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Development of an alternative AM stereo decoder for C-QUAM

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FACULTY OF ELECTRICAL ENGINEERING

EINDHOVEN UNIVERSITY OF TECHNOLOGY

TELECOMMUNICATIONS GROUP EC

Development of an alternative AM stereo decoder for C-QUAM

by : G.H. Schut

Report of a graduation project performed from December 1989 until June 1990 at the Product & Concept Application Laboratory Eindhoven of Philips Components.

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SUMMARY

This report is the conclusion of a graduation project of the study of Electrical Engineering at the Eindhoven University of Technology. The project has been performed at the Product and Concept Application Laboratory Eindhoven (PCALE) of Philips Components.
The goal was to develope an alternative AM stereo decoder for the C-QUAM system, based on a previous feasibility study. This decoder should be competitive with the existing Motorola decoder in price and performance. Part of this decoder should be an IF amplifier/limiter, recently developed at PCALE.
The decoder is based on the simultaneous demodulation of the amplitude and phase information.
A problem is the fact that the phase of the C-QUAM system can contain a DC component, which cannot be detected in the used system. By using secondary properties of the phase, it is possible to regenerate an approximation of this missing DC component. A circuit has been made, which realises this approximation.
A breadboard model of the total decoder has been developed and tested. The performance under ideal conditions is good compared to the Motorola decoder; the channel separation is even slightly better than that of the Motorola decoder.
When built in a radio the decoder does not function as well as the Motorola decoder, although the channel separation is slightly better. Especially the harmonic distortion leaves much room for improvement. The reason for this is the fact that the system used is very sensitive to distortion of the phase, since the DC component is generated from properties of the phase. The distortion is caused by asymmetric filters and the front end.
The conclusion is that this decoder is not competitive with the Motorola decoder on the grounds of its sensitivity to phase distortion. The decoder also needs some variable resistors for fine adjustments. A possible solution that might prove to be too expensive, is reduction of the phase distortion by using better filters and front end in the radio.

LIST OF ABBREVIATIONS

AGC	Automatic Gain Control
AM	Amplitude Modulation
C-QUAM	Compatible QUadrature Modulation
EUT	Eindhoven University of Technology
FCC	Federal Communications Committee
FM	Frequency Modulation
IF	Intermediate Frequency
MF	Medium Frequency
PCALE	Product & Concept Application Laboratory Eindhoven
PLL	Phase Locked Loop
ISB	Independent Sideband Modulation
PM	Phase Modulation
QUAM	QUadrature Amplitude Modulation
THD	Total Harmonic Distortion
VCO	Voltage Controlled Oscillator
VHF	Very High Frequency

ACKNOWLEDGEMENT

I would like to express my thanks to Mr. A. Garskamp, Mr. W. van Dooremolen and Mr. H. van Glabbeek for making it possible for me to perform my graduation project at PCALE.

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Geldrop, 25 July 1990

CONTENTS

1. INTRODUCTION

In vast areas with a low population density it is not efficient to build up an VHF radio net, because of the low reach of VHF transmitters. Listeners in these areas therefore have only MF-AM quality at their disposal. Until recently one of the disadvantages of AM was the fact that listeners could not enjoy stereo audio reproduction.
This resulted in a response from industry by which several companies each developed a different - and mutually incompatible - AM stereo system. Eventually the Motorola C-QUAM system became de facto the standard.
Motorola has developed for its C-QUAM system a special decoder IC, which is now being followed by a whole range of IC's, especially made for C-QUAM stereo. At this moment Motorola is the only manufacturer of such a decoder.
At the Product and Concept Application Laboratory PCALE of Philips Components in Eindhoven an IF-amplifier/limiter has been developed, which makes it possible to include an AM stereo decoder without too much extra cost (as well as components). This AM stereo decoder is also suitable for integration on the same chip as the amplifier/limiter. In 1989 the possibility of developing such a decoder using this IF-amplifier/limiter was investigated [1] and the result was promising enough to continue this work as a graduation project.
In this report the history of AM stereo will be sketched in chapter two together with the C-QUAM system. In chapter three an alternative for the decoding of C-QUAM will be discussed. The realisation of a decoder using this method will be treated in chapter four. In chapter five the performance of the decoder is discussed, based on simulations and experiments with a breadboard model of the decoder. In chapters six and seven the conclusion from this project is drawn and some recommendations to future investigations are proposed.

2. WHAT IS AM STEREO?

Stereo is a feature in audio reproduction, that has been established already a long time ago. It is generally used in commercial broadcasting on the FM-band. It is a system of dividing the sound into two channels to simulate the spatial effect for the human ear. Since two audio channels (left and right) need to be transferred, it takes up more bandwidth than monaural transmissions. Normally the sum (left + right) and the difference (left - right) are transferred instead of left and right independently.

AM stereo is a method of transferring the stereo information in the Medium-Frequency band and it has been introduced only recently. Because of the fact that up to then transmissions were monaural, no reservations in bandwidth have been made to allow stereo transmissions. This means that the stereo system has to fulfill certain requirements in order to become successful:

- the system may not use more bandwitdh than mono transmitters, because of the present standards.
- the system has to be compatible with mono receivers.
- during (and after) the introduction of AM stereo all listeners should have flawless reception.
- AM stereo receivers should be attractively priced in order to ensure acceptance by the listeners.

The requirements listed above make it difficult to develope a system that fulfills all of the requirements and up to today there is no such system available.

Because of the compatibility requirement, the envelope of the signal should be the sum of the left and the right channel. Since the bandwidth is restricted it is not possible to develope a similar system as FM stereo. The use of multiple modulation techniques is necessary, for example Amplitude Modulation combined with Phase Modulation, in order to transmit more information in approximately the same bandwidth.

2.1. The history of AM stereo

The idea of transmitting stereo information in the Medium-Frequency band is not new. In 1926 the American R.K. Potter received a patent for such a system. In the 1950s stereophonic sound reproduction made its way to the audience and as a consequence several companies investigated different techniques of transmitting stereo in the Medium-Frequency band.

Several systems, like those of Philco [2], RCA [3] en Kahn Research Laboratories [4], were proposed to the Federal Communications Committee (FCC), a body instituted to provide the United States with standards for new services. The FCC rejected these systems, arguing that there was no interest from listeners for these systems and that they were not yet technically feasible. This decision ended the discussion about AM stereo for some years, but in the late seventies other companies tried to persuade the FCC of accepting a standard for AM stereo.

Five different, mutually incompatible, systems were proposed, using the following modulation techniques:

- combined amplitude and frequency modulation, whereby the AM is formed by left plus right (L+R) in order to make the envelope compatible for mono receivers and the frequency modulation information by left minus right (L-R). This system was developed by RCA in the fifties and now proposed by Belar.
- combined amplitude and phase modulation, analogous to the RCA/Belar system, but using phase modulation instead of frequency modulation. This system was developed by Magnavox [5].
- quadrature amplitude modulation. The in-phase carrier is amplitude modulated by L+R and the out-of-phase carrier is modulated with L-R. The envelope here is not compatible with monophonic transmission. This system was developed by Harris [6].
- independent sideband modulation (ISB). The left audio channel is mainly transmitted in the lower sideband and the right channel is transmitted in the upper sideband. This makes it possible to use two slightly mistuned mono receivers to receive the stereo information. This system was developed by Kahn [4].
- modified quadrature amplitude modulation (C-QUAM). The pure quadrature signal, similar to the QUAM-signal of Harris, is modified to get a compatible envelope. This system is called C-QUAM and was developed by Motorola [7].

The above systems are all a compromise. Non of the systems fulfill the requirements completely and therefore it is difficult to decide which system is the best.

In 1980 the FCC favoured the Magnavox system, but it revoked its decision a few months later after heavy criticism of the competing companies. In 1982 the FCC decided to let the market make the final decision, risking the abandonment of the idea of AM stereo. After an enormous marketing effort by Motorola the major part of the American AM stereo broadcasters was using the C-QUAM system, although the system is twice as expensive as the Magnavox system. In the same year the Australian government selected the Motorola C-QUAM system as their national standard for AM stereo.

This finally concluded the battle for the AM stereo standard in favour of Motorola.

2.2. The Motorola C-QUAM system

The abbreviation C-QUAM stands for Compatible QUadrature Amplitude Modulation. The C-QUAM system is based on quadrature amplitude modulation and has been modified for compatibility with mono receivers.

In quadrature modulation a carrier with an angular frequency ω_c is in amplitude modulated by a signal X(t) and a carrier, shifted in phase by 90 degrees, is modulated with Y(t). For stereo broadcasts X(t) is represented by the sum of the left and the right audio channel and Y(t) by the difference between left and right.

Thus it is possible to transmit the full stereo information within the same bandwidth as the mono signal.

This leads to the following expression for the total signal with ω_c as angular carrier frequency:

$$[1+L(t)+R(t)]\cos(\omega_c t) - [L(t)-R(t)]\sin(\omega_c t) \tag{2.1}$$

In future the time variable t will be ignored in L(t) and R(t), but silently assumed present.
Formula (2.1) is equivalent to:

$$[(1+L+R)^2 + (L-R)^2]^{½}\cos\{\omega_c t+\phi(t)\} \tag{2.2}$$

$$\text{with: } \phi(t) = \arctan\left[\frac{L-R}{1+L+R}\right] \tag{2.3}$$

The envelope of the signal is not compatible with mono receivers, since:

$$[(1+L+R)^2 + (L-R)^2]^{½} \neq 1+L+R$$

In the Motorola C-QUAM system the signal from formula (2.2) is multiplied by:

$$\cos(\phi(t)) = \left[\frac{1+L+R}{[(1+L+R)^2 + (L-R)^2]^{½}}\right] \tag{2.4}$$

In figure 2.1 both QUAM and C-QUAM are shown in a vector diagram.

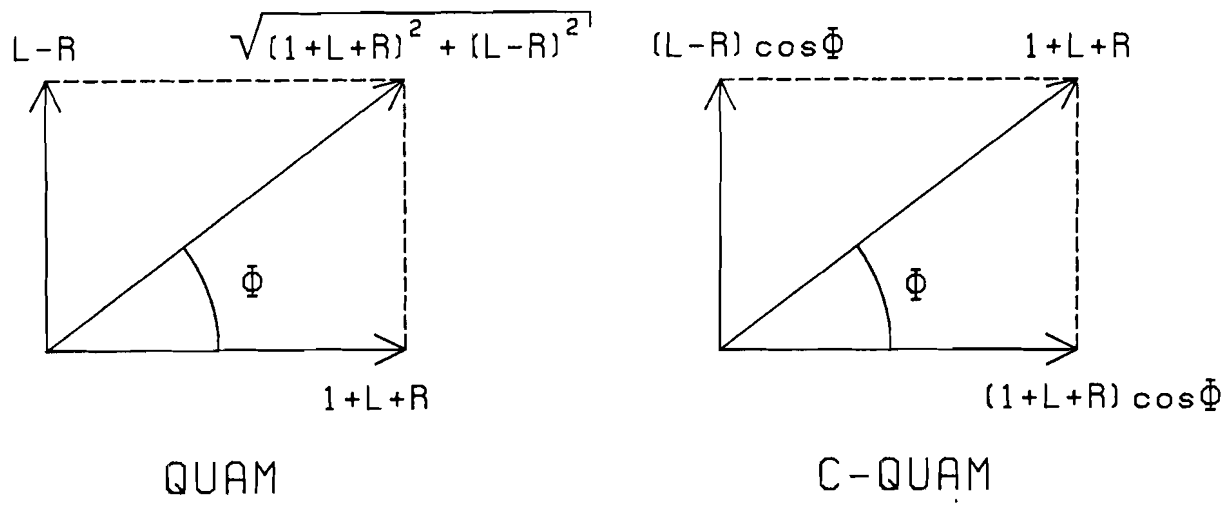

Figure 2.1. Comparison of QUAM and C-QUAM in a vector diagram.

As can be seen in figure 2.1 the multiplication of QUAM with $\cos\{\phi(t)\}$ leads to the following expression for the C-QUAM signal:

$$(1+L+R)\cos\{\omega_c t + \phi(t)\} \tag{2.5}$$

The multiplication by $\cos\{\phi(t)\}$ can be realised technically in a simple way. The pure QUAM-signal is fed to a limiter, which removes the amplitude information from the signal. The resulting signal:

$$\cos\{\omega_c t + \phi(t)\}$$

is then modulated in amplitude by (1+L+R), which results in formula (2.5). Now we have enough information to understand how the C-QUAM transmitter can be realised. In figure 2.2 the diagram of a C-QUAM transmitter is presented.

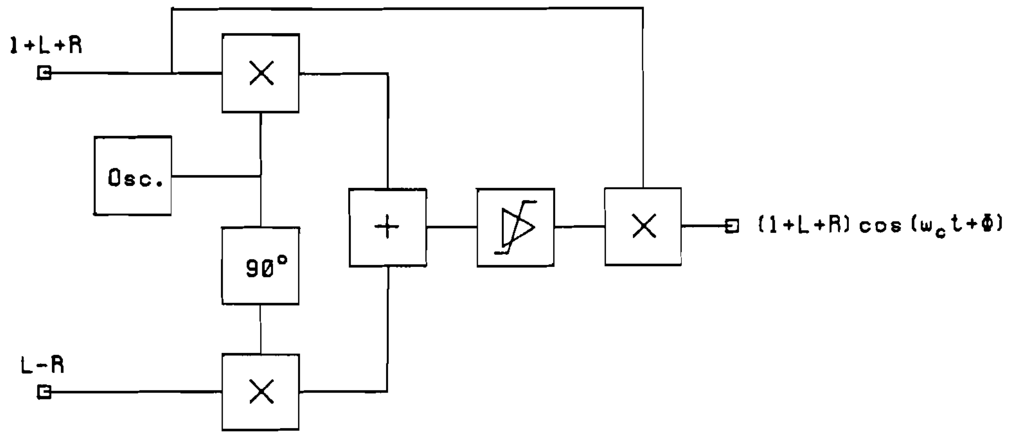

Figure 2.2. Diagram of a C-QUAM transmitter.

The C-QUAM system allows for automatic detection of stereophonic transmissions. This is done by adding to the (L-R) information a pilot signal.
This pilot signal is a tone with a frequency of 25 Hz and a modulation level of 5%, compared to the unmodulated carrier level.
At the receiving end the pilot can be detected, by filtering the pilot from the L-R information using a bandpass filter with a center frequency of 25 Hz. If a correct pilot tone is present the decoder may switch to stereophonic reception.

