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A Design Method for High Audio Quality FM Multiplex Encoders

A.J. Oliviera

Acutron Electroacustica Lda
Sacavem, Portugal

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A. J. Oliveira

Acutron Electroacustica Lda
Urb. Quinta Nova Imp. 1 Lote 134 2685 Sacavem Portugal
Telex 63697 ACTRON P Phone (1)9414087 Fax (1)9417509

FM multiplex encoder design has up to recently relied on either the analog multiplier (matrix) technique or the switching method. Both suffer from significant audio performance drawbacks. An alternative digital sampling method is presented, which closely emulates the operation of an ideal multiplier. An extension of this technique to decoder design is also considered.

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0 Introduction:

Main drawbacks in conventional FM multiplex encoder designs are respectively noise, temperature instability and poor stereo separation in the case of designs using an analog multiplier. Strong intermodulation and interference are their counterparts in the case of units using a switching modulator, needing for heavy brickwall-type filtering. An alternative design method is presented, relying on digital sampling methods, which emulate closely the operation of a high linearity analog multiplier without its drawbacks. Close attention to the effects of filtering and sampling rates is given in order to minimize transient signal degradation, in spite of tight transmission media requirements. Extensive measurements are presented in order to verify predicted results. Finally an extension of this technique to decoder design is also considered.

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1 Background:

Monophonic FM broadcasts

Earliest commercial FM broadcast practice was monophonic, and conformed in what concerns transmission characteristics to standards like the FCC in the US, or the CCIR in Europe. Either standard imposed a minimum 200kHz spacing between adjacent channels, and to further guard against adjacent channel interference an upper limit was also laid on maximum instantaneous frequency deviation. This latter limit was set at 75kHz. If we assume the maximum transmitted audio frequency to be f (in Hz) and the worst case of having all the in-band power concentrated at this frequency this leads to the following simplified calculation:

$$BW=2(\theta f_m+f_m) \qquad (1)$$

where: BW -Transmission bandwidth
θf_m-Carrier frequency deviation
f_m -Modulating frequency

Rearranging equation (1) in order to extract f readily gives:

$$f_m=BW/2-\theta f_m \qquad (2)$$

Considering BW=200kHz and θf_m=75kHz conducts to f_m=25kHz, which of course is well above the audible spectrum. Guarding against interchannel interference however dictates a more realistic upper limit of about 20kHz. Fortunately power density of typical audio signals falls with increasing frequency, helping to prevent adjacent channel interference. This effect is somewhat counteracted by the standard preemphasis network associated with FM broadcasts (75µs in the US, 50µs in Europe). Conjunction of the two factors result for practical purposes in the fact that equation (1) accurately predicts the interference phenomena.

Stereophonic FM broadcasts

From the very first transmission trials, in which two separate carriers were used to vehicle left and right channel information, two facts were readily apparent:

1-Transmission over a single carrier of both channels was necessary to keep the spectrum from being crowded and to avoid unnecessary receiver complexity, keeping the bandwidth the same as in the case of a normal monophonic transmission.
2-Transmitted stereo signal should be compatible with monophonic receivers, meaning that both left and right channel information should be recovered in a receiver without stereophonic decoding capabilities.

Final adopted system employed the following scheme:
Left and right signals, which from now on will be designated respectively by L(t) and R(t) and are both functions of time t, are added and subtracted to result in two (matrix) signals which are L(t)+R(t) and L(t)-R(t). These signals are bandwidth limited to 15kHz by means of more or less complex filters. An oscillator at the transmitter issues a signal at a frequency of f_p=19kHz with an amplitude we shall designate as P. The signal just described is designated as pilot signal, and is further applied to a doubler of some sort which in turn produces a sine subcarrier of $f_c=2f_p$=38kHz. This 38kHz signal is amplitude-modulated by the difference L(t)-R(t) signal by means of a balanced modulator which suppresses the carrier, leaving only the sidebands. Finally this signal is summed back to the L(t)+R(t) signal and to the pilot signal in order to become a so-called composite or multiplex (MPX) signal. This signal shall therefore be designated here by M(t). The essential of such process is summarized in fig.1.
Further signals may be added to the basic MPX signal in order to vehicle additional information, such as a $f_{\blacksquare}$=67kHz FM modulated carrier with an amplitude C conveying either an additional band-limited audio channel(SCA) or asyncronous serial digital information (ARI, RDS) signal A(t). In the end, the composite or MPX signal modulates the main RF (radio frequency) carrier and the product is fed to the transmitting antenna.

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2 Transmission media requirements

As in stereophonic broadcasting the goal to achieve is the compression of all the signals in order for them to occupy the same spectral space as in the case of a regular monophonic signal, stringent requirements shall be imposed on the baseband components. We will now look to the power spectral density of a typical multiplex signal:
This signal is given as explained in the previous section, by:

$$M(t)=L(t)+R(t)+(L(t)-R(t))\sin 2\pi f_c t+P\sin 2\pi f_p t+C\sin(2\pi f_s+A(t))t+\Sigma_0^n B_n \sin 2\pi f_n \qquad (3)$$

where: S - Amplitude of the additional information channel modulated carrier
A(t) - modulating audio signal of the additional information channel

Also it was included in equation (3) an additional term corresponding to other subcarriers of frequency f_n and amplitude B_n, normally used for control purposes.
We shall now and for the purpose of the present discussion disregard as not relevant the last two terms of equation (3), resulting in simplified equation (4):

$$M(t)=(L(t)+R(t))+(L(t)-R(t))\sin 2\pi f_c t+P\sin 2\pi f_p t \qquad (4)$$

Corresponding power spectrum is obtained by Fourier-transforming equation above. Simplifying and calling the sum signal S and the difference D we get:

$$M(t)=S(t)+D(t)\sin w_c t+P\sin w_p t \qquad (5)$$

and assuming the Fourier transform of M(t) to be $F_M(f)$, of S(t) $F_S(f)$ and finally of D(t) $F_D(f)$:

$$F_M(f)=F_S(f)+F_D(f)((f_c+f)+(f_c-f))+P(f_p) \qquad (6)$$

As we can easily induce from equation (6) we have the original baseband spectrum of the sum signal plus two transposed half-amplitude difference signal sidebands around the suppressed carrier f_c plus a pilot frequency tone f_p. This power spectrum is pictured in fig. 2, assuming uniform spectral density inside the band-limited 20-15000Hz baseband signals for the sake of simplicity.
It might be intuitively deducted from both the previous equations and fig. 2 that the baseband signal must be reduced in order to accommodate the remaining transmitted signals. It is not however the case, because of a characteristic of the composite stereo signal known as interleaving [1]. This results directly from the fact that there is in normal conditions a much more strong sum signal than the difference one. To gain some insight on this assumption we will consider the three extreme cases of equation (4):

1- Left and right channel signals are identical in terms of amplitude and phase:
In this case equation (4) reduces to (ignoring for now the pilot):

$$M(t)=2L(t) \qquad (7)$$

2-Left and right channel signals are identical in amplitude but of opposite phase:

$$M(t)=2L(t)\sin 2\pi f_c t \qquad (8)$$

It is evident that as the sine term oscillates between +1 and -1 M(t) will oscillate between +2L and -2L, giving a maximum of 2L.
3-One of the channels is set to zero:

$$M(t)=L(t)+L(t)\sin 2\pi f_c t \qquad (9)$$

In this case maximum is again 2L(t).
Above considerations lead us to see that in order to provide sufficient headroom to allow for a peak deviation identical to a monophonic signal it is only necessary to cut down the amplitudes of the sum and difference signals by one half.
In the prior analysis we neglected the importance of the pilot signal. This signal is unlikely to have a phase correlation to the main MPX signal for most of the time, so it is necessary to provide room for it in the final mix. In a compromise between easiness of processing of the pilot signal within the receiver and at the same time minimize signal to noise degradation caused by the necessity to lower the main multiplex signal in order to accommodate the pilot, this one was fixed at about 10% of the total amplitude of the MPX signal. Chosen amplitude requires a 10% lowering in the main signal and so introduces only a degradation of less than 1 dB upon the signal to noise ratio.

