 12}} \sum_{m=0}^{4} h_i'^2[(14m + k)T/L] \right\} + 5 \times 2^{-2B}/3. \tag{18}$$

The underlying hypothesis in deriving the expressions (17) and (18) is the independence of the noise samples at the input of $H_i(e^{j\omega T/L})$ and $H_i'(e^{j\omega T/L})$ that actually contribute to the filter outputs. This assumption is justified because each noise sample comes from the contribution of different and independent rounding errors.

After the appropriate coefficient scaling, the quantities inside the brackets in (17) and (18) resulted to be slightly less than one. Hence, the maximum noise variance for each channel is from (16), (17), and (18)

$$\begin{aligned}\sigma_c^2 &= \sigma_1^2 + \sigma_2^2 \\ &\cong 2 \times (33 \times 2^{-2B}/3 + 5 \times 2^{-2B}/3) \\ &= 76 \times 2^{-2B}/3.\end{aligned} \tag{19}$$

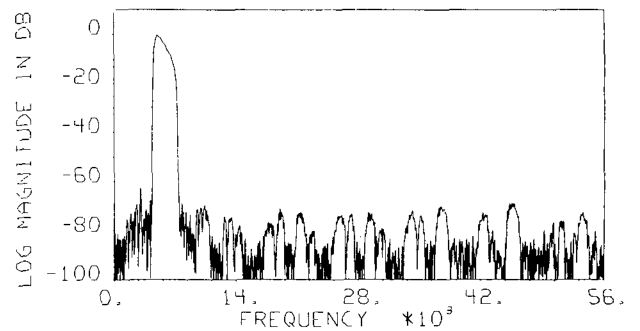

Fig. 7. Output of the TDM-FDM conversion system with only the first channel being active.

It follows that $B = 16$ is sufficient to achieve an output noise level satisfying the chosen design objective.

Furthermore, it can be observed that the variances (16) and (19) correspond to a worst case noise analysis. Indeed, by exploiting the symmetry properties of the filters $G_i(e^{j\omega T})$ and $G_i'(e^{j\omega T})$, the value of σ_g^2 in (16) could be reduced to $17 \times 2^{-2B}/3$ and $16 \times 2^{-2B}/3$, respectively. Consequently, the value of σ_c^2 in (19) would amount to $43 \times 2^{-2B}/3$, giving rise, with $B = 16$, to an arithmetic output noise level 1.73 dB below the input quantization noise level.

If the more stringent overflow constraints should be applied, the preceding analysis can be repeated simply by substituting $B - 1$ for B in (16), because in this case only the filters $G_i(e^{j\omega T})$ and $G_i'(e^{j\omega T})$ require a 1 bit scaling of their products. Hence, the same noise performance can be obtained by using a 17 bit accumulator for the filters $G_i(e^{j\omega T})$ and $G_i'(e^{j\omega T})$ and a 16 bit accumulator for the others. As a comparison, it can be observed that in [2] an input signal quantization of 15 bits is assumed instead of the 13 bit quantization suitable for a transmultiplexer implementation. This results in more severe requirements on the rounding errors that would imply an 18 bit arithmetic in the present structure, compared to the 20–22 bit arithmetic estimated in [2]. It can be pointed out that the choice of direct form structure has the advantage that an improvement of the rounding noise performance is simply obtained by increasing the length of the accumulator without modifying the remaining hardware.

The above noise variances have been also measured through a computer simulation of the overall conversion system with fixed-point arithmetic. In addition, Fig. 7 shows the system output obtained by using in the simulation the filter characteristics of Table I, when only the first channel ($i = 1$) is active. This simulation was carried out with 16 bit fixed-point arithmetic and using an input signal spectrum of a triangular form extending from 300 to 3400 Hz with the maximum at 1000 Hz.

The group delay for one conversion (e.g., TDM to FDM) is

$$D_g = 16/8000 + 34/112\,000 = 2.3 \text{ ms}. \tag{20}$$

If this delay should not meet the CCITT specifications, two suitable modifications could be considered: 1) an IIR design, or 2) a minimum-phase FIR design of the low-rate complex filters $\bar{G}_i(e^{j\omega T})$, which alone produce a delay of 2 ms.

V. CONCLUSIONS

A new non-FFT method for the transmultiplexer implementation has been described, based on the generation of a complex baseband SSB signal (analytic signal) and on its successive allocation in the FDM format bands by digital interpolation and complex bandpass filtering with relatively wide transition bandwidths. The proposed system implementation, involving only operations on real signals, uses FIR digital filters, because of their inherent high noise immunity, the absence of limit cycles and of stability problems [10], and their simpler hardware realization.

Theoretical analysis and computer simulations have shown the feasibility of the system implementation that essentially requires a 16 bit arithmetic for the filter operations for the 60-108 kHz primary group. The method allows an implementation of the transmultiplexer on a channel-by-channel basis. This solution is very attractive from the point of view of system reliability, fault recognition, and elimination, and only requires that the number of multiplications per channel be within the performance of inexpensive standard digital circuits.

The proposed system, requiring filters with asymmetrical transition bands wider than the FDM channel width, allows the design of filters with a much reduced number of coefficients with respect to methods which use a bank of conventional bandpass filters [11]. Moreover, its computational complexity results are of the same order as the digital Weaver method as applied in [2]. However, the present approach uses only FIR linear-phase filters in direct form, which, in addition to their inherent advantages, can be realized to introduce a negligible amount of noise on the processed signal by only using a double length accumulator. Furthermore, an improvement of the system computational load could be achieved by 1) using FIR minimum-phase filters [12] and 2) increasing the sampling rate in two steps.