2.3. The Motorola stereo decoder

The Motorola MC13020P stereo decoder is based on the principal of synchronous detection of the QUAM signal. This means that the QUAM-signal in formula 2.1 is

multiplied with an in-phase carrier without modulation. This gives us the AM information of the in-phase carrier (L+R in formula 2.1). The signal is also multiplied with a 90 degrees out-of-phase carrier without modulation. This gives us the AM information of the out-of-phase carrier (L-R in formula 2.1).
The received C-QUAM signal therefore has to be converted to pure QUAM, since C-QUAM is QUAM, multiplied by $\cos\{\phi(t)\}$. This means that the signal has to be divided by $\cos\{\phi(t)\}$. This division can be achieved by comparing the output of the in-phase synchronous detector with the envelope (see figure 2.3).
In order to understand this principle, we have to rewrite the expression for the C-QUAM signal in the following manner:

$$(1+L+R)\cos\{\omega_c t + \phi(t)\} = (1+L+R)\cos\{\phi(t)\}\cos(\omega_c t) - (L-R)\cos\{\phi(t)\}\sin(\omega_c t) \qquad (2.6)$$

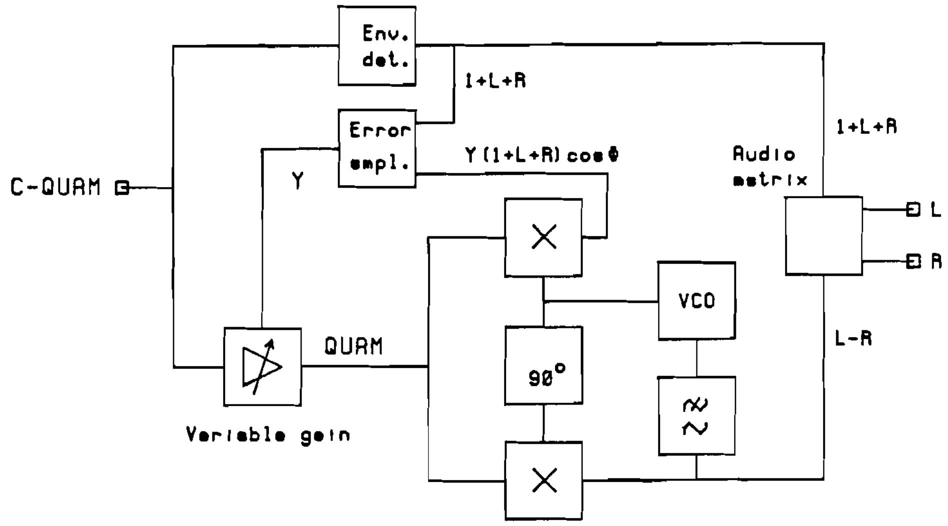

Figure 2.3. Diagram of the Motorola C-QUAM decoder.

The output of the in-phase detector equals $(1+L+R)\cos\{\phi(t)\}$. This result is compared with the output of the envelope detector, which equals 1+L+R. The difference is used to modulate the C-QUAM signal in amplitude in such a way that the difference is minimised. The output Y of the error amplifier (which amplifies the difference between the envelope and the output of the in-phase detector) is then given by:

$$(1+L+R) - Y(1+L+R)\cos\{\phi(t)\} = Y/A \qquad (2.7)$$

If the amplification A of the error amplifier is sufficiently large, then Y/A will approach zero. This means that Y will equal $1/\cos\{\phi(t)\}$.

The out-of-phase detector gives us the difference information (L-R). Now L+R and L-R are fed to a matrix that adds and subtracts them to get L and R respectively.
The reason that the decoder uses such a complicated system is, that the sum and the difference are both contained in an AM signal. This means that the same modulation technique is used to transfer L+R and L-R. Thus they will undergo the same kind of distortion on the way from the transmitter to the decoder. This means that the difference between L+R and L-R will be conserved better than in a mixed demodulation system, where two different modulation techniques are used to transfer L+R and L-R and these different modulation techniques are susceptible to different kinds of distortion.

3. A DECODING ALTERNATIVE

3.1. Introduction

From the description of the C-QUAM signal it is clear that the signal is modulated both in amplitude and in phase.
If these two parts can be independently demodulated at low cost, it is possible to derive L and R in an easy way.
At the application laboratory PCALE an IF-amplifier/limiter with some interesting features has been developed. The amplifier has a dynamic range of over 80 dB, which makes the use of an Automatic Gain Control unnecessary. It automatically demodulates the AM-information and gives at the same time a square wave, which contains the PM (or FM) information.
These properties allow us to design an AM stereo decoder in a different and probably a more economical way than Motorola.
Now we only need to detect the phase information and then it is possible with the aid of some relatively simple circuitry to obtain L-R from the phase information. This method was proposed in [1] and will be developed further.
In order to be able to judge the decoder on its performance we need to set some targets. One way of obtaining those targets is to analyse the performance of the Motorola decoder (see appendix A). The new decoder will have to fulfill the following requirements:

- the signal-to-noise ratio should be 60 dB or more
- the channel separation α should be at least 30 dB for frequencies, ranging from 300 Hz to 10 000 Hz.
- the total harmonic distortion (THD) should be less than 1% for a modulation index $m \leq 0.75$. This upper limit for the individual peaks of Left and Right is recommended by Motorola of L and R to a maximum of $m=0.75$.

3.2. Structure of the decoder

The goal of the decoder is to derive the sum (L+R) and the difference (L-R) information and to combine these in a matrix in order to get the left and right audio channel.
Since the sum is already present at the envelope output of the IF-amplifier/limiter we only need to obtain L-R.
As described in paragraph 2.2, L-R can be found in the phase of the C-QUAM signal, according to the following formula for the phase:

$$\phi(t) = \arctan\left[\frac{L-R}{1+L+R}\right] \qquad (2.3)$$

Looking closely at expression 2.3, it will be clear what steps must be taken to obtain L-R.

First we have to put the phase through a circuit, which operates as a tangent function.
We then get at the output of the tangent:

$$\tan\{\phi(t)\} = \frac{L-R}{1+L+R} \tag{3.1}$$

Now we only have to multiply this result with (1+L+R) to get L-R.
In figure 3.1 we can see the structure of the decoder.

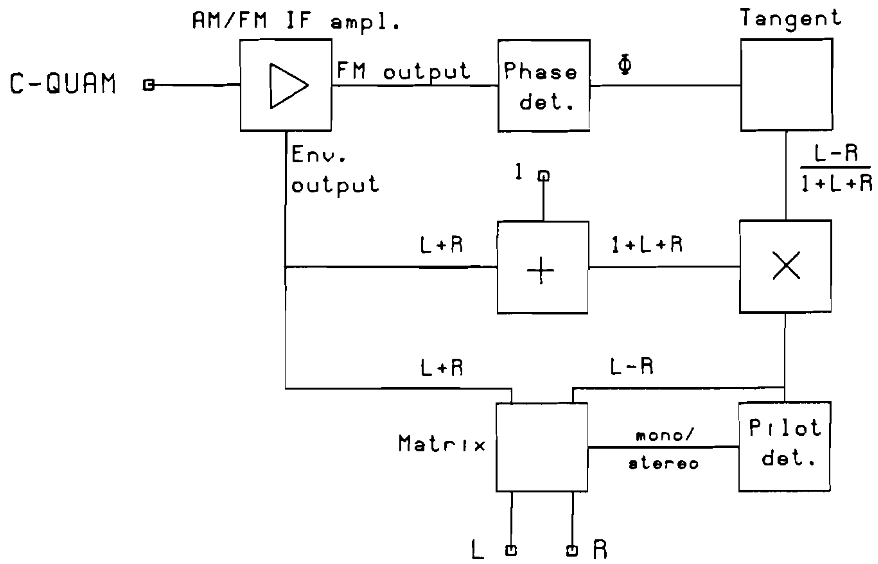

Figure 3.1. Diagram of the alternative decoder.

The PM-output of the IF-amplifier/limiter is fed to a phase detector, which demodulates the phase information. This phase information is fed to a tangent function and its output is multiplied with 1+L+R. This 1+L+R is made by adding a DC bias to L+R, which comes directly from the envelope output of the IF-amplifier/limiter.
The result of the multiplication (L-R) and the envelope output of the amplifier/limiter (L+R) are fed to a matrix to get the left and right audio channel.
A pilot detection circuit is added to identify whether or not the received signal is a C-QUAM AM stereo signal. Its function is to switch the decoder to stereo reception when a correct pilot tone is detected.
In the next chapter the blocks in figure 3.1 will be studied in more detail.

3.3. A flaw in the system

The system described in the previous paragraph does not function properly under certain conditions. When receiving a signal which has pure AM information (L=R)

or pure PM information (L=-R) then there is no problem. However when transmissions occur in which only one channel (left or right) is being broadcast then a problem arises. This problem is caused by the fact that the detected phase is not entirely correct: the shape is O.K. but the DC level is not.
For single channel (L-/R-only) single tone modulation (L = m $\sin(\omega_M t)$, R = 0) $\phi(t)$ is given by:

$$\phi(t) = \arctan\left[\frac{m\sin(\omega_M t)}{1+ m\sin(\omega_M t)}\right] \tag{3.2}$$

where: m = modulation index ($0 \leq m \leq 1$)
ω_M = angular frequency of the sine wave

An example of the phase is given in figure 3.2, where m=0.8 and $\omega_M = 2\pi$ 1000 Hz.

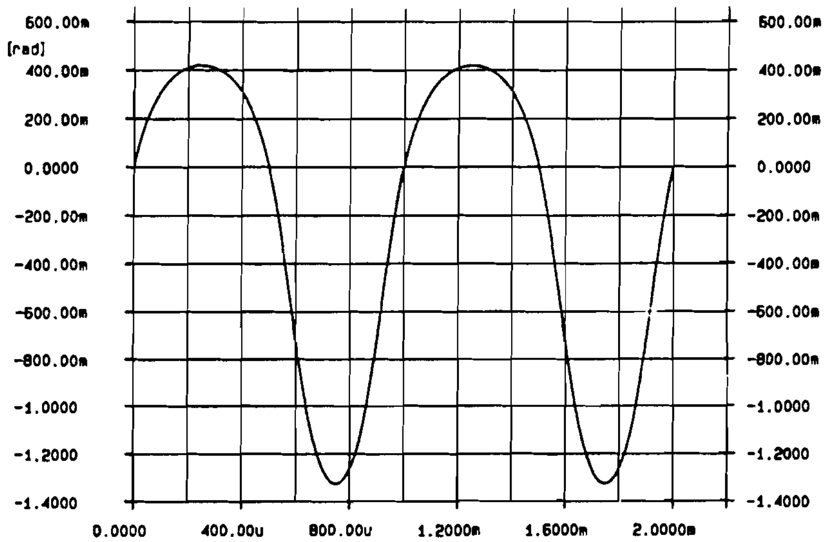

Figure 3.2. The phase of the C-QUAM signal at L-only and m=0.8.

The average value of the phase in figure 3.2 is not equal to zero, because the surface of the curve above the horizontal axis is smaller than the surface under the horizontal axis. An average value not equal to zero means:

$$\phi_{AVG} = \frac{1}{T}\int_0^T \phi(t)\, dt \neq 0 \tag{3.3}$$

This integral is best calculated with the aid of a computer and it gives the following curve of the average phase versus modulation index (see figure 3.3).

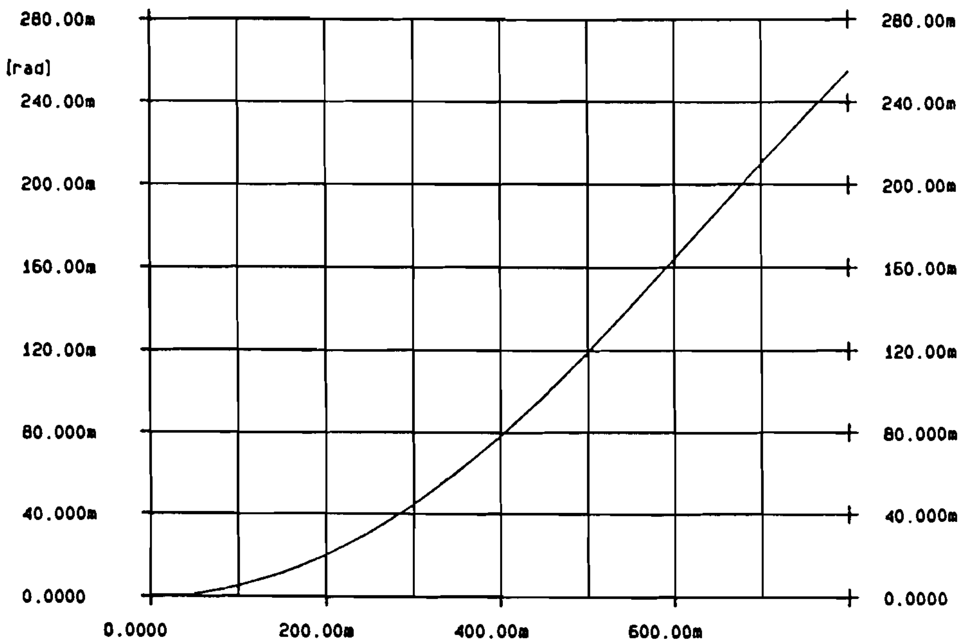

Figure 3.3. Average phase versus modulation index at L-only.

A constant in the phase is equal to a constant phase shift or time delay. Unfortunately there is no way of establishing this phase shift or time delay, because there is no reference available. It is therefore theoretically impossible to obtain the true phase, i.e. the phase including the constant value.
The consequence of the missing DC value lies in the fact that the tangent will not compensate the arctangent distortion of the phase, because the phase will now be in the wrong part of the tangent curve. This means that the distortion increases rapidly with increasing modulation index.
A possible remedy is to make a DC value afterwards, using properties of the detected phase. By adding this DC value to the phase at the input of the tangent circuit, it may be possible to restore the phase.
When we look again at figure 3.2 we see two possibly useful properties:

- the zero crossings of the detected phase are no longer spaced equidistantly, since the whole curve will be shifted upwards.
- the positive and negative peak value are not the same in absolute value.

3.3.1. Solution 1: a zero-crossing detector

Now we need to measure in one way or another how much the zero crossings of the detected phase are shifted, due to the missing DC component in the phase.
One way of measuring this is to convert the phase into a square wave having the same zero crossings as the phase. The average value of this square wave will then be unequal to zero, since the zero crossings within one period of the square wave will not be spaced equidistantly.

The zero crossings of the detected phase $\phi_{DET}(t)$ are given by the following equations:

$$\phi_{DET}(t) = 0 <=> \phi(t) - \phi_{AVG} = 0 <=> \tag{3.4)(3.5}$$

$$\arctan\left[\frac{m\sin(\omega_M t)}{1+ m\sin(\omega_M t)}\right] = \phi_{AVG} \; (< 0) \tag{3.6}$$

The values t_o when the zero crossings occur, are those values of t within the period T of the sine wave when the following equation holds:

$$m \sin(\omega_M t_o) = \frac{\tan(\phi_{AVG})}{1 - \tan(\phi_{AVG})} \tag{3.7}$$

The time t during which the square is negative, is given by:

$$\omega_M t_o < \omega_o t < 3\pi - \omega_M t_o$$

The time t during which the square is positive, is given by:

$$0 < \omega_M t < \omega_M t_o \quad \text{and} \quad 3\pi - \omega_M t_o < \omega_M t < 2\pi$$

The time during which the square is negative is not equal to the time during which the square is positive. This means that if the square wave is symmetric, the average value will not be zero. With this information we can calculate the average value of the square wave at different modulation indices (see figure 3.4).