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3 Encoder types

Analog matrix encoders

Historically the first encoders applied faithfully the technique pictured in fig. 1. They are known as matrix encoders, due to the sum/difference matrix network at their inputs. With time the pilot oscillator/doubler configuration was replaced by a digital quartz controlled oscillator running at 38kHz or at a multiple frequency and using dividers to generate that signal, a subsequent digital divide by two circuit to generate the pilot signal and a filter bank to filter out unwanted harmonics from both the 38kHz carrier and the pilot signal. An analog multiplier was employed for the balanced modulator and very careful summing of the sum signal and modulated difference signal was necessary in the output section in order to insure acceptable stereo separation. Careful adjustment of the pilot phase was also critical in order not to degrade this last parameter. These designs are still in use and feature a very quiet operation in what harmonic and spurious products generation is concerned. However, they suffer from separation degradation caused by analog multiplier temperature drift, from distortion generated within the multiplier itself and caused by poor linearity, and from noise due to small signal levels present again within the multiplier[4].
If carefully built, they don't need too much filtering on inputs to prevent aliasing with the 38kHz carrier, and none on the output, resulting in very good frequency and phase response characteristics considering the tightness of the transmission media. In a direct result from that transient behavior is

normally excellent, due to the use of standard Bessel or Butterworth filters.

Switching encoders

This type of encoders use a modulator consisting basically of an electronic switch, connecting alternately the two stereo channels to the output at a 38kHz rate, such as pictured in fig. 3. Extensive filtering is needed at the output to prevent carrier harmonics from reaching the output. This is normally done with the aid of Cauer filters, profiting of the nulls present in their transfer functions, which are order-dependant. Resulting phase response is poor and high frequency components are not even in the best designs sufficiently attenuated, giving rise to odd aliasing effects that degrade harmonic and intermodulation distortion performance. Noise performance is also degraded in inverse proportion with filtering amount. As frequency response, phase and high frequency stereo separation degrade with increased filtering, this topology is a continuing headache for the designer. Nevertheless, switching encoders offer, if well designed, complete independence from temperature effects and do not need critical summing of signals in the output section to provide good channel separation. Some designs exist that use the basic topology of the analog matrix encoder and use a time division multiplier simply in place of the analog multiplier. This approach needs of course careful signal summing as in the case of the matrix encoder. Others avoid a part of filtering by applying an amount of the third harmonic of the carrier (114kHz) to the output, with inverse phase. If adjustment of the amplitude of this 114kHz signal and the time symmetry of the 38kHz carrier (responsible for the second harmonic at 76kHz) is carefully made is it is possible to reduce filtering to a point where performance becomes very good. However, a small change in circuit layout or in oscillator duty cycle because of temperature negates these advantages. In all cases, extreme care must be taken in the switch section so that channels do not overlap because of different switching times, degrading separation, or that switching itself is too slow, introducing distortion.

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4 The digital sampling concept

Previous considerations about current encoder types readily revealed the necessity of finding a new circuit topology that features the advantages of both analog matrix and switching encoders without their drawbacks. As the problems inherent to carrier harmonic production on the time division encoder are hard to solve we looked forward to use the old matrix technology. To do that it was necessary to solve the temperature drift, distortion and noise problems encountered with this design. As they all reside in the analog multiplier, the logic alternative was to implement a new multiplier. Fortunately one of the signals to be multiplied (the carrier) is totally predictable and time invariant. Being this, chosen approach was to sample the difference L(t)-R(t) signal at a multiple of the 38kHz carrier, at equal time intervals and at a frequency that is very high in relation to the Nyquist rate, imposing to each successive time sample a multiplicative factor following an algorithm synchronous with a perfectly sinusoidal 38kHz waveform. Output filter

requirements in this case can be relaxed to the extent of the relation of the sampling frequency to the 38kHz carrier, allowing the use of complex pole pair denominator low pass filtering without the time and phase problems inherent to elliptic (also known as Cauer) filters. The filter order can be relaxed in the same previously stated proportion, avoiding unnecessary circuit complexity.

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5 Sampling frequency versus output filtering

In order not to degrade the signal it is necessary to pass 15kHz and its image frequencies, 23 and 53kHz with the least possible attenuation (say 0.5dB at the most). It is also necessary to cut the sampling signal itself and its harmonics to a level at least 80dB below the nominal output signal, to avoid interference. If for example a 12X oversampling (456kHz) is used we must use a 4th order Chebyshev, a 5th order Butterworth or a 10th order Bessel filter to do the job, listed by increasing order of complexity [2]. When transient response is taken into account, the 5th order Butterworth seems to be a reasonable compromise.
Theoretical attenuation for this type of filter is expressed by:

$$AT=20\log(1+(f/f_c)^{2n})^{1/2} \qquad (9)$$

where: n is the filter order.
f is the frequency at which attenuation is to be determined
f_c is the filter -3dB frequency

Now the problem resumes to apply equation (9) to the frequency where the attenuation is desired to be 0.5dB (53kHz), and to the frequency where it is desired to be 80dB (456kHz), terms which conduct to the following equations:

$$20\log(1+(f_1/f_c)^{2n})^{1/2}=0.5 \qquad (10)$$

$$20\log(1+(f_2/f_c)^{2n})^{1/2}=80 \qquad (11)$$

$$f_2/f_1=8.6 \qquad (12)$$

Solving equations (10) and (11) in terms of extracting $f_1$and f_2:

$$f_1=f_c(10^{5/100}-1)^{1/2n} \qquad (13)$$

$$f_2=f_c(10^{8}-1)^{1/2n} \qquad (14)$$

Simultaneous solution of the three equations (13), (14), and (12) in terms of n results in:

$$n=\ln((10^{8}-1)/(10^{5/100}-1))/2\ln(8.6) \qquad (14)$$

which solved gives n=4.77, a good approximation of our first assumption that a 5th order Butterworth filter would be a good choice.

Now we must determine what, with a 5th order filter, shall be the -3dB frequency to give -0.5dB at 53kHz. The problem resumes to solve equation (13) in terms of f_c. This gives:

$$f_c=65.4kHz \qquad (15)$$

Now is the time to check attenuation at the 456kHz sampling frequency. This can be done by solving equation (9) with n set to 5 and f set to 456kHz. Result is:

$$AT=84.3dB \qquad (16)$$

which represents quite good performance taking into account the fact that the sampling frequency is already sufficiently removed from the useful signal so as not to alias with it producing products falling after decoding in the audio band. This filtering is therefore only needed both to avoid interchannel interference and to prevent interference with SCA transmission at all costs. Final measured filter response is shown in fig.4. Attenuation of this filter alone at the 67kHz SCA subcarrier frequency is about 3.6dB, needing accessory filtering to lower it further.