Other approaches [13]-[15] are based on the processing of complex signals and on the use of low-pass or bandpass filters in structures not leading to a channel-by-channel implementation. In [13] complex filters with narrow transition bands are used to produce the input signal to an FFT processor. The same observation applies to the first-step filters in the tree structure described in [14]. A tree structure is also proposed in [15], where recursive filters with wider transition bands are used.

The present approach also applies to the 12-252 kHz and 312-552 kHz 60-channel supergroups through an appropriate new design of the complex filters $\bar{G}_i(e^{j\omega T})$ and $\bar{H}_i(e^{j\omega T/L})$. The design and performance of the 60-channel transmultiplexer are presently under evaluation.

Finally, it must be noted that the method applies equally well to the FDM-to-TDM conversion, simply by reversing the block diagram of Fig. 2, and substituting adding points with branch ones and sampling rate increase operations with sampling rate decrease ones [16].

ACKNOWLEDGMENT

The authors wish to thank Dr. D. F. Maffucci for his valuable cooperation in the filter design and system simulation.

REFERENCES

[1] Special Issue on Digital Signal Processing, *IEEE Trans. Commun.*, vol. COM-26, May 1978.
[2] S. L. Freeny, R. B. Kieburtz, K. V. Mina, and S. K. Tewksbury, "System analysis of a TDM-FDM translator/digital A-type channel bank," *IEEE Trans. Commun. Technol.*, vol. COM-19, Dec. 1971.
[3] M. G. Bellanger and J. L. Daguet, "TDM-FDM transmultiplexer: Digital polyphase and FFT," *IEEE Trans. Commun.*, vol. COM-22, Sept. 1974.
[4] E. Del Re, "A new approach to transmultiplexer implementation," in *Proc. Int. Conf. Commun.*, Denver, CO, June 1981.
[5] A. V. Oppenheim and R. W. Schafer, *Digital Signal Processing*. Englewood Cliffs, NJ: Prentice-Hall, 1975.
[6] R. W. Schafer and L. R. Rabiner, "A digital signal processing approach to interpolation," *Proc. IEEE*, vol. 61, June 1973.
[7] J. H. McClellan, T. W. Parks, and L. R. Rabiner, "A computer program for designing optimum linear phase digital filters," *IEEE Trans. Audio Electroacoust.*, vol. AU-21, Dec. 1973.
[8] M. T. McCallig, "Design of digital FIR filters with complex conjugate pulse responses," *IEEE Trans. Circuits Syst.*, vol. CAS-25, Dec. 1978.
[9] V. Cappellini, A. G. Constantinides, and P. L. Emiliani, *Digital Filters and Their Applications*. London, England: Academic, 1978.
[10] A. Fettweis and K. Meerkötter, "On parasitic oscillations in digital filters under looped conditions," *IEEE Trans. Circuits Syst.*, vol. CAS-24, Sept. 1977.
[11] C. F. Kurth, K. J. Bures, and P. R. Gagnon, "Per-channel, memory-oriented, transmultiplexer with logarithmic processing—Architecture and simulation," in *Proc. Int. Conf. Commun.*, Denver, CO, June 1981.
[12] R. Boite and H. Leich, "A new procedure for the design of high order miminum phase FIR digital or CCD filters," *Signal Processing*, vol. 3, Apr. 1981.
[13] R. Maruta and A. Tomozawa, "An improved method for digital SSB-FDM modulation and demodulation," *IEEE Trans. Commun.*, vol. COM-26, May 1978.
[14] T. Tsuda, S. Morita, and Y. Fujii, "Digital TDM-FDM translator with multistage structure," *IEEE Trans. Commun.*, vol. COM-26, May 1978.
[15] A. G. Constantinides and R. A. Valenzuela, "A new modular low complexity transmultiplexer design," in *Proc. Int. Conf. Commun.*, Denver, CO, June 1981.
[16] T. A. C. M. Claasen and W. F. G. Mecklenbräuker, "On the transposition of linear time-varying discrete-time networks and its application to multirate digital systems," *Philips J. Res.*, vol. 33, nos. 1/2, 1978.

★

Enrico Del Re (M'78) was born in Florence, Italy, in 1947. He received the Dr. Ing. degree in electronic engineering from the University of Pisa, Pisa, Italy, in 1971.

After military service in the Italian Air Force, involved in the maintenance and supervision of telecommunication and air traffic control equipment, from 1973 to 1975 he worked for the Officine Galileo, engaged in the analysis and design of control circuits and systems for space applications. In 1975 he joined the Istituto di Elettronica of the University of Florence, Florence, first with a research fellowship and since 1978 as Associate Professor of Radio-Aids to Navigation. His main research interests are digital signal processing, digital communications, and radar signal processing.

Dr. Del Re is a member of the Italian Electrotechnical and Electronic Association (AEI) and EURASIP.

★

Pier Luigi Emiliani (M'71) received the degree in physics from the University of Florence, Florence, Italy, in 1968.

Since 1968 he has been with the National Research Council at the Istituto di Ricerca sulle Onde Elettromagnetiche, Florence, where he has been involved in the theory and applications of data processing techniques. From 1970 to 1971 he was with the Italian Navy, working on communications problems. His main interests are now in digital filtering and image processing with biomedical and communications applications. He is also involved in the application of speech processing to aids for the blind. He is coauthor of the book, *Digital Filters and Their Applications*.

Mr. Emiliani is a member of EURASIP.