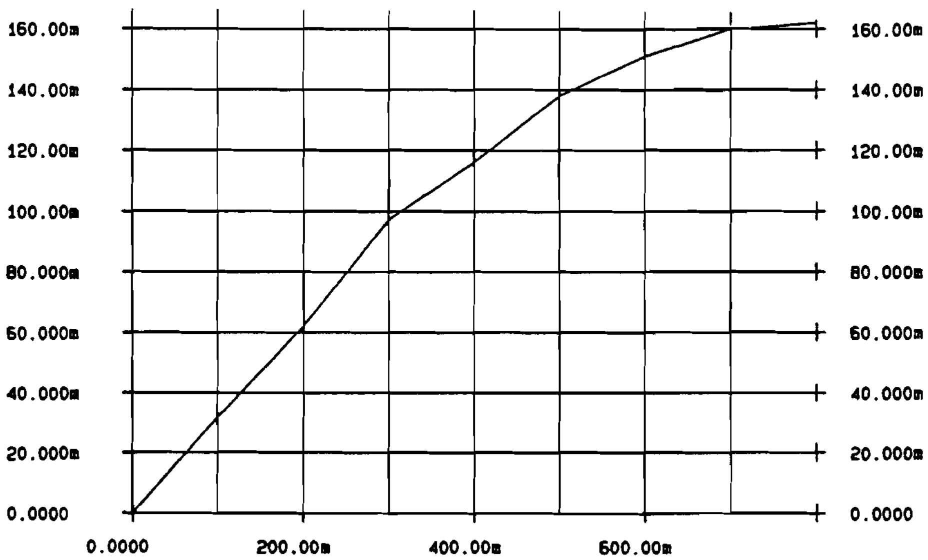

Figure 3.4. Average value of square wave versus modulation index (L-only).

3.3.2. Solution 2: using the peak difference

Another solution is to determine the difference in positive and negative peak value of the detected phase. The maximum and minimum value of $\phi_{DET}(t)$ occur when the sine wave reaches its maximum (+1) and minimum value (-1) respectively. Thus:

$$\phi_{DET,MAX} = \arctan[m/(1+m)] + \phi_{AVG} \tag{3.8}$$

$$\phi_{DET,MIN} = -\arctan[m/(1-m)] + \phi_{AVG} \tag{3.9}$$

This gives us the following curve of the peak difference versus modulation index (see figure 3.5).

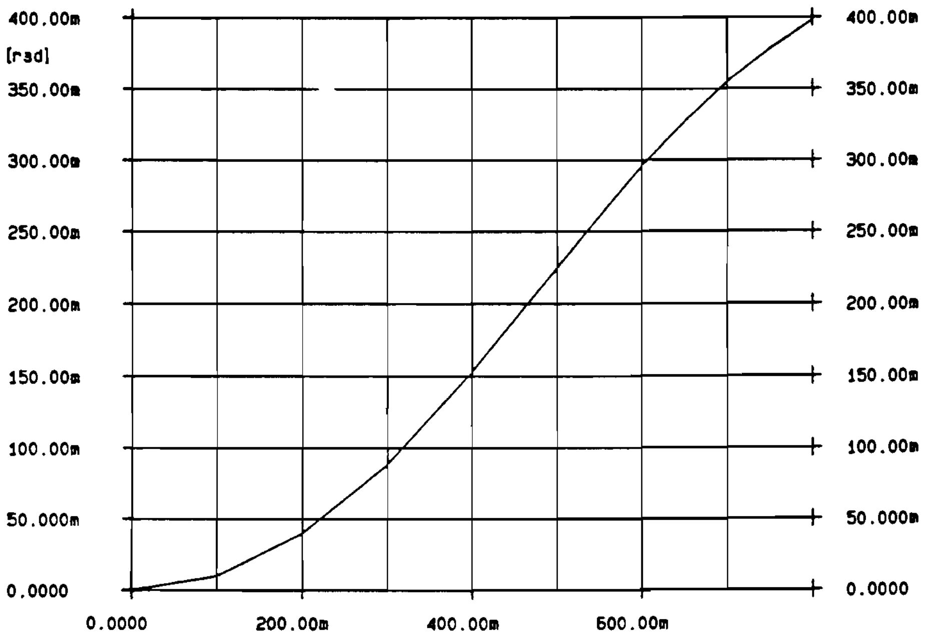

Figure 3.5. Peak difference versus modulation index (L-only).

3.3.3. Conclusion

In the previous two paragraphs two methods are described for regenerating the missing DC value from properties of the detected phase. Comparing figures 3.4 and 3.5 with figure 3.3 it is clear that the peak difference gives us the best curve for the approximation of the missing DC value in the phase. Only at high modulation index the curve starts to deviate from the desired curve. It remains to be seen whether this method gives sufficient compensation at the input of the tangent to reduce the distortion sufficiently. But without a method of correcting the phase the whole system will not be capable of meeting the demands, given in paragraph 3.2. Therefore the system of using the peak difference will be used.

4. FUNCTIONAL BLOCKS OF THE DECODER

4.1. The phase detector

For the detection of the phase information we can use two methods [1]:
- direct phase detection
- indirect phase detection.

Direct phase detection can be done using a P.L.L. with a very small loop filter bandwidth. The major disadvantage of this method is the fact that it requires a very stable signal for the P.L.L. to remain in lock, since the bandwidth of the loop filter is very small. This means that the mixing oscillator must be very stable, which makes the system expensive. Therefore this solution is not practical.
Indirect phase detection can be achieved by using an FM detector followed by an integrator. The FM detector is sensitive to information contained in the frequency of the signal. The relation between frequency $\Delta\omega(t)$ and phase $\phi(t)$ is given by:

$$\phi(t) = \int_{-\infty}^{t} \Delta\omega(\sigma)\, d\sigma \tag{4.1}$$

Since the frequency is the derivative of the phase, it is possible to use an FM-detector, followed by an integrator, to obtain the phase information.

4.1.1. The FM detector

As detector we can use a P.L.L. with a larger loop bandwidth or - even simpler - a quadrature detector. The quadrature detector (or Q-detector) is a multiplier with a tuned circuit at one of its inputs, which causes a phase shift, which is a function of the frequency (see figure 4.1).

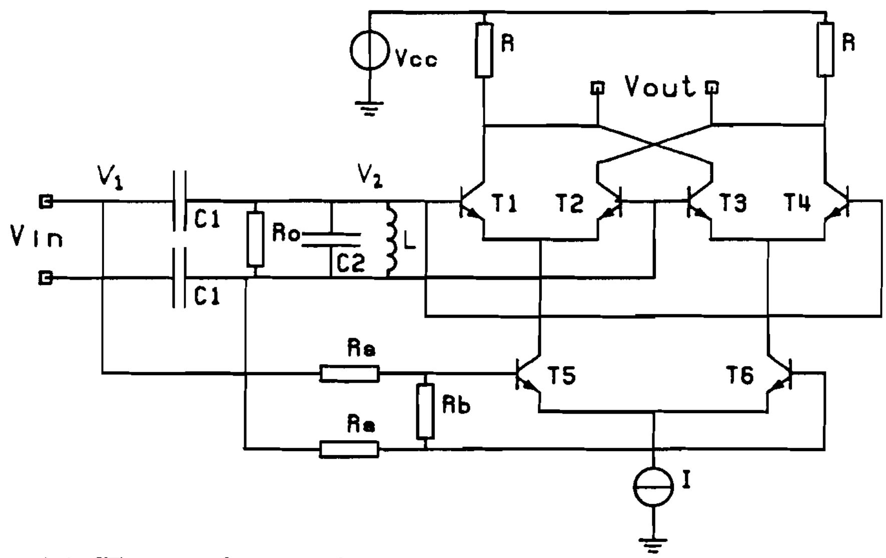

Figure 4.1. The quadrature detector.

The transfer function of the filter network is:

$$\frac{V_2}{V_1} = \frac{-\tfrac{1}{2}\omega^2 LC_1}{1-\omega^2 L[C_2+\tfrac{1}{2}C_1] + j\omega L/R_0} \tag{4.2}$$

When at the inputs of the multiplier two signals $E_1\cos(\omega_C t)$ and $E_2\cos(\omega_C t+\theta)$ with the same frequency but with a phase difference θ are presented, the output is then given by:

$$V_{OUT} = K\,E_1\,E_2\cos(\omega_C t)\cos(\omega_C t + \theta) = \tfrac{1}{2}K\,E_1\,E_2\,\{\cos(2\omega_C t+\theta)+\cos\theta\} \tag{4.3}$$

K is the amplification of the multiplier.
A low-pass filter removes the component of frequency $2\omega_C$, giving an output of:

$$V_{OUT} = \tfrac{1}{2}K\,E_1\,E_2\cos\theta \tag{4.4}$$

Now the phase shift θ of the tuned circuit is:

$$\theta = -\pi - \arctan\left[\frac{\omega L/R_0}{1-\omega^2 L[C_2+\tfrac{1}{2}C_1]}\right] \tag{4.5}$$

The resonant frequency is chosen to be the carrier frequency and we are interested only in the case where ω is near ω_C. We can apply the first two terms of the series expansion for $\arctan(x) = \tfrac{1}{2}\pi - (1/x)$ to give:

$$\theta \approx -3\pi/2 + \left[\frac{1-\omega^2 L[C_2+\tfrac{1}{2}C_1]}{\omega L/R_0}\right] \tag{4.6}$$

If we define the frequency variation $\Delta\omega$ by:

$$\omega = \omega_0 \pm \Delta\omega$$

and:

$$\omega_0^2 = (L[C_2+\tfrac{1}{2}C_1])^{-1}$$
$$\omega_0 >> \Delta\omega$$
$$\omega \approx \omega_0$$

then equation 4.6 becomes:

$$\theta \approx -3\pi/2 \pm R_0C/\omega_0\,(2\omega_0\Delta\omega + \Delta\omega^2) = -3\pi/2 \pm 2R_0C\,\Delta\omega \tag{4.7}$$

From equation 4.4 we see that the output of the multiplier after the low-pass filter will be:

$$V_{OUT} = K' \cos(-3\pi/2 \pm 2R_0C\,\Delta\omega) \approx -K'\sin(\pm 2R_0C\,\Delta\omega) \approx -2K'R_0C\,\Delta\omega \qquad (4.8)$$

Now we have a circuit that produces an output that is proportional to the modulation frequency. A disadvantage is the fact that if the carrier frequency is not equal to the resonant frequency an offset in the output will appear, because now there will be a phase shift between the two inputs of the multiplier even if the incoming signal is unmodulated. This means that we have to make an AC coupling between the Q-detector and the rest of the circuit to eliminate any DC changes in the phase, due to mistuning of the radio.
The output is made asymmetric by using current mirrors. The resistors with values R are replaced by current mirrors. This does not alter the functioning of the Q-detector.

4.1.2. The integrator

Since the Q-detector produces an output that is proportional to the modulation frequency, we need an integrator to make the output amplitude constant over the desired frequency range (0 to 10 KHz).
As integrator we can use several options.
The Q-detector has an asymmetric output, which can be modelled as a current source. The output current flows through a resistor and produces a voltage across the resistor, which can be buffered and used in the system further on. If we connect a capacitor parallel to the resistor, we get for V_{OUT} as function of the output current I:

$$V_{OUT} = \frac{R}{1 + j\omega RC} I \qquad (4.9)$$

For angular frequencies $\omega >> (RC)^{-1}$ the output voltage V_{OUT} is the integrated version of the current I.
For a good low frequency performance the product RC must be as large as possible (and practical). R is limited by the fact that for too great a value of R the distortion (especially at higher frequencies) will become too large. C cannot be made too small because of the low frequency performance. If C is too big, we need much amplification to obtain the correct voltage of 392 mV_{RMS} at 1.326 radians of sine wave phase modulation. Still it is an interesting solution, although it is not used here, because the amplification needed would be too large.
Another possibility is a separate integrator using an operational amplifier (opamp).

Here we have the choice between two possible configurations:

- the non-inverting integrator.

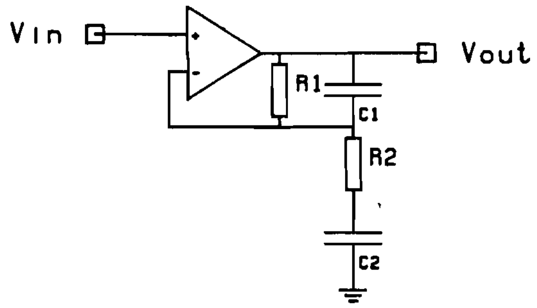

Figure 4.2. The non-inverting integrator.

This integrator does not need an external reference voltage and the amplification is large. A disadvantage is the fact that the amplification is dependent on a timing constant, which determines the highest frequency, for which the circuit still functions as an integrator. For higher frequencies the circuit functions as an amplifier, which means that very high frequencies are not suppressed.
The transfer function is given by:

$$\frac{V_{OUT}}{V_{IN}} = \frac{(1+j\omega R_1C_1)(1+j\omega R_2C_2) + j\omega R_1C_2}{(1+j\omega R_1C_1)(1+j\omega R_2C_2)} \tag{4.10}$$

When $\omega >> (R_1C_1)^{-1}$, $\omega >> (R_2C_2)^{-1}$ and $\omega << (R_2C_1)^{-1}$ the transfer function will reduce to:

$$\frac{V_{OUT}}{V_{IN}} \approx \frac{j\omega R_1C_1\, j\omega R_2C_2 + j\omega R_1C_2}{j\omega R_1C_1\, j\omega R_2C_2} = \frac{1 + 1/(j\omega R_2C_1)}{1} \approx \frac{1}{j\omega R_2C_1} \tag{4.11}$$

Now it is clear that under these conditions the circuit in figure 4.2 is an integrator.

- an inverting integrator (figure 4.3).

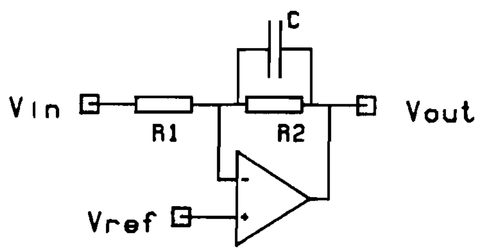

Figure 4.3. The inverting integrator.

This circuit works as a true integrator for angular frequencies $\omega >> (R_2C)^{-1}$. The transfer function is given by:

$$\frac{V_{OUT}}{V_{IN}} = \frac{-R_2/R_1}{1 + j\omega R_2C} \tag{4.12}$$

For $\omega >> (R_2C)^{-1}$ the circuit will perform like an integrator, since the transfer function will reduce to:

$$\frac{V_{OUT}}{V_{IN}} = \frac{-1}{j\omega R_1C} \tag{4.13}$$

The amplification is now determined by R_1C. C can be made small, since R_2 can be made quite large. There is still need for some extra amplification, but not very much (about 20 dB). The main advantage is that the amplification of the integrator can be made independent of the timing constant R_2C.
In the decoder the inverting integrator is used. In appendix C the decoder is shown with the coupling of the quadrature detector and the integrator.

4.2. The correction circuit

The correction circuit is based on the detection of the positive and the negative peak value of the phase. The difference between the two absolute values is determined and used as a correction voltage. Since the positive and the negative peak values do not occur at the same instant, they need to be stored temporarily in order to determine the difference between them.
A circuit that implements this function is given in figure 4.4.