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6 Sampling multiplier design

Now we will turn our attention to the sampling multiplier design task. Assuming a sampling frequency of 12 times the carrier frequency we will have twelve steps per cycle in order to regenerate the multiplicative process. As each cycle is 360o each time slot corresponds to a phase angle of 30o. Two issues are now possible to do the job:
1- Use the L(t) and R(t) signals directly to produce the MPX signal.
2- Use the L(t)-R(t) signal to produce the suppressed carrier difference signal
As it is now obvious issue 1 must transform the two channel signals in an output signal conforming to equation (4). In the second case sampling multiplier must issue a signal of the form:

$$D(t)=(L(t)-R(t))\sin 2\pi f_c t \qquad (17)$$

corresponding to the second term of equation (4).
Let's consider this case first because it is the only one with direct application to the matrix encoder canonic structure:
First we will construct a table of the sine function sampled at 30o interval starting at 0o point.

Table 1

Angle	sine
0	0
30	0.5
60	0.866
90	1
120	0.866
150	0.5
180	0
210	-0.5
240	-0.866
270	-1
300	0.866
330	-0.5

It seem thus possible, after looking into table 1 to use a 6 input multiplexer to do the job. Basic circuit of one of the possible implementations is pictured in fig. 5. In every time interval, a multiplicative step corresponding to the sine value of a 38kHz sine wave is discretely multiplied by the L(t)-R(t) signal. The corresponding multiplied 38kHz waveform is pictured in fig.6.
If we turn now to the other possibility (construct the multiplex signal directly from the L(t) and R(t) signals, and thinking on a configuration similar to fig. 5 we are led to see that the multiplicative function is no longer an approximation of a sine function. Let's rewrite equation (4) in terms of L(t) and R(t), ignoring the third term corresponding to the pilot:

$$M(t)=L(t)(1+\sin 2\pi f_c t)+R(t)(1-\sin 2\pi f_c t) \qquad (18)$$

This time we have two independent multiplicative factors which force us to modify the multiplier design. Fortunately they have a constant sum and they are shifted by 180o one from the other. We will construct a table again to gain some knowledge of the possible implementation:

Table 2

Angle	1+sine	1-sine
0	1	1
30	1.5	0.5
60	1.866	0.134
90	2	0
120	1.866	0.134
150	1.5	0.5
180	1	1
210	0.5	1.5
240	0.134	1.866
270	0	2
300	0.134	1.866
330	0.5	1.5

Again one of the possible implementations of this multiplier is shown in fig. 7. A similar but substantially reworked and condensed version of it was the object of a prototype but was abandoned because of being extremely critical on layout in order to insure a correct sum-difference signal relationship. It may be possible to implement this modulator in a single chip, minimizing inter-connection capacitances responsible for the inaccuracies, because all necessary components are resistors and semiconductors. This could result in a minimal parts count MPX encoder/decoder featuring superior performance.

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7 Input anti-aliasing filtering

Concepts stated in section 5 about output filtering are also valid here. The sampling frequency here is so much removed that it will not dictate in any way the filter design criteria. The problem we now have is to avoid that input signal frequencies in excess of 19kHz translate in the multiplier in lower sideband frequencies well into the audio band. We are helped to counteract this effect by the fact that typical audio signal power density falls with increasing frequency, but have as we already saw the preemphasis network working against us. Logical solution would consist on a Cauer brickwall filter. However these filters have the inconvenience of having a constant stopband attenuation which to be made high demands a very high order filter, with the associated transient response problems. We will examine requirements to see if they can be solved without resorting to brute-force solutions.
Problem region is the audible band top at 15kHz and above. In order for a signal in the baseband to translate back after multiplied with the 38kHz subcarrier to an audio signal again, its frequency must be higher than 23kHz. Being so, it is convenient to have a rapid attenuation for signals exceeding 15kHz, and one proportional to frequency because "garbage" of this type (appearing as intermodulation) is most audible in the midrange than in the extreme high. Still another problem is the extreme proximity of the pilot tone (19kHz) relative to the bandwidth extreme (15kHz). This is perhaps the worst problem encountered in the transmission process, because we must keep audio signals from modulating the 19kHz pilot tone. Another annoyance possibility is the proximity of the 53kHz sideband endpoint from the 67kHz subcarrier frequency used in SCA transmission, that we already mentioned in the section where we discussed output filtering. This last signal has the additional drawback of being of much lower amplitude than the main MPX spectrum.
In the case of the 19kHz subcarrier there is no solution using a Butterworth filter as before. The constant stopband of a Cauer would negate (unless a very high order filter was used) a good 38kHz rejection, and one increasing with frequency. The solution found to this dilemma was to use the same 5th order Butterworth filter as before, transposed in frequency so as to give 0.5 dB loss at 15kHz so giving about 24dB of attenuation at 38kHz. This filter will be placed at the inputs. Additionally we will cascade with the output an high Q notch filter, so as to obtain about -35dB attenuation at 19kHz. This guarantees that both the pilot signal and the suppressed subcarrier will not be disturbed by an out of range baseband signal. Frequency and phase response of the notch filter is shown in fig. 8, of input filter in fig. 9 and total cascade input/output filter in fig. 10. SCA problems were attenuated through the cascade of the input and output filters, because the response of the first transposes in frequency to a

filter identical to the one already described for use in the output section, giving thus a 10th order behavior, as we can see in fig. 10. This induced frequency response difference between sum and difference channels will unfortunately create us some phase and in result of that, some separation problems. A possible issue to that problem could be the introduction of an additional 5th order Butterworth in the way of the sum signal, just before the main summing network, but that would complicate the final design even further. Solution was to choose carefully the constants of the notch filter so as to compensate the phase differences of the added difference channel filtering, at least up to 15kHz. Of course if we are really not concerned with complexity we could introduce an additional 5th order high pass filter at the output, working with a 10.6kHz -3dB frequency, but the added advantage would not worth the effort.

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8 Pilot signal generation

The 19kHz pilot signal is critical to the decoding process, for two main reasons. First it must absolutely be free from the 2nd harmonic which would if present degrade the suppressed subcarrier rejection, and second because its phase must be steady as a rock in respect to the 38kHz suppressed subcarrier itself, to allow for a correct channel separation in the final decoder matrix adding/subtracting process. Of course not only the second harmonic of the pilot signal is dangerous; the third is at 57kHz and can interact with a control subcarrier placed at that frequency. Taking the exposed considerations into account, we generated the pilot signal using the a similar process to the one already used to obtain the suppressed 38kHz subcarrier, and taking care in the 24X sampling sine algorithm to synchronize pilot waveform so as it crosses zero each time the MPX signal has a positive slope, this for an MPX signal corresponding to positive L and R=-L. This synchronization process will depend on the kind of control logic used to generate the signal.
As we chose to place the 19kHz trap filter on the output some form of filtering not associated with the main output filter (placed before the trap) was necessary to produce a pure pilot signal. Used filter was also a fifth order Butterworth, giving 0.1dB of attenuation at 19kHz, and -120dB of rejection at the 456kHz sampling frequency. Attenuation of this filter is 14dB at the second harmonic and 31dB at the third. As the purity of the synthesized waveform is already very high (a spectrum analysis of it before filtering is presented in fig. 11) the net result is a total THD+N of less than 0.1% for the pilot signal, measured over a 500kHz bandwidth.
The final step was to guarantee phase tracking between pilot and main MPX signal. This was obtained through the use of a simple first-order phase equalizer with variable time constants, using precision components in order to minimize temperature drift. An alternative could consist on an automatic phase tracking circuit, but this was not considered necessary for the present purpose, because 0.1o phase tracking re 38kHz was very easy to obtain.
Lots of misconceptions have been stated about the importance of phase tracking to a perfect stereo separation. I recall one advertising sheet which stated that 0.6o would result on a separation degradation to 55dB. Let us check this in more detail:

We will start by rewriting equation (9), which describes the multiplex signal in the absence of the right channel considering the signal expressed by this equation is being recovered in a decoder by multiplying the second (difference) term again with the $\sin 2\pi f_c t$ factor, only with this last term replaced with $\sin(2\pi f_c t+\delta)$, where δ is the phase error. We will call this recovered signal $K(t)$.