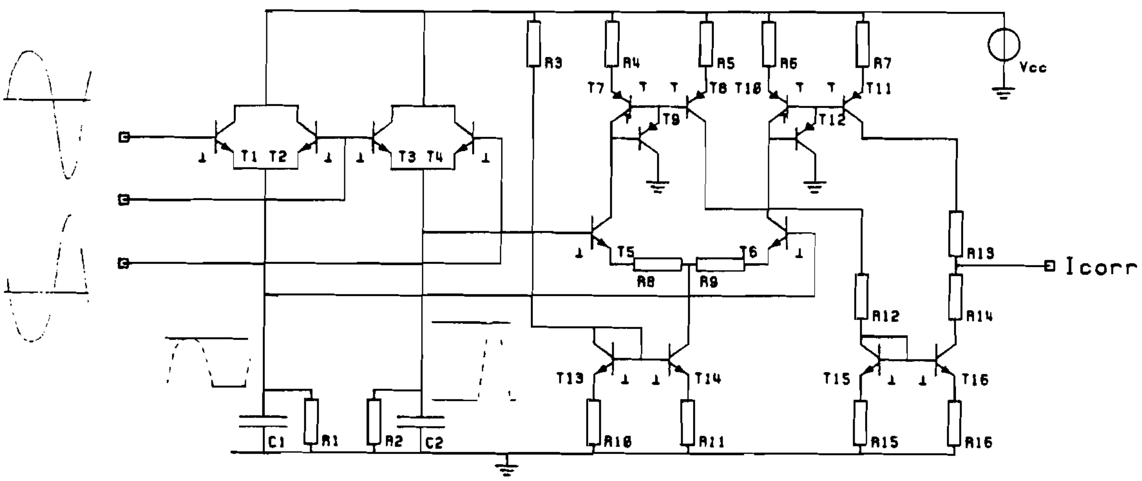

Figure 4.4. A circuit for establishing the difference between positive and negative peaks.

The used circuits are level detectors. The phase appears at the base of transistor T_1. At the base of transistors T_2 and T_3 the DC value of the detected phase is presented. T_1 is blocked until the voltage at its base exceeds the voltage at the base of T_2. This will charge capacitor C_1 until the capacitor is fully charged and the voltage at the base of T_1 decreases. C_1 will slowly discharge itself through resistor R_1, but since the product R_1C_1 is 1 second, this does not happen very quickly. The same story applies to transistors T_4 and T_3. Now the inverted phase is offered at the base of T_4, which means that the negative peak value will be stored. The differential voltage between C_1 and C_2 is then fed to a differential amplifier, which converts the voltage to a current. Thus the current I_{CORR} will be proportional to the difference in positive and negative peak value of the detected phase. Resistors R_{13} and R_{14} are added to give the output approximately the correct DC bias.

4.3. The tangent approximation

The implementation of the tangent function poses a problem, since there is no circuit that performs this function. This means we have to make do with an approximation.
In [1] this topic has been investigated and this resulted in the approximation of the tangent function by the arctanh-function.
If we develope both functions into powers series, we get:

$$\tan(x) = x + \frac{1}{3}x^3 + \frac{2}{15}x^5 + \frac{17}{315}x^7 + \ldots. \tag{4.14}$$

$$\operatorname{arctanh}(x) = x + \frac{1}{3}x^3 + \frac{1}{5}x^5 + \frac{1}{7}x^7 + \ldots. \tag{4.15}$$

For low values of x the arctanh(x) approaches tan(x) reasonably well. For high values of x the approximation is not so good.
A simple implementation of the arctanh is the following circuit (see fig. 4.5):

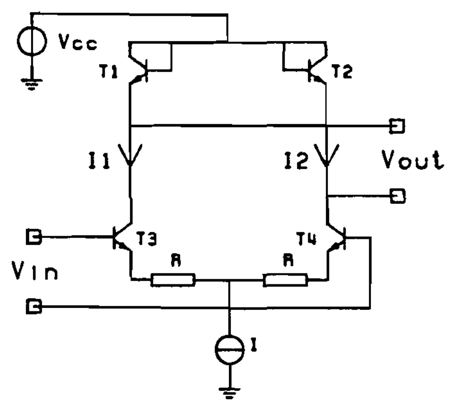

Figure 4.5. Implementation of arctanh(x).

The relation between the input voltage V_{IN} and the output voltage V_{OUT} is given by [1]:

$$V_{OUT} = -\frac{2kT}{q} \operatorname{arctanh}\left[\frac{V_{IN}}{I\,R + 2kT/q}\right] \tag{4.16}$$

where: k = Boltzmann's constant = 1.38×10^{-23} [J/K]
q = elementary charge = 1.6×10^{-19} [C]
T = absolute temperature [K]

Transistors T_1 and T_2 are responsible for the implementation of the arctanh-function. They have to approach the ideal exponential curve of a diode well and therefore they are bigger than the other transistors. This is done to minimise the influence of the base resistance.
It must be noted that expression 4.16 is valid, only if the current variation in I_1 and I_2 is much smaller than current I. Only then transistors T_3 and T_4 perform a good voltage to current conversion.
But when this condition is not entirely met, the approximation of the tangent is improved, due to the extra distortion in the voltage to current converter. In figure 4.6 a comparison of the true tangent and its approximation is shown, using data given by simulations. The values of R and I are: R = 375 Ω, I = 1 mA.

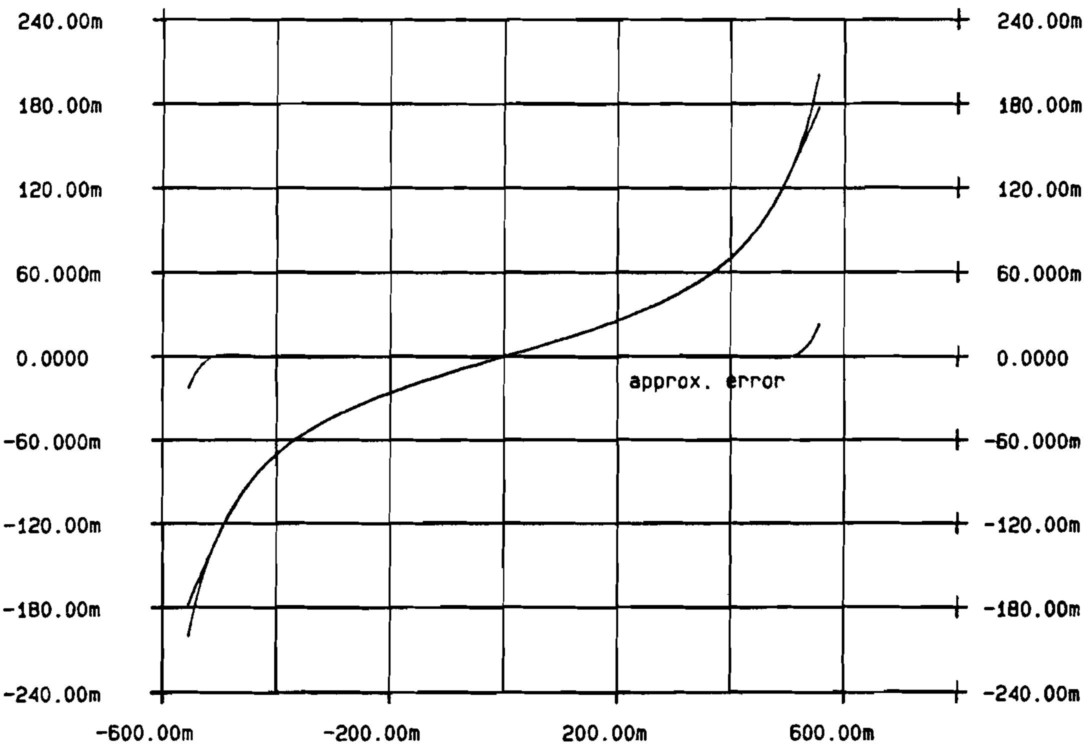

Figure 4.6. Comparison of tan(x) with arctanh(x).

An input voltage of 555 mV equals a phase angle of 1.326 radians.
Figure 4.6 shows that we have realised good tangent approximation; only at extreme phase angles (close to 1.326 radians) the approximation deviates from the desired curve.
The result is that we now have a rather simple circuit to approximate the tangent function with sufficient accuracy.
A disadvantage of the tangent circuit is that it requires a buffer, since a load at the output will distort the tangent curve. A simple buffer can be made using a Darlington configuration (see appendix C for the buffer in the total circuit).

4.4. The multiplier

A simple solution for the multiplier is the well known Gilbert Cell multiplier, as shown in figure 4.7.

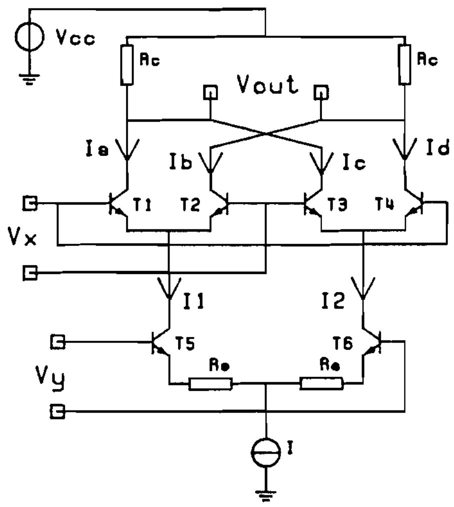

Figure 4.7. The Gilbert Cell multiplier.

The output voltage V_{OUT} as function of V_X and V_Y is given by [9] (see also [1]):

$$V_{OUT} = \frac{R_C}{R_E + 2kT/qI} \tanh(qV_X/2kT)\, V_Y \tag{4.17}$$

For small values of V_X ($|V_X| << 2kT/q$) we can approximate tanh(x) by x. This means that for low values of V_X we have a linear multiplier with a small input range. Since the output of the tangent can be as much as 185 mV, we need to increase the range of V_X by using a circuit that compensates the tanh(x). Such a circuit has been described in the previous paragraph.

The relation between V_{OUT} and V_{IN} of this circuit is given by:

$$V_{OUT} = -\frac{2kT}{q} \operatorname{arctanh}\left[\frac{V_{IN}}{I\,R + 2kT/q}\right] \qquad (4.16)$$

Now we obtain for the total multiplier the following expression:

$$V_{OUT} = \frac{-R_C}{(R_E + 2kT/qI_M)(I_A R_A + 2kT/q)}\, V_X\, V_Y \qquad (4.18)$$

where: R_A = resistor in arctanh-circuit
I_A = current in arctanh-circuit
I_M = current in multiplier

Simulations show that the multiplier is now linear within a range of:

$|V_X| < 0.9$ V and $|V_Y| < 0.9$ V.

This means we can amplify the output of the tangent in order to obtain more signal using a buffer/amplifier. The amplification of this buffer can be as much as 5.

4.5. The audio matrix

In the audio matrix L+R and L-R are combined to give the left and right audio channel. This has to be done accurately since any error will affect the channel separation.
The ideal case would be two identical circuits in which L-R and -(L-R) are added with L+R to give L and R respectively.
The output of the multiplier is symmetric, which means that it gives us both L-R and -(L-R).
The circuit used here as matrix is based on the addition of two currents. This means that L+R, L-R and -(L-R) have to be converted into currents, using current mirrors. The principal of a current mirror can best be explained by figure 4.8:

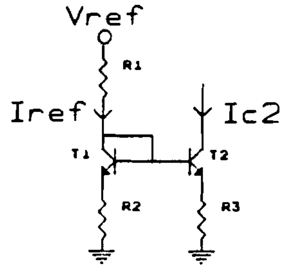

Figure 4.8. A current mirror.

If transistors T_1 and T_2 are identical, the following equation can be derived from figure 4.8:

$$I_{REF} R_2 - I_{C2} R_3 + kT/q \ln(I_{REF}/I_{C2}) = 0 \tag{4.19}$$

When $R_2 = R_3$ then I_{C2} will equal I_{REF}, since $kT/q \ln(I_{REF}/I_{C2})$ is small compared to the other terms in equation 4.19. If V_{REF} contains a DC bias which is much larger than the AC-information, then the voltage drop across T_1 will be constant and the variation in I_{REF} will be proportional to the variation in V_{REF}. In this way we have realised a voltage to current converter.
Now the outputs of two converters are connected and mirrored to a current through a resistor.
This leads to the following circuit for the complete audio matrix (see figure 4.9):

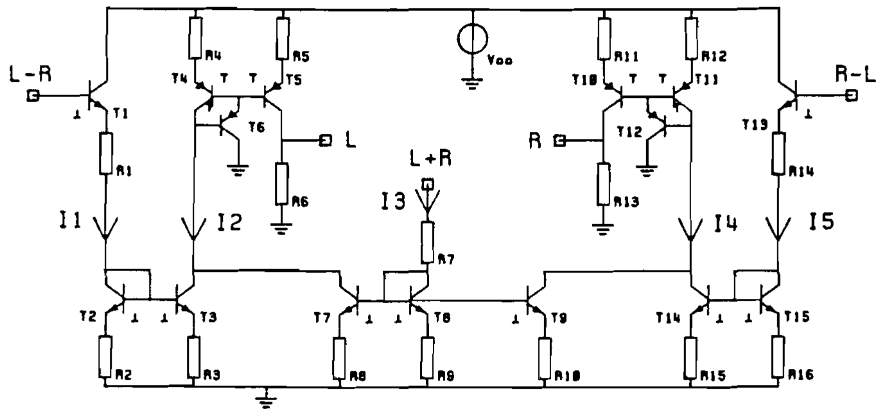

Figure 4.9. The complete audio matrix.

Current I_2 is the sum of I_1 and I_3 and is therefore proportional to L. Current I4 is the sum of I_3 and I_5 and thus proportional to R. This circuit has the advantage that it is totally symmetric. If necessary the addition of L+R can be fine adjusted by varying R_9. The matrix can be easily switched to mono by short circuiting the bases of T_2 and T_{15} to the ground.

4.6. The pilot detection circuit

The task of the pilot tone detection circuit is to make the decision of switching from monaural reception to stereo or vice versa. This can be done by detecting whether there is a pilot tone present or not. In addition it must be possible to switch from stereo to mono on the basis of external conditions. For example when the radio is searching for a station it is wise to switch to mono. Below a certain signal level the reception may be garbled by too much noise, which makes monaural reception preferable to stereo.
Here we will concentrate on the task of accurately detecting the pilot tone.
It is not sufficient to switch to stereo as soon as a pilot tone is detected and switch to mono when the pilot is not present. This will cause serious chattering when the

decoder switches repeatedly from stereo to mono and vice versa whenever the pilot is for example cancelled by program material. This means that a kind of hysteresis has to be incorporated.
There are basically two configurations for the pilot tone detection circuit:
- a digital way (see figure 4.10).

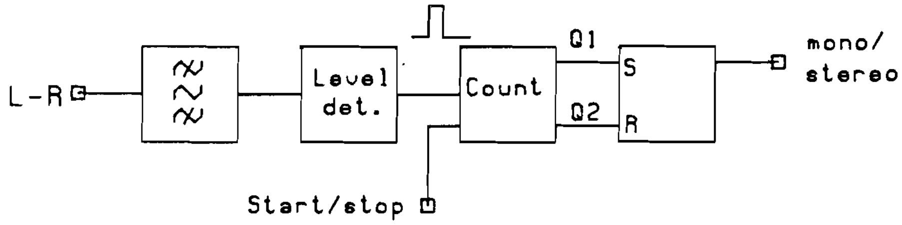

Figure 4.10. A digital treatment of the pilot tone.

The pilot tone is filtered from L-R and is passed onto a level detector to check whether the pilot exceeds a certain threshold. If it does, then it is shaped into a pulse. During a certain interval the number of pulses is counted and if this number exceeds a predetermined value, the decoder will switch to stereo. When in stereo mode the number of pulses crosses a lower bound (i.e. not enough pulses have arrived) the decoder will switch to mono. This method is used in [10]. The main disadvantage of this system is the need for an interval timer, which is usually derived from the carrier.
A somewhat different solution, which does not need an interval clock, is the one chosen by Motorola (see figure 4.11).