$$K(t)=L(t)\sin(2\pi f_c t)\sin(2\pi f_c t+\delta) \qquad (19)$$

Equation (19) can be rewritten successively as:

$$K(t)=L(t)(\cos\delta-\cos(2\pi f_c t+\delta))/2 \qquad (20)$$

$$K(t)=L(t)(\cos\delta-\cos(2\pi f_c t)\cos\delta+\sin(2\pi f_c t)\sin\delta)/2 \qquad (21)$$

Equation (21) consists of three terms, one representing the recovered baseband signal, which is proportional to $\cos\delta$, and two other terms of much higher frequency, which are sidebands of $2f_c$. These last two will be eliminated by filtering. When the recovered difference signal K is again matrixed with the baseband signal $L(t)$ we get:

$$L(t)=L(t)/2+L(t)(\cos\delta)/2 \qquad (22)$$

and:

$$R(t)=L(t)/2-L(t)(\cos\delta)/2 \qquad (23)$$

Separation between left and right channels is then, expressed in dB:

$$S_{LR}=20\log((1-\cos\delta)/(1+\cos\delta)) \qquad (24)$$

Let us now apply the example given earlier of 6o phase difference re 38kHz. Applying equation (24) S_{LR}=-51dB, representing a net 4dB difference from the figure quoted in the above mentioned add. All the previous considerations and attention paid to the noise levels present in FM transmission conduct us to aim for a more than enough stereo separation of about 70dB (crosstalk signal in the order of the noise level). Rewriting equation (24) as:

$$\delta=\arccos(1-\log^{-1}(S_{LR}/20))/(1+\log^{-1}(S_{LR}/20)) \qquad (24)$$

Maximum δ will thus be ±2o. Obtained limit of 0.1o will thus correspond to a theoretical separation of 122dB, which will obviously never be obtained in practice, because of other factors disregarded in the present analysis, such as phase differences induced by different filter paths.

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9 Preemphasis

Standard preemphasis (50µS in Europe, 75µS in the US) has to be applied, in order to ensure correct frequency response. In order to avoid side-effects at high frequency which would counteract filter action, it was decided to cut it off at a frequency as low as possible without degrading the audio spectrum. Preemphasis

will be applied using the feedback network around an operational amplifier, as can be seen in fig. 12. We will start by writing the basic transfer function of the network:

$$H(s)=(1+sC(R_1+R_2))/(1+sR_1C) \qquad (25)$$

Clearly this function has a pole and a zero, respectively at $s=-R_1C$ and $s=-1/(R_1+R_2)C$. The magnitude of the transfer function was obtained after some algebra and is:

$$|H(w)|=((1+w^2C^2R_1(R_1+R_2))^2+w^2C^2R_2^2)^{1/2}/(1+w^2C^2R_1^2) \qquad (26)$$

Obviously the zero time constant $(R_1+R_2)C$ will equal one of the specified preemphasis time constants. We will choose accordingly the other so $|H(w)|$ after multiplication with the complementary deemphasis function will not deviate more than 0.5dB from a straight line.
Complementary deemphasis function is (see fig. 12):

$$G(s)=1/(1+sC(R_1+R_2)) \qquad (27)$$

so the product $G(s)H(s)$ is:

$$G(s)H(s)=1/(1+sCR_1) \qquad (28)$$

Corresponding magnitude is:

$$|G(w)H(w)|=1/(1+w^2C^2R_1^2)^{1/2} \qquad (29)$$

And rewriting in terms of $f=w/2\pi$, calling $F(f)=G(f)H(f)$:

$$f=(|F(f)|^{-2}-1)^{1/2}/2\pi R_1C \qquad (30)$$

Or equivalently expressing $|F(f)|$ in dB and calling it F:

$$f=(10^{-F/10}-1)^{1/2}/2\pi R_1C \qquad (31)$$

Applying equations above to the European case, requiring $(R_1+R_2)C$=50µS and setting f=15kHz and F=-0.5 gives:

$$R_1C=(10^{-F/10}-1)^{1/2}/2\pi f=3.7\mu S \qquad (32)$$

(R2+R1)C=50µS implies R2/R1=(50-3.7)/3.7=12.5

Of course in the US case R1/R2=19.3. Final plot of preemphasis function can be seen in fig. 13, along with deviation from linearity after deemphasis in fig. 14.

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10 Input, output stages and power supply

Input stages were built using a variable gain differential amplifier using a single operational amplifier, as described in [3]. This design was capable of more than 50dB CMMR measured over the whole audio band. Input and output passive filters were incorporated to prevent interference. Input filter has its -3dB point at 200kHz, well outside the audio spectrum, and output filter was placed accordingly at 1MHz. Output configuration is electronically balanced for minimum distortion. Input and output stages are powered from separate symmetrical regulated power supplies and logic/switching circuitry, which was implemented using CMOS technology, is powered from its own symmetrical regulated power supply. All indicators and level meters are powered by their own unregulated supplies. Power supply configuration adopted avoids crosscoupling between stages, which is critical at working frequencies of the order used here (456kHz). Also very careful layout had to be used to avoid performance degradation due to mutual inductance and capacitive coupling. Final adopted block diagram is shown in fig. 15. An SCA input was also provided for expansion.

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11 Measurements

Several parameters were measured at the output of a top-grade commercial hi-fi tuner, (details available upon request) after using the output of the prototype encoder to feed an FM transmitter. These measurements, in spite of not reflecting the encoder quality, are representative of the standard quality obtained in the present state of technology at the listening end, taking all of the transmission chain into account. These measurements have nothing to do with the ones usually presented by manufacturers of such units, that reflect characteristics of the encoder itself, that never turn into real advantage unless the decoder uses a similar decoding process. Thus presented measurements strictly answer the radio station manager question : Will it improve my station's signal clarity?. Figs. 16-20 show the most important measurements taken. We can thus conclude that in a typical transmission/reception chain, (not in laboratory conditions) the present encoder can move the present performance figures to the following values:

70dB unweighted signal to noise ratio

Frequency response flat within 1dB from 20Hz to 15kHz

Phase accuracy within ±1o from 20Hz to 15kHz

THD+N < 0.1% in band, rising to 0.5% at 15kHz

CCIF and SMPTE IMD <0.2%

Channel crosstalk < 45dB at 1KHz, degrading to 15dB at 20Hz and to 25dB at 15kHz

Total dynamic range from 20 to 15kHz, with maximum output level measured at 0.5% distortion: 75dB

We were extremely pleased by the fact that most of the figures we measured, especially signal to noise ratio, THD+N, CCIF and SMPTE IMD were the same with or without the encoder inserted in the transmitting chain. This last fact is a proof that this configuration exceeds by far presently available transmitter and receiver performance capabilities, making it the most transparent link in the chain. Fig. 19 will draw our attention as being the comparison of the dynamic capabilities of a present state of the art encoder, which performance is shown in dashed lines, with the present topology ones, shown in solid traces. It is important to

state that the upper curves, which represent the maximum output level at constant 0.5 distortion, are limited in the case of the present configuration by transmitter VCO non-linearity, conducting to figures about 10dB lower than expected.

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12 Conclusion

A method of FM multiplex encoder/decoder design for high quality audio applications has been described, using standard circuitry. It can provide increased performance in relation to conventional designs, both analog and switching. It has great advantages in what transient behavior is concerned, always one of the most negative aspects of FM stereo transmission, and is exceptionally quiet and transparent in actual operation. Technology used lends itself to manufacture of most of the present encoder in integrated circuit form, possibility the author hopes will be exploited in the future. A similar decoder is very easy to implement using the same techniques described, and one is being built now in order to investigate what would be the perceptible performance advantage if all of the encoding/decoding process was implemented this way.