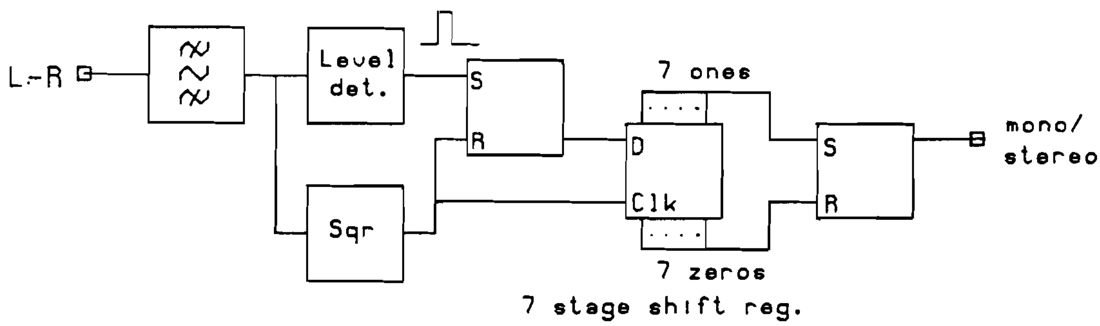

Figure 4.11. The Motorola pilot tone detection.

The idea is basically the same: when a number of consecutive good cycles of the pilot tone is detected, the decoder switches to stereo. When a number of consecutive bad cycles is detected then the decoder switches to mono. The question is now: what is a good cycle and what is a bad one?
Whenever a tone of the right frequency (25 Hz) is detected, it is shaped into a clock pulse (by the block Sqr in figure 4.11). It is also checked if the level of the pilot is correct. If it is correct, then we have a so called good cycle. If the level is not correct then we have a bad cycle. In this way whenever 25 Hz is present it is checked whether it is a pilot tone or not. This means that the pilot tone detection can be fooled: when it is in stereo mode and the pilot is

put off, the decoder will remain in stereo mode until a false pilot tone is detected, which can be caused by the modulation, containing 25 Hz or by noise for example.

- an analogue way (see figure 4.12).

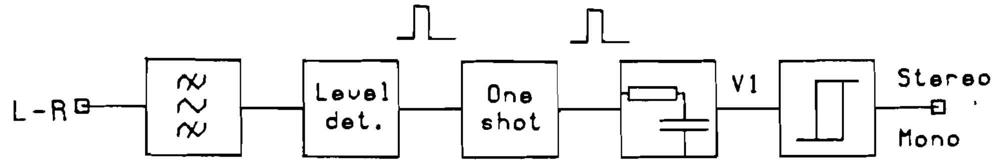

Figure 4.12. Analogue pilot tone detection.

Another method of treating the pilot tone is in an analogue way. Now a capacitor is used as memory instead of a counter.
The pilot tone is also passed through a filter and a level detector. Each time the level is correct a capacitor is charged. When a number of consecutive cycles are correct, the capacitor voltage will exceed a certain threshold and the decoder will switch to stereo. When the voltage drops below a certain voltage when in stereo mode, the decoder will switch to mono. This solution is less complex, but it requires some external components, such as a capacitor. Another problem is that the pulse, coming from the level detector will not always have the same width. This means that the capacitor will not always be charged in the same amount of time. It is therefore necessary to include a "one shot" circuit (a monostable multivibrator) which makes the pulse into a pulse with a constant width. A comparator with hysteresis will then decide whether to switch to stereo or mono.

The method used is the analogue method. A major advantage is the fact that it does not require many components, although it needs two external capacitors, apart from the band-pass filter. Furthermore it is easy to build.
As described above the detection circuit consists of a band pass filter, a level detector followed by a monostable multivibrator. This is followed by a circuit, that allows the charging of a capacitor during a certain pulse. The voltage across this capacitor is then used as input of a comparator with hysteresis. This comparator has two thresholds:

- the upper threshold will cause the output of the comparator to switch from low to high when the input voltage is increased and passes this threshold.
- the lower threshold will cause the output of the comparator to switch from high to low when the input voltage is decreased and passes this threshold.

The used circuitry here is implemented mainly using operational amplifiers, since there was not enough time to design the blocks with simple transistors. In appendix B the complete pilot detection circuit is described in more detail.
Adding an external mono switch is simple to do by placing a switch over the capacitor. By short circuiting the capacitor to earth with a transistor switch it is possible to switch the decoder from stereo to mono.

5. THE DECODER'S PERFORMANCE

5.1. Performance analysis

When we look at the expression for the C-QUAM signal and at the structure of the Motorola decoder we see that there is no theoretical limit to the performance of the C-QUAM system. The performance of the C-QUAM system only depends on the limitations of the practical implementation of the total system from encoder, transmitter to decoder.
The situation for the alternative decoder is somewhat different, since we needed a correction circuit to overcome a fundamental problem. This means that the performance is now limited by the used method to regenerate the missing DC value in the phase of the C-QUAM signal.

5.1.1. Possible sources of errors

As stated earlier, the performance also depends on the limitations of the practical implementation. When we look again at the structure of the alternative decoder, it is possible to identify some possible errors:

- the level at the input of the tangent is not correct. This causes a non-linear error.
- a phasing error between L+R and (L-R)/(1+L+R) since these two signals should arrive at the inputs of the multiplier at exactly the same time, but they travel along different routes.
- the matching of 1 to L+R in order to make 1+L+R at the input of the multiplier may not be as accurate as desired.
- the matching of L+R and L-R in the audio matrix is not correct. This error is also a representation of deviations in calculated amplifications, since these deviations will eventually result in a mismatch in level between L+R and L-R.

The first two errors are not easily described and can best be simulated in order to establish their influence on distortion and channel separation. In appendix E the results of these simulations are shown.
The second two errors can be analysed mathematically without too much trouble.
When we look at a possible error through a mismatch of 1 and L+R we will assume that only the left channel contains information:

$$L(t) = A(t)$$
$$R(t) = 0$$

where A(t) is an arbitrary function with the following property:

$$|A(t)| \leq 1 \text{ for all } t.$$

The time variable t will be ignored in future, but tacitly assumed to be present.

In the addition of 1 and L+R an error δ is introduced. If the demodulation introduces no further errors, we can write for the restored left channel L' and right channel R':

$$L' = (L+R) + (1+\delta+L+R) \times \frac{L-R}{1+L+R} = A + (1+\delta+A)\frac{A}{1+A} \tag{5.1}$$

$$R' = (L+R) - (1+\delta+L+R) \times \frac{L-R}{1+L+R} = A - (1+\delta+A)\frac{A}{1+A} \tag{5.2}$$

These expressions can be derived from figure 3.1.
The cross talk from the left to the right channel related to the left channel is then given by:

$$\frac{R'}{L'} = \frac{A - (1+\delta+A)\, A\,/\,(1+A)}{A + (1+\delta+A)\, A\,/\,(1+A)} = \frac{-\delta}{2 + 2A + \delta} \tag{5.3}$$

When $|\delta| << 1$ the above expression may be written as:

$$\frac{R'}{L'} = \frac{-\delta}{2\,(1+A)} \tag{5.4}$$

For low values of A and for $A>0$ the error will be relatively small. However when A approaches -1, which may happen at high modulation when the modulation index comes close to 100%, the error will have big consequences, although δ may still be small. One consequence will be that the positive amplitude of L' will be unequal to the negative amplitude. This causes large distortion to the desired L'.
This means that the error δ must be made as small as possible.
When the matching of L+R and L-R in the audio matrix is not exact then this will cause a reduction in channel separation. The distortion will not increase, since this is a purely linear error.
When such an error is expressed in a factor $1+\delta$, this leads to similar expressions for L' and R' with $L=A$ and $R=0$:

$$L' = (L+R) + (1+\delta)(1+L+R)\frac{L-R}{1+L+R} = A + (1+\delta)\,A \tag{5.5}$$

$$R' = (L+R) - (1+\delta)(1+L+R)\frac{L-R}{1+L+R} = A - (1+\delta)\,A \tag{5.6}$$

The cross talk is now given by:

$$\frac{R'}{L'} = \frac{A - (1+\delta)A}{A + (1+\delta)A} = \frac{-\delta}{(2+\delta)} \approx -\tfrac{1}{2}\delta \qquad (5.7)$$

This means that the error is not so large and independent of the amplitude of L
An error of about 5% will reduce the maximal separation to about 32 dB.

5.1.2. Sensitivity to temperature variations

Another source of concern is the temperature dependency of the circuits used. In an car radio the temperature may vary over a wide range. Here the decoder must function properly in a temperature range from minus 20° celsius to 85° celsius. It is now important to investigate in what way temperature variations effect the functioning of the decoder and how this can be compensated. In the analysis we will concentrate our attention on the circuitry after the IF-amplifier/limiter and phase detector.
When we look at the expression for the tangent it is clear that the temperature has a great influence, as can be seen in expression 4.15 here below.

$$V_{OUT} = -\frac{2kT}{q} \operatorname{arctanh}\left[\frac{V_{IN}}{I\,R + 2kT/q}\right] \qquad (4.16)$$

For a proper approximation of the tangent function, the argument of the above expression should remain constant when the temperature is varied. This can be done by making the input voltage V_{IN} as well as the current I proportional to the absolute temperature. This causes the numerator and the denominator to be equally proportional to the temperature.
What remains is the fact that the amplification contains the absolute temperature.
The buffer after the tangent is almost independent from the temperature, since the amplification is R_C/R_E, because $I\,R_E >> kT/q$.
The expression for the multiplier is given by:

$$V_{OUT} = \frac{-R_C}{(R_E + 2kT/qI_M)(I_A R_A + 2kT/q)} V_X V_Y \qquad (4.18)$$

Since $R_E >> 2kT/qI_M$ and $I_A R_A >> 2kT/q$, V_{OUT} is independent of the temperature, but inversely proportional to the current I_A.
The result is now that the output of the multiplier is directly proportional to the absolute temperature. If we make I_A in the multiplier proportional to the temperature, then it is possible to compensate the temperature dependency completely.

In the mean time we have ignored the temperature dependence of 1+L+R at the input of the multiplier and L+R at the input of the audio matrix.
Fortunately the temperature dependence of 1+L+R and L+R is not important as long as they are exactly the same.
This can be explained by the fact that the audio matrix is symmetric. The voltage to current converters are identical for L+R, L-R and -(L-R) (see paragraph 4.5). Therefore any temperature variation will have the same effect on all three signals. If L-R and L+R have the same temperature dependence, the net effect will be negligible, because the difference between the two will remain the same.
From the analysis of the decoder it has been shown that the only remaining temperature effect on L-R will be caused by the temperature dependence of 1+L+R. If this dependence is equal to the one of L+R, then it is clear that the net result is negligible.
Now we have reduced the temperature effect to nil by the following measures:

- the currents in the tangent and the arctanh-circuit in the multiplier as well as the input voltage of the tangent must be made proportional to the absolute temperature.
- the temperature dependence of 1+L+R and L+R must be identical.

Some small modifications must be made in the total circuit in appendix C, because the DC values at several point in the circuit are a function of several currents. In order to keep them at approximately the correct level, it is necessary to use a level shift circuit instead of pure resistors.

5.2. Measured performance

In chapter four the functional blocks of the decoder have been described. These blocks have been added together and with breadboard components a working model of the decoder has been made. The used process for the breadboard components is a Bimos process, used in Nijmegen. Actually the process is not critical, since only low frequencies are being used here, although it is important to use good diodes in the tangent approximation. In appendix C the circuit diagram of the decoder is given, excluding the IF-amplifier/limiter and Q-detector. The output resistor of the Q-detector is here 5.6KΩ. There is also a capacitor of 0.22nF connected from the output to the ground.
The AM output of the IF-amplifier/limiter has also a capacitor of 1nF connected to the ground. The values here are chosen in such a way that the phase difference between $\phi(t)$ and L+R is minimal. In the circuit several resistors are used as level shifters in order to ensure a correct operation of the total system. This means that these circuits are fed from a supply, followed by a resistor, instead of a supply only. There are at the moment two variable resistors in the decoder. One is used to adjust L+R in the audio matrix. The other is used to adjust L+R to 1 at the input of the multiplier. The power supply voltage should be 8.5V. The complete circuit has been built on a printed circuit board, using breadboard components.

5.2.1. The breadboard model without IF filters

To see what the performance of the alternative decoder is, it has been tested, using a C-QUAM signal generator which is connected with the input of the IF-amplifier/limiter. The following measurements have been carried out:

- harmonic distortion THD as function of the modulation index at 400 Hz (figure 5.1A) and 1 KHz (figure 5.1B).
- channel separation α as function of the modulation index m at 400 Hz (figure 5.2A) and 1 KHz (figure 5.2B).
- channel separation α as function of the modulation frequency f_M at m=0.3 (figure 5.3).

These measurements have been carried out with a C-QUAM signal with a carrier frequency of 450 KHz and a level of 106 dB EMF for both Left only and Right only modulation. The results can be seen in figures 5.1-5.3.

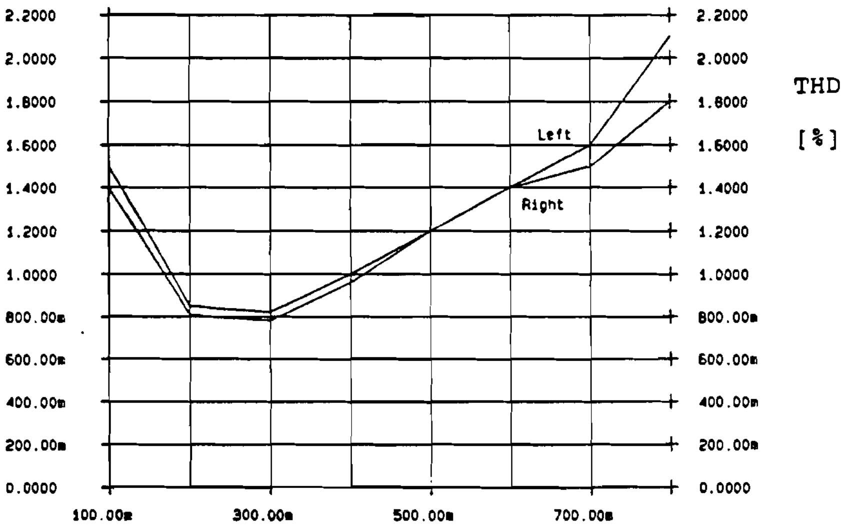

Figure 5.1A. Harmonic distortion (THD) vs. modulation index m at 400 Hz.

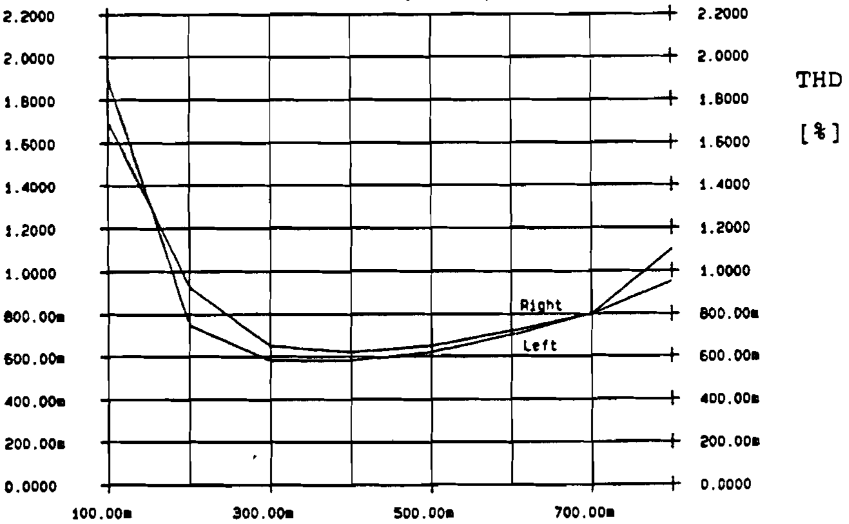

Figure 5.1B. Harmonic distortion (THD) vs. modulation index m at 1 KHz.

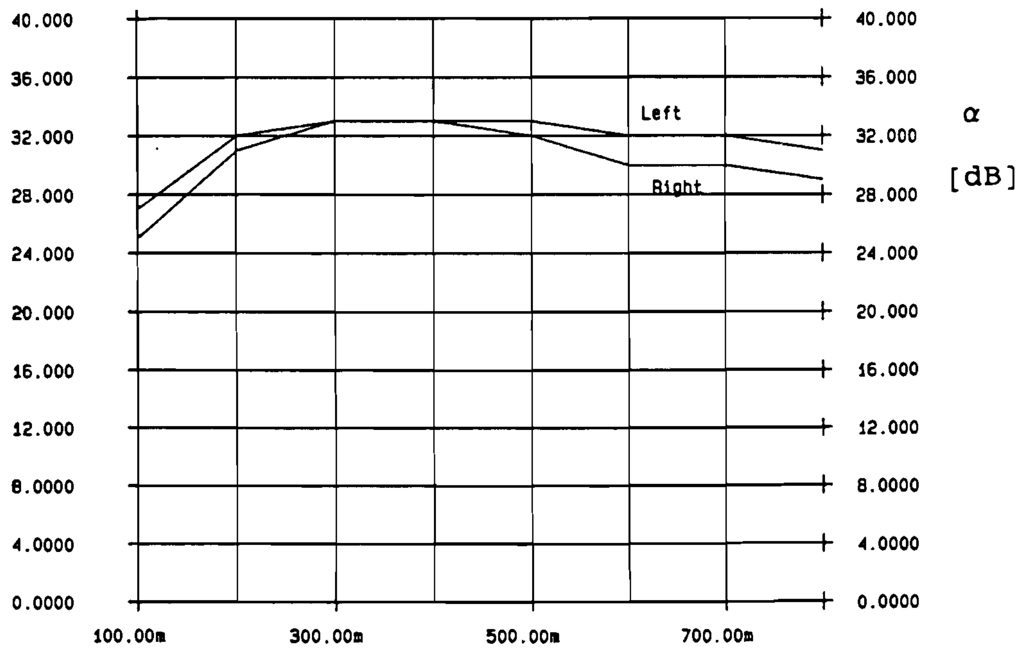

Figure 5.2A. Channel separation α vs. modulation index m at 400 Hz.

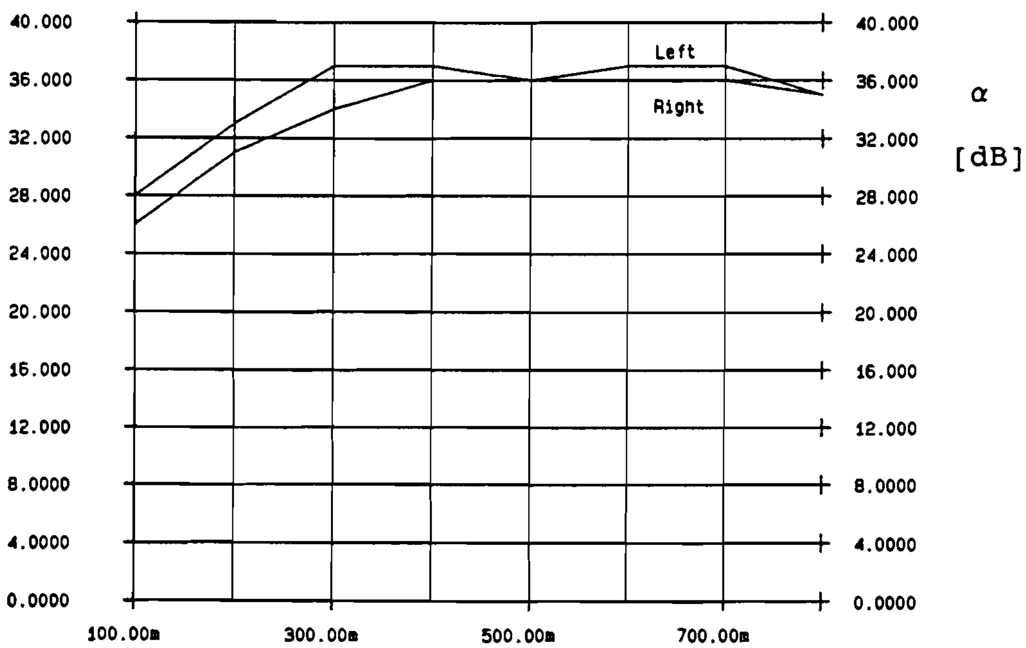

Figure 5.2B. Channel separation α vs. modulation index at 1 KHz.

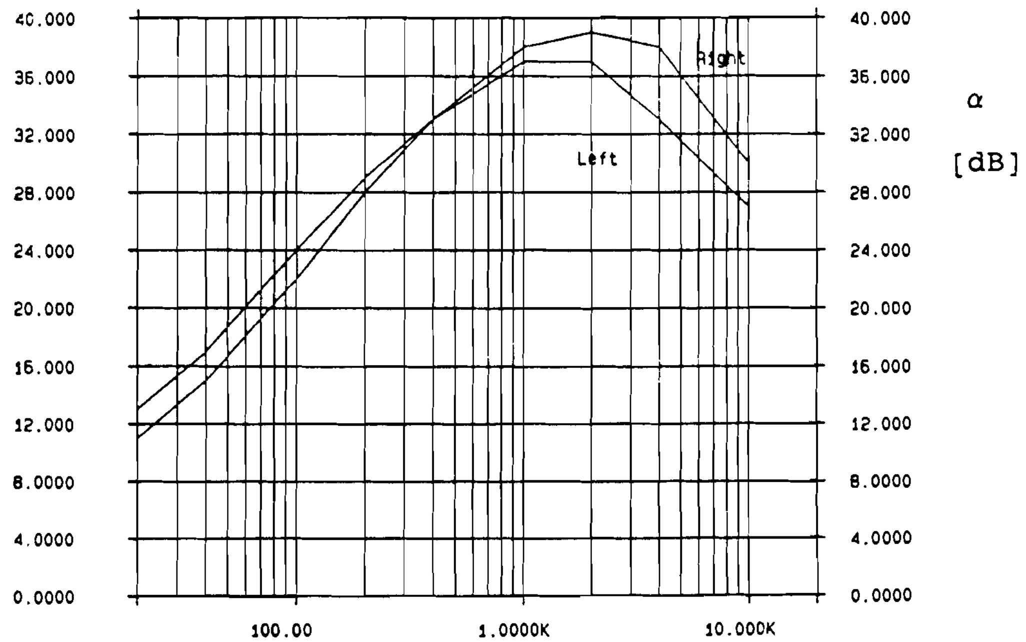

Figure 5.3. Channel separation α vs. modulation frequency f_M at m=0.3.

The distortion is acceptable; for m=0.8 at 1 KHz the distortion is only about 1%. Without the correction circuit this figure would be around 5-7%.
The channel separation as function of the modulation index is also good. The difference between the situation at 400 HZ and 1 KHz can be caused by the fact that the timing constant of the correction circuit is not large enough. This does not have to be a problem since the stereo information is mostly present at higher frequencies (f > 300 Hz).
From figure 5.3 it is clear that the decoder is highly competitive with the Motorola decoder. In appendix D.1 the results of a measurement are presented, in which the output level of the C-QUAM signal generator is varied. Appendix D.1 shows that the breadboard model functions well. There is only a slight variation in channel separation as function of the input level. This is caused by the fact that the AM-output is not constant as a function of the input signal level. The noise level lies around 57 dB.

5.2.2. The breadboard model in a radio environment

Since the decoder is to be used in a radio, it is also important to see how the performance is degraded by the front-end and the IF-filters in the radio. therefore the decoder has been connected to the IF of a Philips 90DC788 car radio.
In this configuration the following measurements have been performed as function of the input signal level:
- signal-to-noise ratio at m=0.3
- channel separation at m=0.3
- harmonic distortion at m=0.8.

These measurements have been carried out in stereo mode. The results can be seen in appendix D.2.
The results in appendix D.2 show that there is a certain threshold in the input signal level, above which the performance is fairly good and below which the performance decreases abruptly. It appears that this threshold level coincides with the level above which a feedback loop in the front end of the radio is activated. This feedback keeps the output signal of the front end constant for strong signals. The threshold level is about 15 mV (EMF) at the antenna input.
Above this threshold the performance of the decoder is not too bad, although the difference between L-only and R-only is rather large. A harmonic distortion figure of 6% for R-only compared to 2.6% for R-only is not good. The channel separation is 30 dB. The signal-to-noise ratio is 44 dB.
This threshold can also be found in the performance of the Motorola decoder as function of the input signal level in appendix A. The channel separation is reduced by only 2 dB, compared to 5-6 dB for the alternative decoder. This means that the Motorola decoder is not as sensitive as the breadboard model.
Comparing appendix A with appendix D.2 it is clear that the alternative decoder scores better on channel separation but worse on distortion. The distortion figures of the breadboard model for L-only are not acceptable.
When searching for an explanation of the difference between L-only and R-only an explanation is provided by the fact that the phase of the C-QUAM signal undergoes more distortion for R-only than L-only. On a spectrum analyser it can be seen that the lower harmonic components do not have the same magnitude as the upper harmonic components of the signal. This means that the signal undergoes asymmetric distortion. Even pure sine wave phase modulation undergoes asymmetric distortion when the modulation level is about 1 radian or more.
Asymmetric phase distortion will have serious consequences for the performance of the decoder, since the correction circuit for the phase is based on an adequate detection of the phase. It uses secondary properties of the phase for the regenerating of the missing DC component. When the phase of the signal is distorted the correction will not be done well. Thus the performance will be affected more seriously in the alternative decoder than in the Motorola decoder.

6. CONCLUSIONS

The objective of this graduation project was to realise an alternative decoder for C-QUAM AM stereo signals in a different way than Motorola has done, using the IF-amplifier/limiter developed at PCALE. Furthermore the total concept of radio with stereo decoder should be at least equal in performance and cheaper than a radio using the Motorola decoder.
The method for the decoding of C-QUAM followed here is based on the simultaneous demodulation of the phase and amplitude information.
It has been shown that it is theoretically impossible to do this perfectly, since the phase of the C-QUAM signal can - under certain conditions - contain a DC component which cannot be detected, although it is possible by using secondary properties of the detected phase to restore the missing DC component.
A complete breadboard model of the alternative decoder has been developed and tested. It incorporates a circuit, which generates a DC component out of secondary properties of the phase. This DC is then added to the detected phase. This reduces the distortion of the decoder dramatically under ideal conditions (i.e. without IF filtering). Now the decoder performs better than the Motorola decoder on channel separation and equal on distortion at high modulation levels. The signal-to-noise ratio of the decoder is unfortunately 3 dB worse. The breadboard model contains several points which need to be carefully realised. One is the precise matching of 1 and L+R in order to make 1+L+R at the multiplier input. Another point is the matching of L+R and L-R in the audio matrix. A 5% mismatch will decrease the channel separation to a maximum of 32 dB. The accuracy of the phase level needs to be within less than 5% (see Appendix E.1). Another problem which can be circumvented by using good opamps, is the fact that the phase should not contain any DC offset at the input of the tangent. An offset will cause different performance figures for L-only and R-only. An offset of a few millivolts is still acceptable.
When used in a radio the decoder's performance is reduced dramatically. The performance is no longer symmetric. For R-only the performance remains acceptable but for L-only the distortion becomes too large (about 6%). The channel separation remains better than the separation of the Motorola decoder.
The distortion is mainly caused by asymmetric filters, who provide the selectivity of the radio. The front end is also partly responsible for the extra distortion at lower signal levels.
The pilot detection circuit functions reasonably well, although it is more susceptible to pilot falsing than a digital solution. This is caused by the fact that a strong and broad pulse might extend the charging period of the capacitor to more than the calculated period. This will cause the capacitor to be charged too quickly and the following level detector will switch the decoder to stereo, although no stereo broadcasting takes place. The present detection circuit is not very complex but requires two external capacitors. It will switch to mono as soon as the pilot is gone for a certain period of time. This is not the case for the pilot detection circuit in the Motorola decoder.
The overall conclusion now must be that the decoder is not competitive with the Motorola decoder, since the distortion figures are not good enough.

7. RECOMMENDATIONS

The main cause of the extra distortion can be found in the fact that the phase of the C-QUAM signal suffers distortion due to IF filter asymmetry. The alternative decoder is more susceptible to this kind of distortion than the Motorola decoder. This distortion may be reduced by using more symmetric filters. Furthermore it is important to investigate the distortion caused by the front end, since this is also an important factor.
Still it remains to be seen, whether the low cost of the alternative decoder is enough compensation for the extra costs of improving the front end and the IF filters.
Another possible solution might be to use the zero crossings as a measure for the missing DC component. The thus achieved curve can be shaped by a non-linear device in order to get the desired approximation of the DC component in the phase. This shaping might prove to be somewhat difficult.
The advantage of this method is that it is not susceptible to the phase distortion caused by the IF filters and front-end, since only the high peaks of the phase are affected but not the zero crossings.
The procedure for realising this circuit is the following:

- first we have to convert the phase into a square wave
- next we establish the average value of the square wave
- the average value is then shaped into the desired curve
- the resulting DC voltage is then added to the phase.

The conversion of the phase into a square wave (see paragraph 3.3.2) can be done by a limiter. When we compare figures 3.3 and 3.4, it is possible to describe how the shaping function should work. The shaping function should attenuate at a low modulation index and give extra amplification at a high modulation. The addition of the resulting DC to the phase can be done in a similar way as it is done in the present decoder. In this way it is possible to realise an alternative correction circuit that is less susceptible to phase distortion than the present circuit.

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APPENDIX A: Data on the Motorola decoder

- Appendix A.1: data sheets of the decoder
- Appendix A.2: Measurements of the decoder

MOTOROLA
SEMICONDUCTORS
P.O. BOX 20912 • PHOENIX, ARIZONA 85036

MC13020P

Advance Information

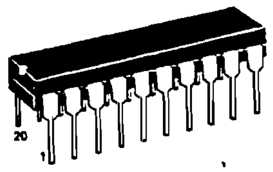
MOTOROLA CQUAM® AM STEREO DECODER

This circuit is a complete one-chip full-feature AM stereo decoding and pilot detection system. It employs full-wave envelope signal detection at all times for the L + R signal, and decodes L−R signals only in the presence of valid stereo transmission.