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13 References

[1] H. Taub and D. Schilling, "Principles of communication systems" Mc Graw-Hill, 1971

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[3] G. Tobey, J. Graeme and L. Huelsman, "Operational amplifiers-Design and applications" Burr-Brown/Mc Graw-Hill, 1971

[4] A. J. Oliveira "A feedforward side-chain Limiter/Compressor/De-esser with improved flexibility" Journal of the Audio Engineering Society, Volume 37, Number 4, pages 226-240

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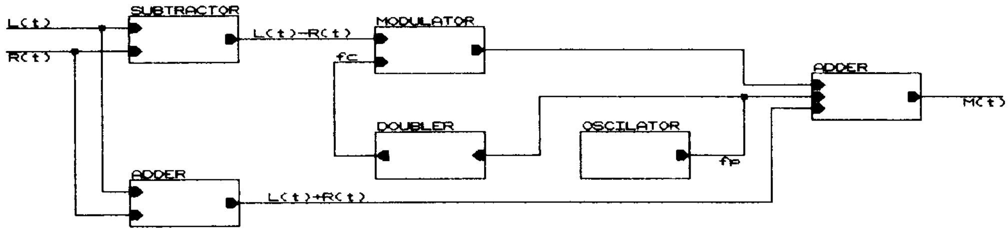
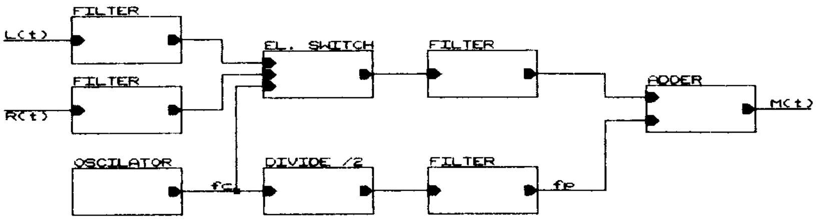
14 Acknowledgements

I sincerely wish to thank Acutron broadcast customers for useful suggestions, which helped to make the SME2D encoder, where the described topology was applied, a successful and useful product.



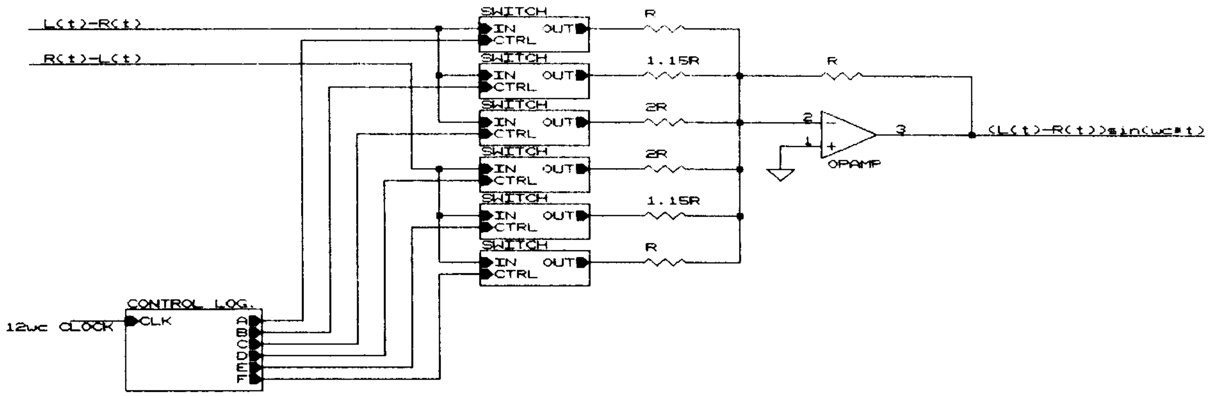
Basic operation of a matrix-type FM stereo encoder

Fig. 1



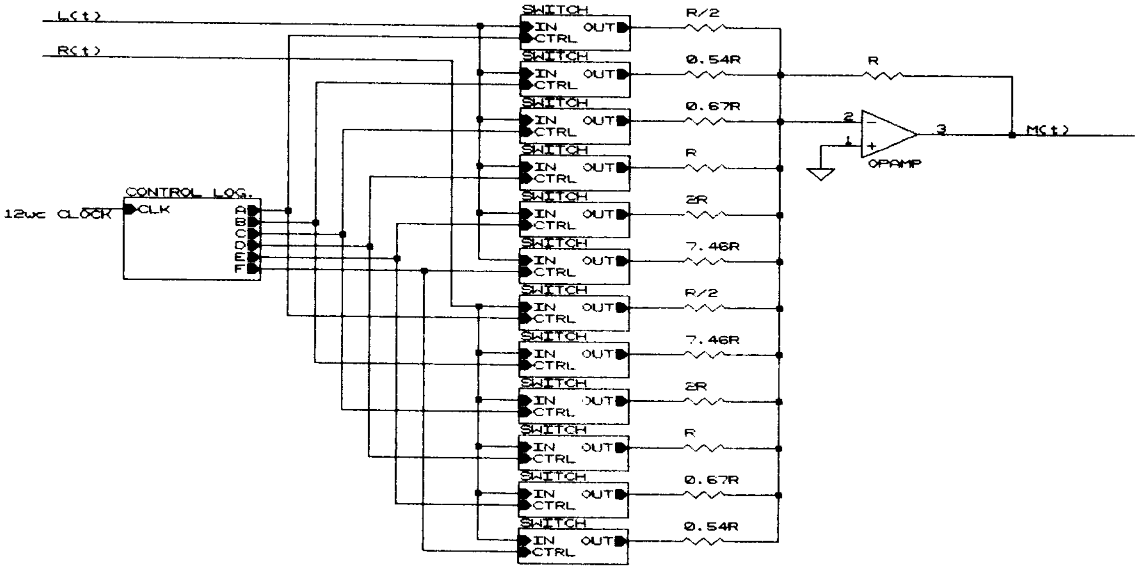
Basic operation of a switching type FM stereo encoder