- No Adjustments, No Coils
- Few Peripheral Components
- True Full-Wave Envelope Detection for L+R
- PLL Detection for L−R
- 25 Hz Pilot Presence Required To Receive L−R
- Pilot Acquisition Time 300 ms For Strong Signals, Time Extended For Noise Conditions To Prevent "Falsing"
- Internal Level Detector Can Be Used As AGC Source

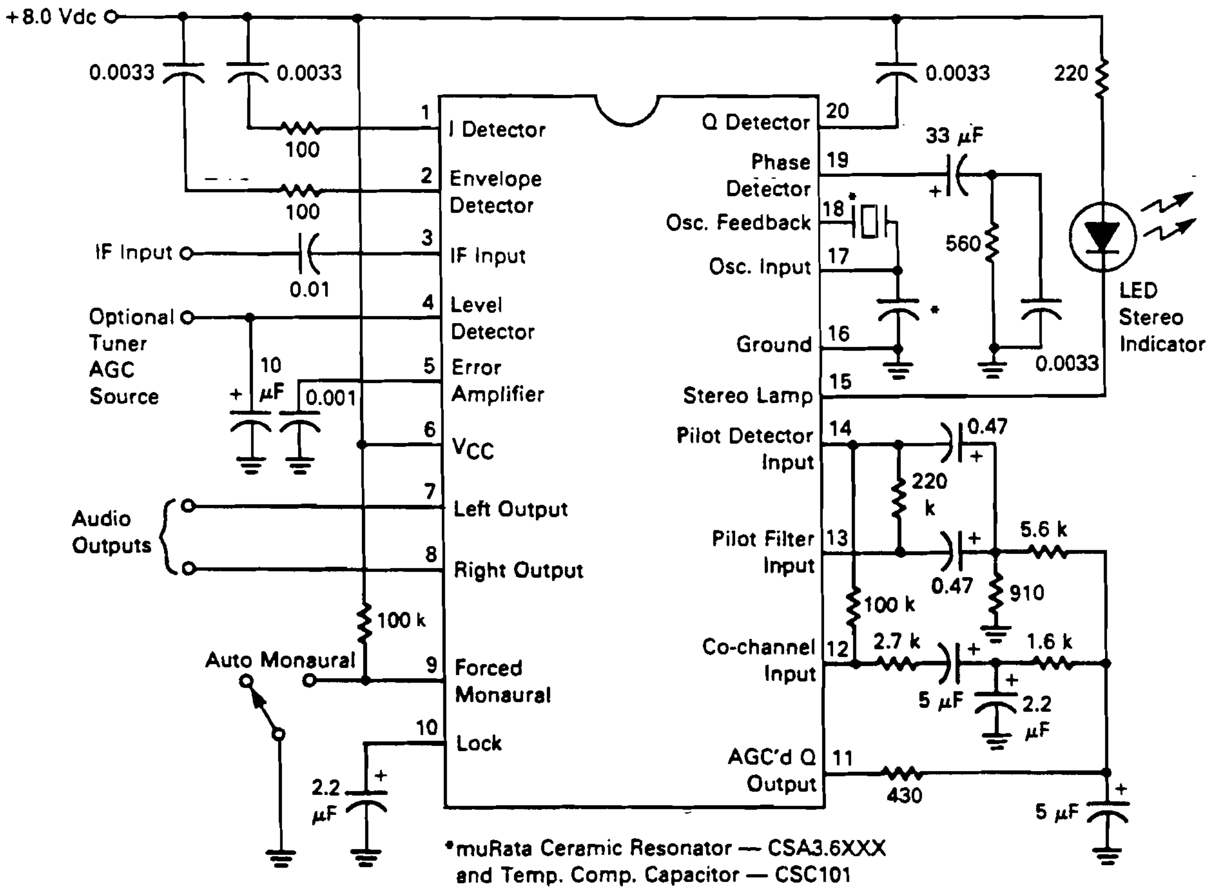
MOTOROLA CQUAM®
AM STEREO
DECODER

SILICON MONOLITHIC
INTEGRATED CIRCUIT



P SUFFIX
PLASTIC PACKAGE
CASE 738-01

FIGURE 1 — TYPICAL APPLICATION



*muRata Ceramic Resonator — CSA3.6XXX
and Temp. Comp. Capacitor — CSC101

is document contains information on a new product. Specifications and information herein is subject to change without notice.

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ADI-725
(Replaces NP144)

MAXIMUM RATINGS

Rating	Symbol	Value	Unit
Supply Voltage	V_{CC}	14	Vdc
Pilot Lamp Current, Pin 15		50	mAdc
Operating Temperature	T_A	−40 to +85	°C
Storage Temperature	T_{stg}	−65 to +150	°C
Junction Temperature	$T_{J(max)}$	150	°C
Power Dissipation Derate above 25°C	P_D	1.25 10	W mW/°C

ELECTRICAL CHARACTERISTICS (V_{CC} = 8.0 Vdc, T_A = 25°C, Circuit Of Figure 1 Unless Otherwise Noted.)

Characteristic	Min	Typ	Max	Unit
Power Supply Operating Range	6.0	8.0	12.0	Vdc
Supply Line Current Drain, Pin 6	—	30	—	mAdc
Input Signal Level, Unmodulated, Pin 3	—	200	350	mVRMS
Audio Output Level, 50% Modulation, L only or R only	—	220	—	mVRMS
Audio Output Level, 50% Modulation, Monaural	—	110	—	mVRMS
Output THD Monaural Stereo	— —	0.5 1.0	— —	%
Channel Separation	—	30	—	dB
Pilot Acquisition Time	—	300	—	ms
Input Impedance R_{in} C_{in}	20 —	27 6.0	— —	kΩ pF
Output Impedance	—	100	150	Ω
Level Detector Filter Voltage, Pin 4, 0 signal 200 mVRMS Signal	— —	1.7 2.5	— —	Vdc
Lock Detector Filter Voltage, Pin 10 In Lock Out of Lock	— —	4.3 0.8	— —	Vdc
Force to Monaural, Pin 9, Pull Down for Monaural Mode	— —	<2.5 150	— —	Vdc nA
Force to Monaural, Pin 9, Pull Up for Automatic Mode	— —	>3.5 <1.0	— —	Vdc nA

FIGURE 2 — BASIC QUADRATURE AM (QUAM)

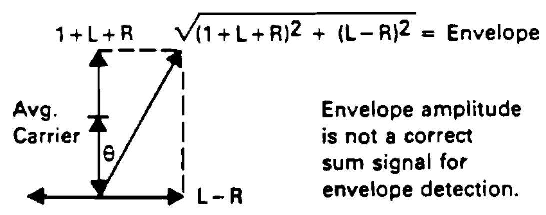
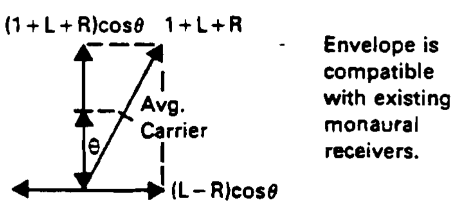
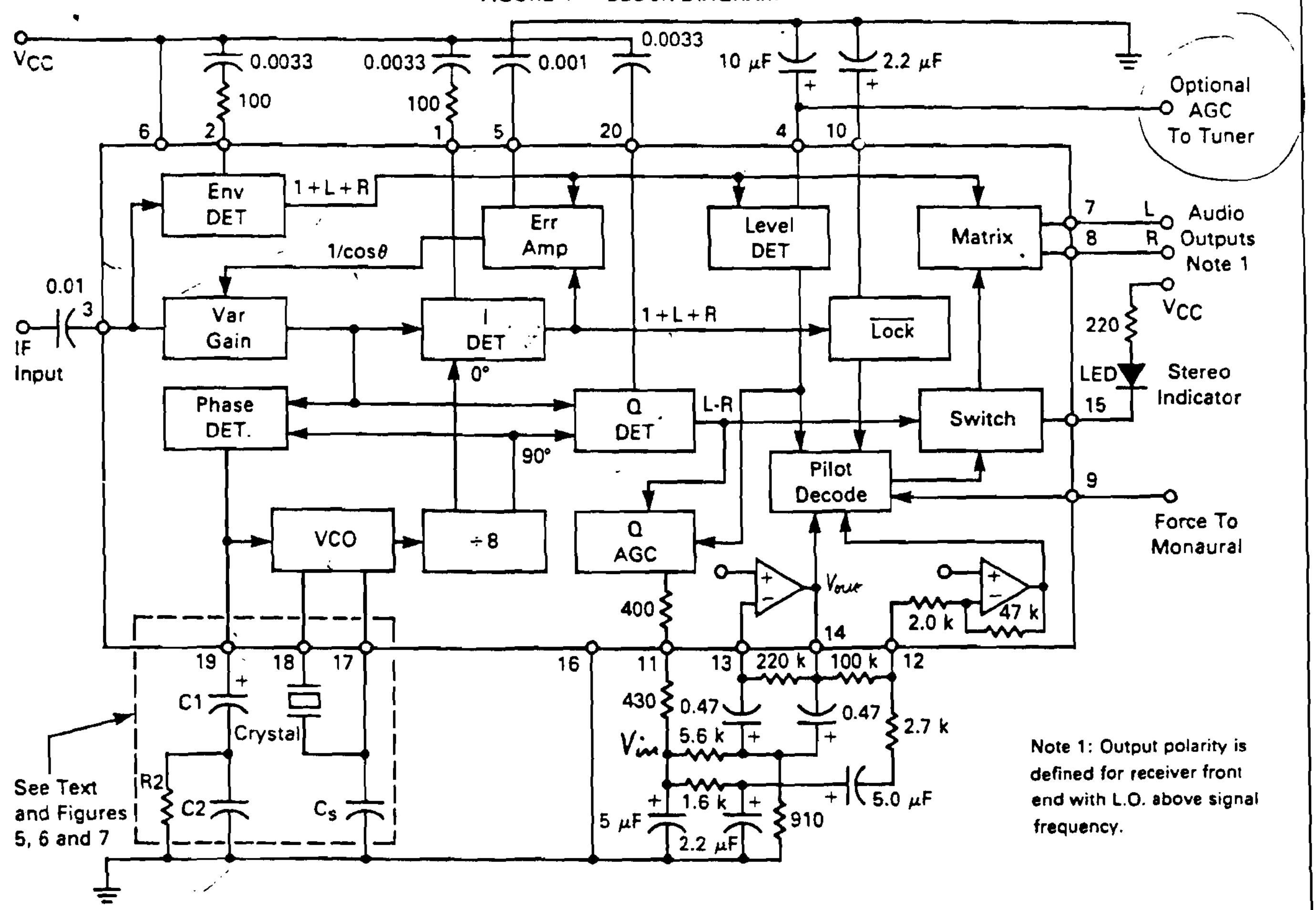


FIGURE 3 — MOTOROLA CQUAM®



$\Theta \neq \omega_c t - \varphi$

FIGURE 4 — BLOCK DIAGRAM



MOTOROLA CQUAM® — COMPATIBLE QUADRATURE AM STEREO

INTRODUCTION

In CQUAM®, conventional quadrature amplitude modulation has been modified by multiplying each axis by cosθ as shown in Figures 2 and 3. The resulting carrier envelope is 1+L+R, i.e., a correct sum signal for monaural receivers and for stereo receivers operating in monaural mode. A 25 Hz pilot signal is added to the L−R information at a 4% modulation level.

THE DECODER

The MC13020P takes the output of the AM IF amplifier and performs the complete CQUAM® decoding function. In the absence of a good stereo signal, it produces an undegraded monaural output. Note in Figure 4 that the L+R information delivered to the output always comes from the envelope detector (Env DET).

The MC13020P decodes the stereo information by first converting the CQUAM® signal to QUAM, and then detecting QUAM. The conversion is accomplished by comparing the output of the Env DET and the I DET in the Err AMP. This provides the 1/cosθ correction factor, which is then multiplied by the CQUAM® incoming signal in the Var Gain block. Thus, the output of the Var Gain block is a QUAM signal, which can then be synchronously detected by conventional means. The I and Q detectors are held at 0° and 90° relative demodulation angles by reference signals from the phase-locked, divided-down VCO. The output of the I DET is 1+L+R, with the added benefit (over the Env DET) of being able to produce a negative output on strong co-channel or noise interference. This is used to tell the Lock circuit to go to monaural operation. The output of the Q DET is the L−R and pilot information.

THE VCO

The VCO operates at 8 times the IF input frequency, which ensures that it is out-of-band, even when a 260 kHz IF frequency is used. Typically a 450 kHz IF frequency is used with synthesized front ends. This places the VCO at 3.6 MHz, which permits economic crystal and ceramic resonators. A crystal VCO is very stable, but cannot be pulled very far to follow front-end mistuning. Pull-in capability of ±100 Hz at 450 kHz is typical, and de-Q-ing with a resistor (see Figure 7) can increase the range only slightly. Therefore, the crystal approach can only be used with very accurate, stable front-ends. By comparison, ceramic and L−C VCO circuits offer pull-in range in the order of ±2.5 kHz (at 450 kHz). Ceramic devices accurate enough to avoid trimming adjustment can be obtained with a matched capacitor for Cs (see Figures 1 and 5).

In the PLL filter circuit on Pin 19, C1 is the primary factor in setting a loop corner frequency of 8–10 Hz, in-lock. An internally controlled fast pull-in is provided. R2 is selected to slightly overdamp the control loop, and C2 prevents high frequency instability.

The Level DET block senses carrier level and provides an optional tuner AGC source. It also operates on the Q AGC block to provide a constant amplitude of 25 Hz pilot at Pin 11, and it delivers information to the pilot decoder regarding signal strength.

PILOT AND CO-CHANNEL FILTERS

The Q AGC output drives a low pass filter, made up of 400 Ω internal, and 430 Ω and 5 μF external. From this point, an active 25 Hz band-pass filter is coupled to the Pilot Decoder, Pin 14, and another low-pass filter is connected to the Co-channel Input, Pin 12. A 2:1 reduction of 25 Hz pilot level to the Pilot Decode circuit will cause the system to go monaural, with the components shown. Refer to Figure 8 for the formulas governing the active band-pass filter. The co-channel input signal contains any low frequency intercarrier beat notes, and, at the selected level, prevents the Pilot Decode circuit from going into stereo. The co-channel input, Pin 12, gain can be adjusted by changing the external 2.7 k resistor. The values shown set the "trip" level at about 7% modulation. The 25 Hz pilot signal at the output of the active filter is opposite in phase to the pilot signal coming from the second low-pass filter. The 100 k resistor from Pin 14 to Pin 12 causes the pilot to be cancelled at the co-channel input. This allows a more sensitive setting of the co-channel trip level.

THE PILOT DECODER

The Pilot Decoder has two modes of operation. When signal conditions are good, the decoder will switch to stereo after 7 consecutive cycles of the 25 Hz pilot tone. When signal conditions are bad, the detected interference changes the pilot counter so as to require 37 consecutive cycles of pilot to go to stereo. In a frequency-synthesized radio, the logic that mutes the audio when tuning can be connected to Pin 9. When this pin is held low it holds the decoder in monaural mode and switches it to the short count. This pin should be held low until the synthesizer and decoder have both locked onto a new station. A 300 ms delay should be sufficient. If the synthesizer logic does not provide sufficient delay, the circuit shown in Figure 9 may be added. Once Pin 9 goes high, the Pilot Decoder starts counting. If no pilot is detected for seven consecutive counts, it is assumed to be a good monaural station and the decoder is switched to the long count. This reduces the possibility of false stereo triggering due to signal level fluctuation or noise. If the PLL goes out of lock, or interference is detected by the co-channel protection circuit before seven cycles are counted, the decoder goes into the long count mode. Each disturbance will reset the counter to zero. The Level Detector will keep the decoder from going into stereo if the IF input level drops 10 dB, but will not change the operation of the pilot counter.