Fig. 3



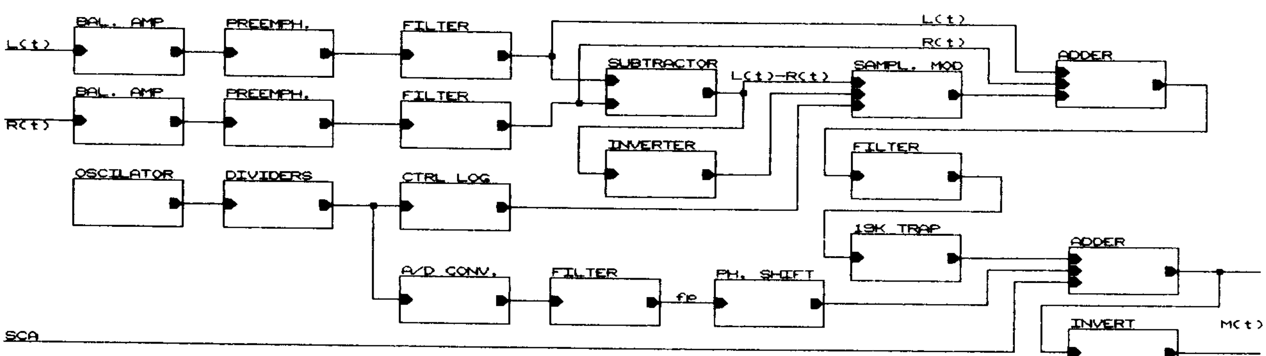
Sampling multiplier implementation

Fig. 4



Sampling encoder implementation

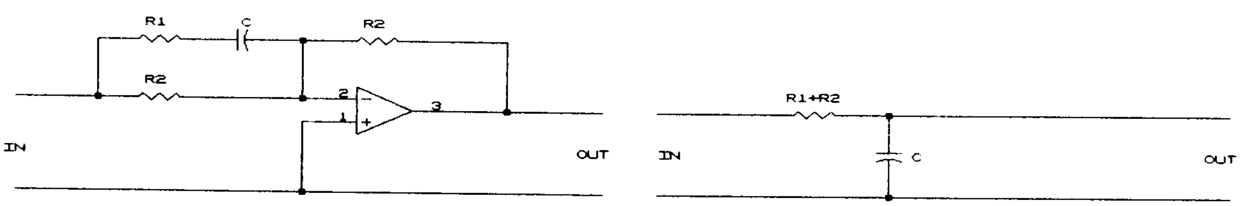
Fig. 7



Power supplies and measuring circuits ommited for clarity

Final block diagram of sampling MPX encoder

Fig. 15



Preemphasis circuit

Deemphasis circuit

Fig. 12

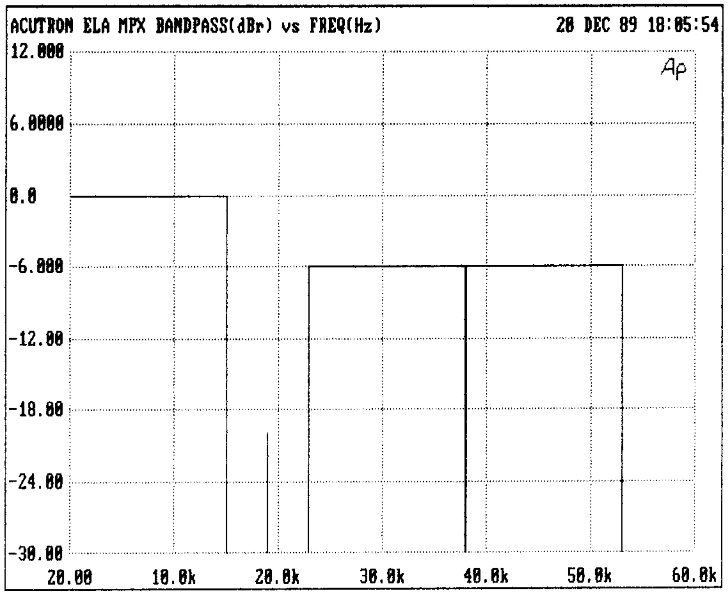


Fig.2-Simulated spectra of a multiplex signal

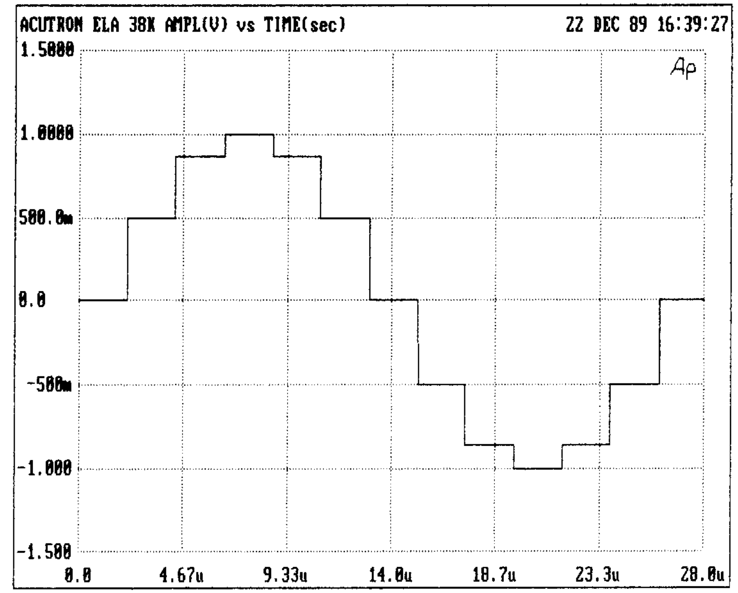


Fig. 4 Multiplier 38kHz sampled waveform

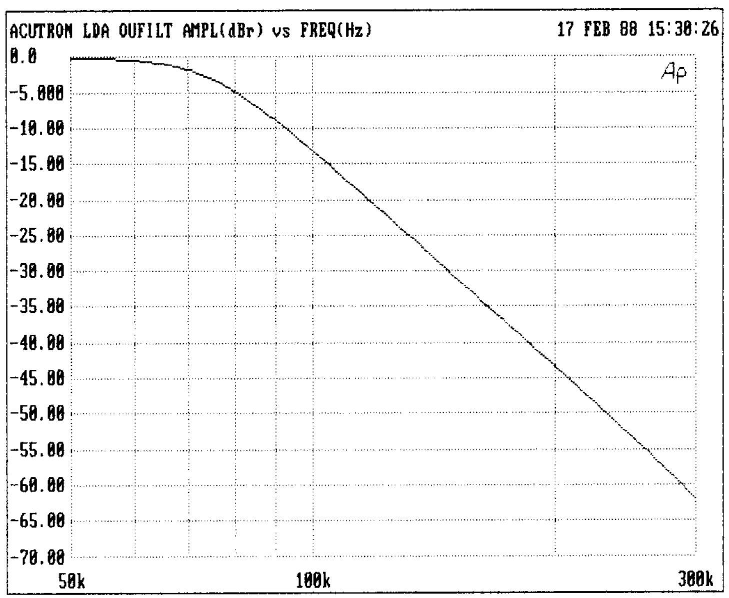


Fig. 4 - Frequency response of output filter

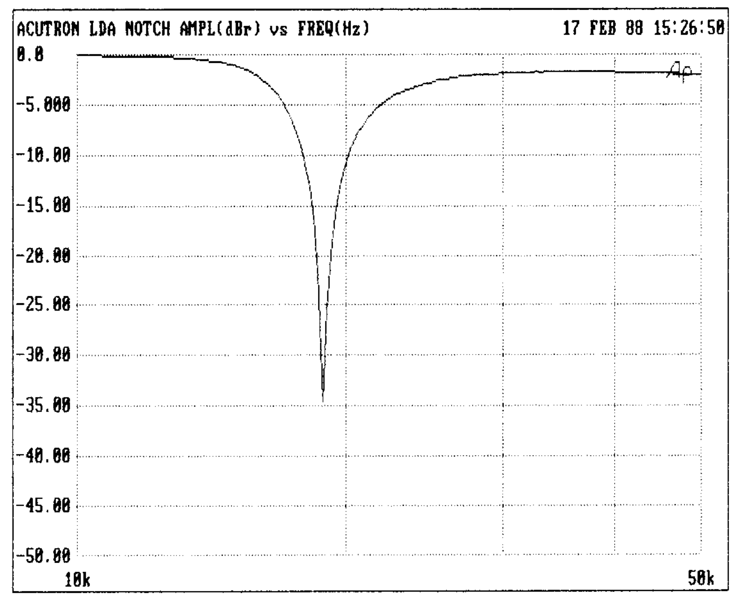


Fig. 8 - Frequency response of notch filter

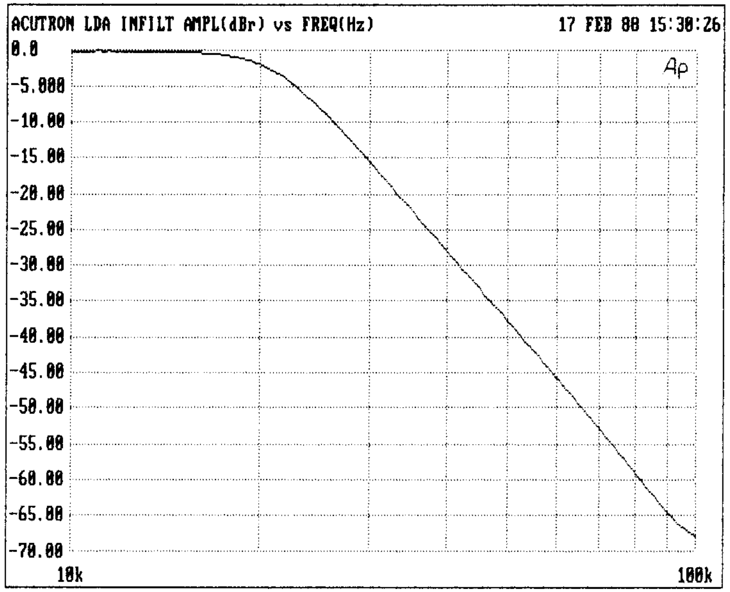


Fig. 9 - Frequency response of input filter

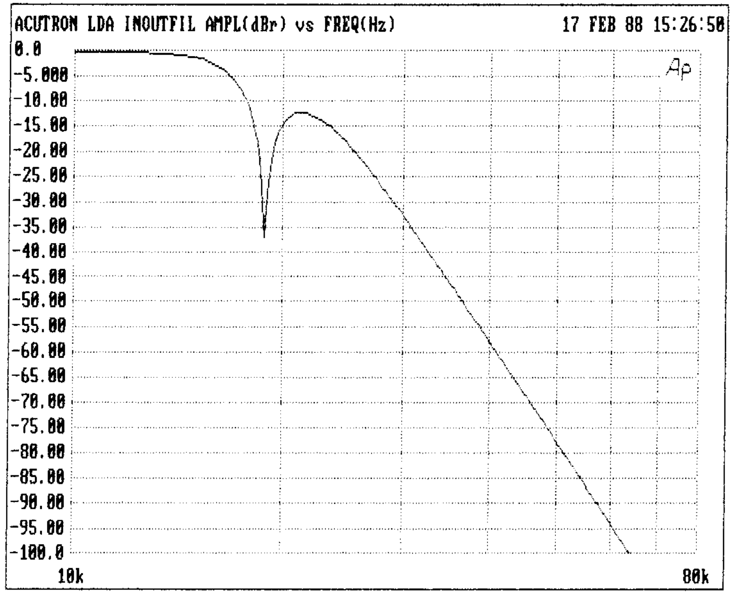


Fig. 10 - Frequency response of combined input/output filters

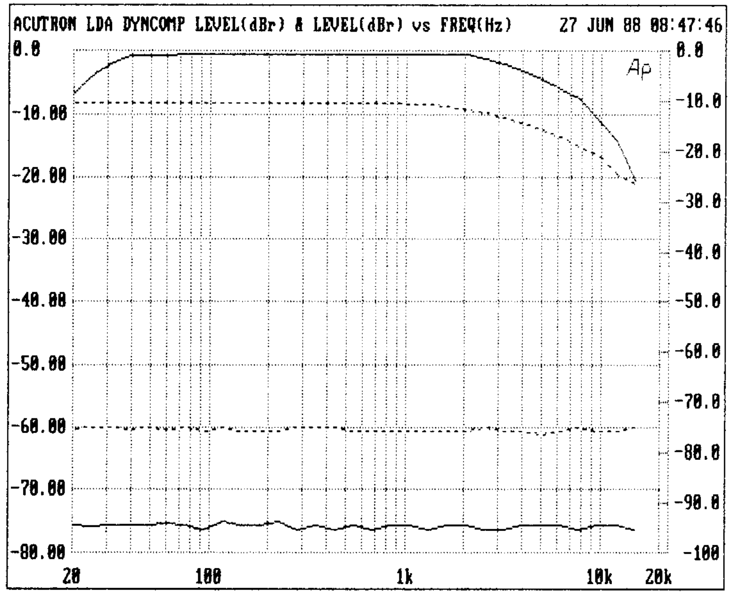


Fig. 19 - Maximum output level (upper trace) and noise (lower trace) of the new sampling encoder (solid), and of another top grade unit (dashed) measured again over the complete chain.

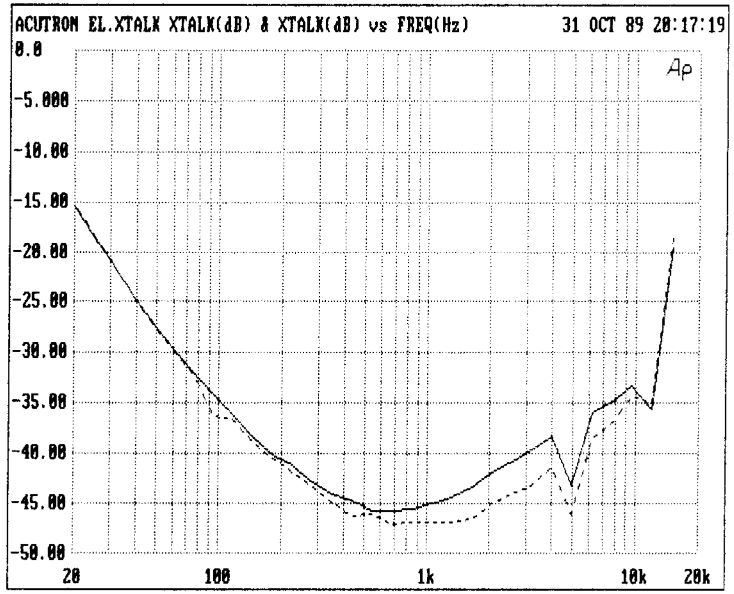


Fig. 20 - Crosstalk between channels, measured over the complete transmitting-receiving chain.

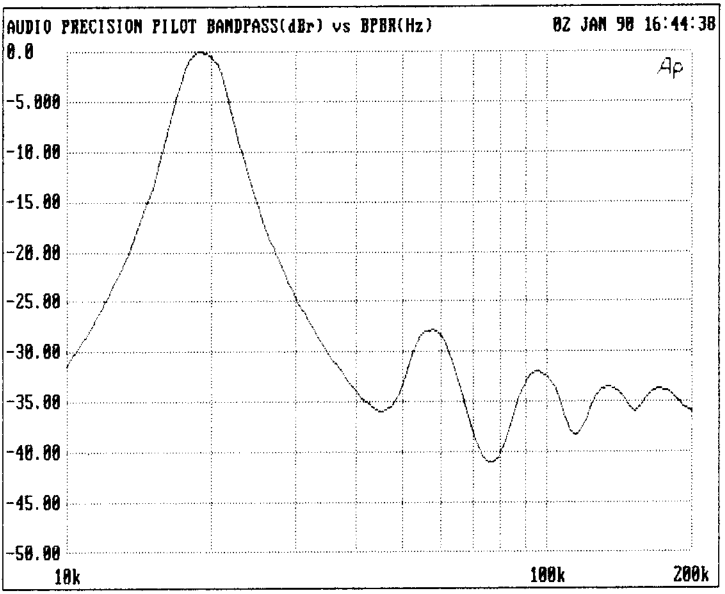


Fig. 11 - 1/3 octave spectral analysis of pilot waveform

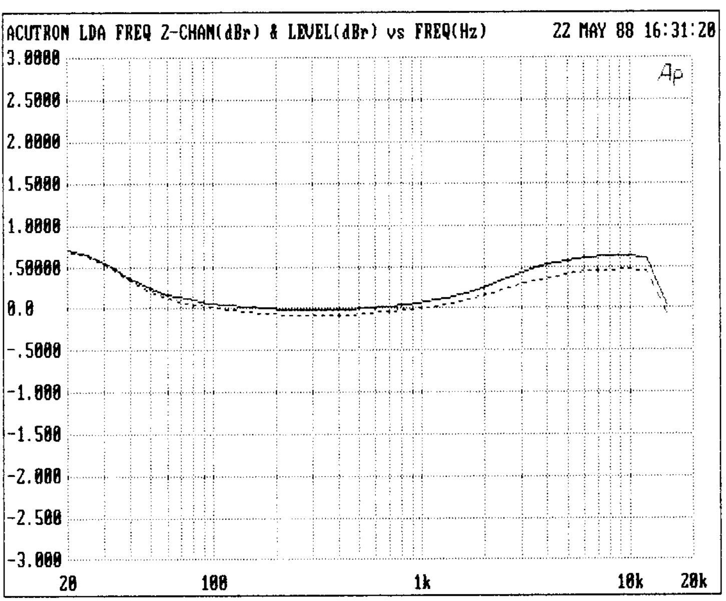


Fig. 16 - Frequency response of both channels, measured over the complete transmitting-receiving chain.

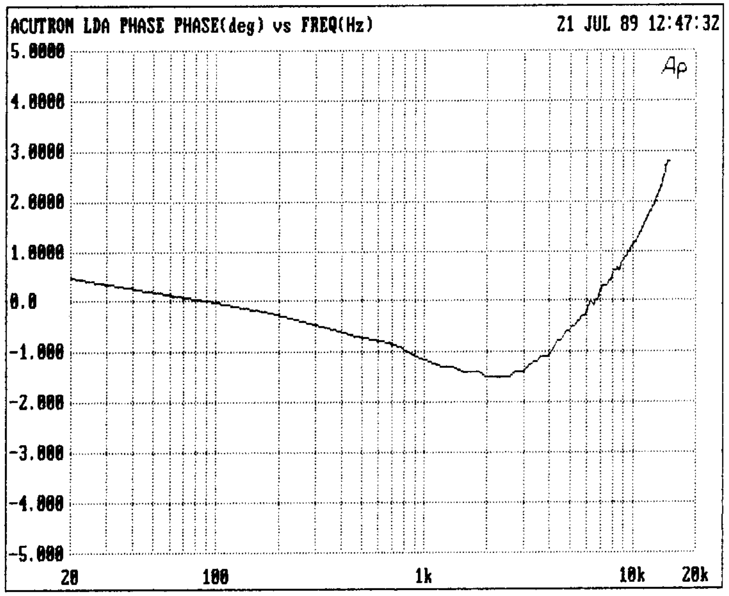


Fig. 17 - Phase deviation between channels, measured over the complete transmitting-receiving chain.

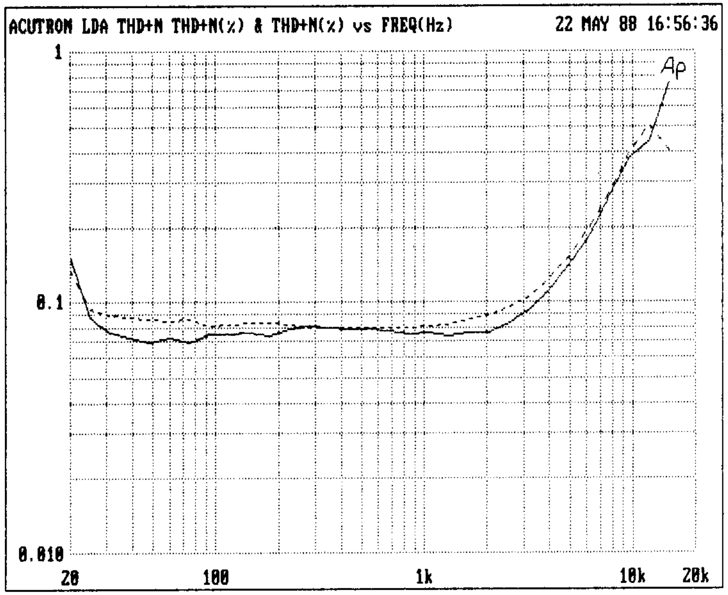


Fig. 18 - THD+N of both channels, measured over the complete transmitting-receiving chain, 80kHz measuring bandwidth.

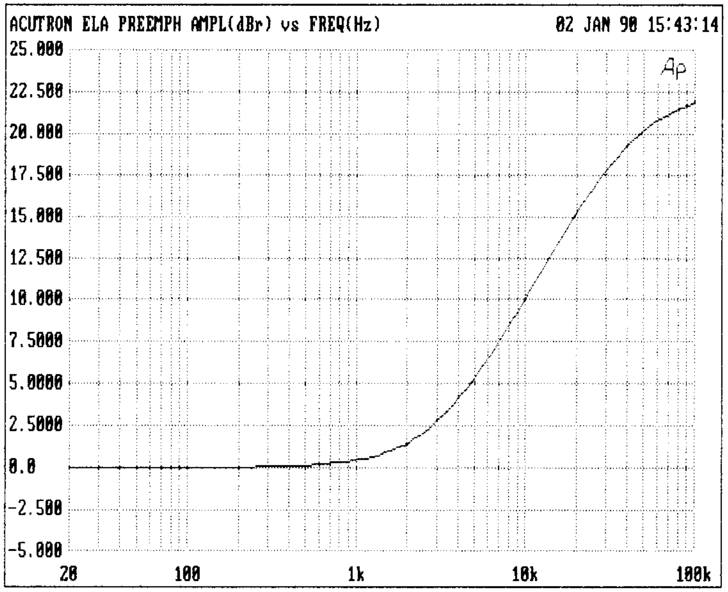


Fig. 13 - 50uS preemphasis function with high frequency cut.

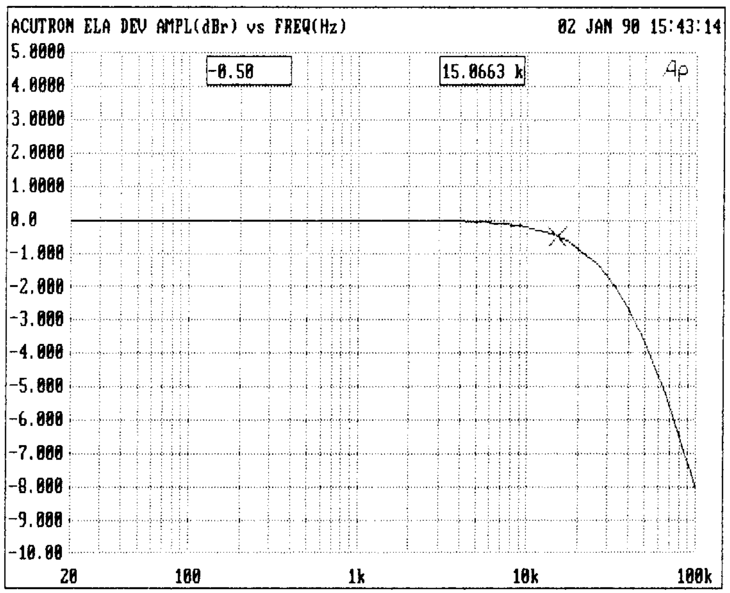


Fig. 14 - 50uS preemphasis/deemphasis function deviation.