Once the decoder has gone into the stereo mode, it will go instantly back to monaural if either the lock detector on Pin 10 goes low, or if the carrier level drops below the preset threshold. Seven consecutive counts of no pilot will also put the decoder in monaural. In stereo, the co-channel input is disabled, and co-channel or other noise is detected by negative excursions of the I DET, as mentioned earlier. When these excursions reach a level caused by approximately 20% modulation of co-channel, the lock detector puts the system in monaural, even though the PLL may still actually be locked. This higher level of co-channel tolerance provides the hysteresis to prevent chattering in and out of stereo on a marginal signal.

When all inputs to the Pilot Decode block are correct, and it has completed its count, it turns on the Switch, sending the L−R to the Matrix, and switches the pilot lamp pin to a low impedance to ground.

SUMMARY

It should be noted that in CQUAM®, with both channels AM modulated, the noise increase in stereo is a maximum of 3.0 dB, less on program material. Therefore, this is not the major concern in the choice of monaural to stereo switching point as it was in FM, and blend is not needed.

PIN DESCRIPTIONS

Pin 1, 2	— Detector Filters, R_{out} = 4.3 k, recommen 0.0033 μF to V_{CC} to filter 450 kHz compc nents and 100 resistors for loop stability.
Pin 3	— IF Signal Input
Pin 4	— Level Detector filter pin, R_{out} = 8.2 k, 10 μ to ground sets the AGC time constant. Hig impedance output, needs buffer.
Pin 5	— Error Amp compensation to stabilize the Vε Gain feedback loop
Pin 6	— V_{CC}, 6–12 Vdc, suitable for low V_{batt} au tomotive operation, but must be protecte from "high line" condition.
Pin 7, 8	— Left and Right Outputs, NPN emitter followe
Pin 9	— Forced Monaural, MOS or TTL controllab
Pin 10	— Lock detector filter, R_{out} = 27 k, recon mend 2.2 μF to ground.
Pin 11	— AGC'd Q output, NPN emitter follower wi 400 Ω from emitter to Pin 11
Pin 12	— Co-channel Input, 2.0 k series in and 47 feedback
Pin 13	— Pilot Filter Input to op amp, see Figure 8
Pin 14	— Pilot Decode Input (op amp output) emitt follower, R_{out} = 100 Ω
Pin 15	— Stereo Lamp, open-collector of an NP common emitter stage, can sink 50 mA, V_s = 0.3 V at 5.0 mA
Pin 16	— Ground
Pin 17	— Oscillator input, R_{in} = 10 k, do not dc co nect to Pin 18 or ground
Pin 18	— Oscillator feedback, NPN emitter, R_{out} 100 Ω
Pin 19	— Phase Detector Output, current source filter
Pin 20	— Detector Filter, R_{out} = 4.3 k, recomme 0.0033 μF to V_{CC} to filter 450 kHz

FIGURE 5 — CERAMIC VCO

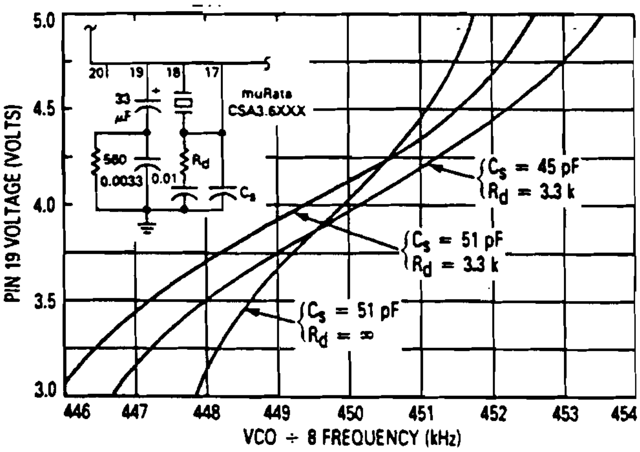


FIGURE 6 — L–C VCO

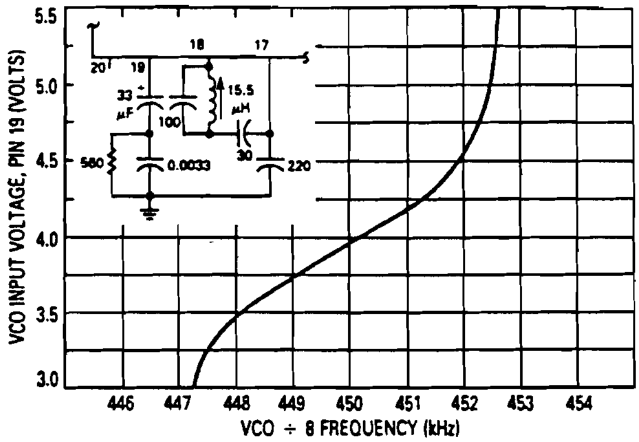


FIGURE 7 — CRYSTAL VCO

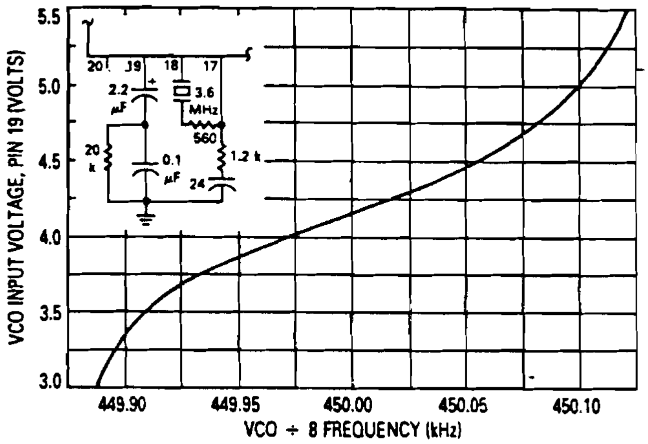
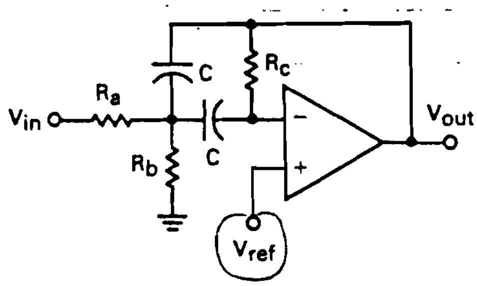


FIGURE 8 — ACTIVE BAND-PASS FILTER



$$R_c = \frac{Q}{\pi f_o C}$$

$$R_a = \frac{R_c}{2 A_o}$$

$$R_b = \frac{R_a R_c}{4Q^2 R_a - R_c}$$

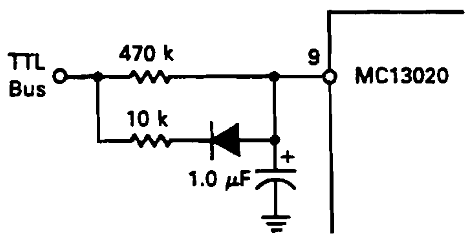
where:
- f_o = center frequency
- A_o = gain at f_o
- $Q \leq 10$

Choose values for F_o, A_o, Q, and convenient C, solve for resistors

In this application:

f_o = 25 Hz	R_a = 5.6 k
A_o = 20	R_b = 910
Q = 8.2	R_c = 220 k
C = 0.47 μF	

FIGURE 9 — FORCED MONAURAL OPTIONAL DELAY CIRCUIT



A.2. Measurements of the decoder

The measurements are performed on the MC13020P decoder in the radio. The input at pen 3 is connected to the output of the C-QUAM signal generator through a capacitor of 10nF. The carrier frequency is 468 KHz.
The performed measurements are:
- Harmonic distortion vs. modulation index at 400 Hz (figure A.1) and 1000 Hz (figure A.2).
- Channel separation vs. modulation frequency at m=0.4 (figure A.3).
- Performance (S+N, N ,channel separation, THD) vs. input signal level when in the radio (figure A.4). Here the performance of the total radio is measured.

For the first two measurements the input signal level is 117 dB EMF (=0.7 V EMF).

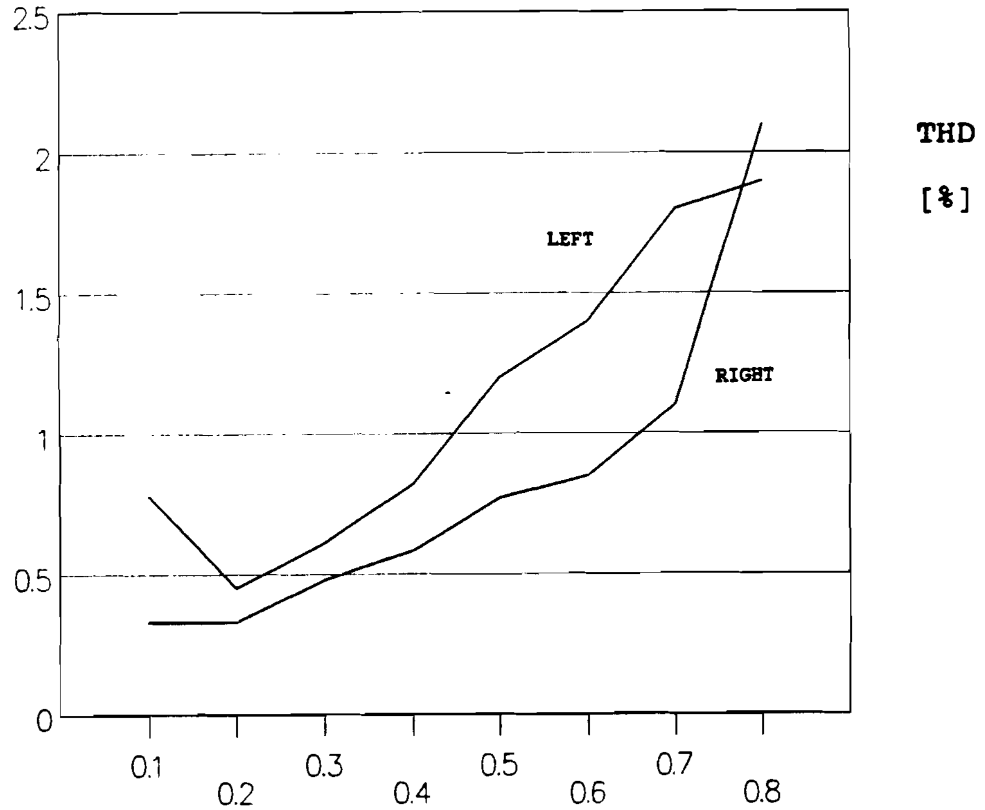


Figure A.1. Harmonic distortion vs. modulation index m at 400 Hz.

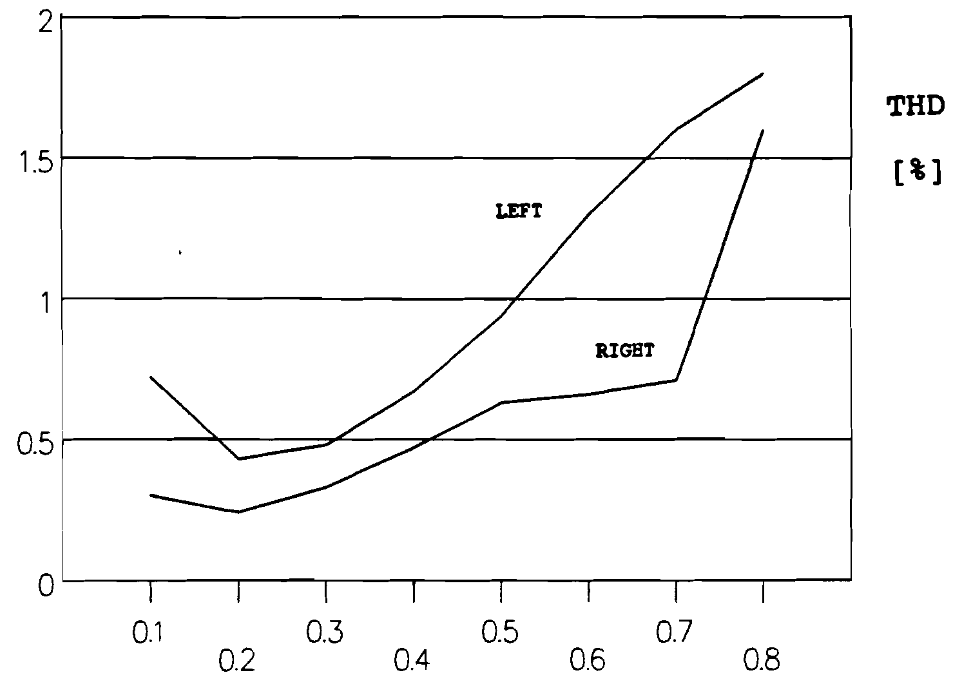


Figure A.2. Harmonic distortion vs. modulation index at 1000 Hz.

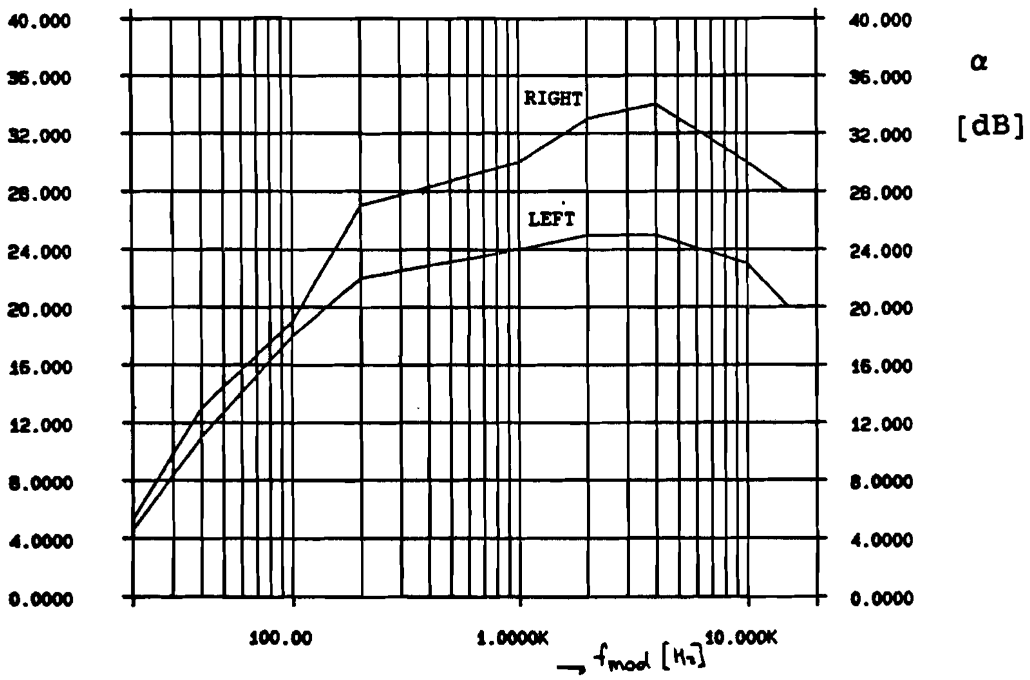


Figure A.3. Channel separation α vs. modulation frequency at m=0.4.

AM performance Car Radio   90DC788 LEFT (Stereo)   05/12/1990

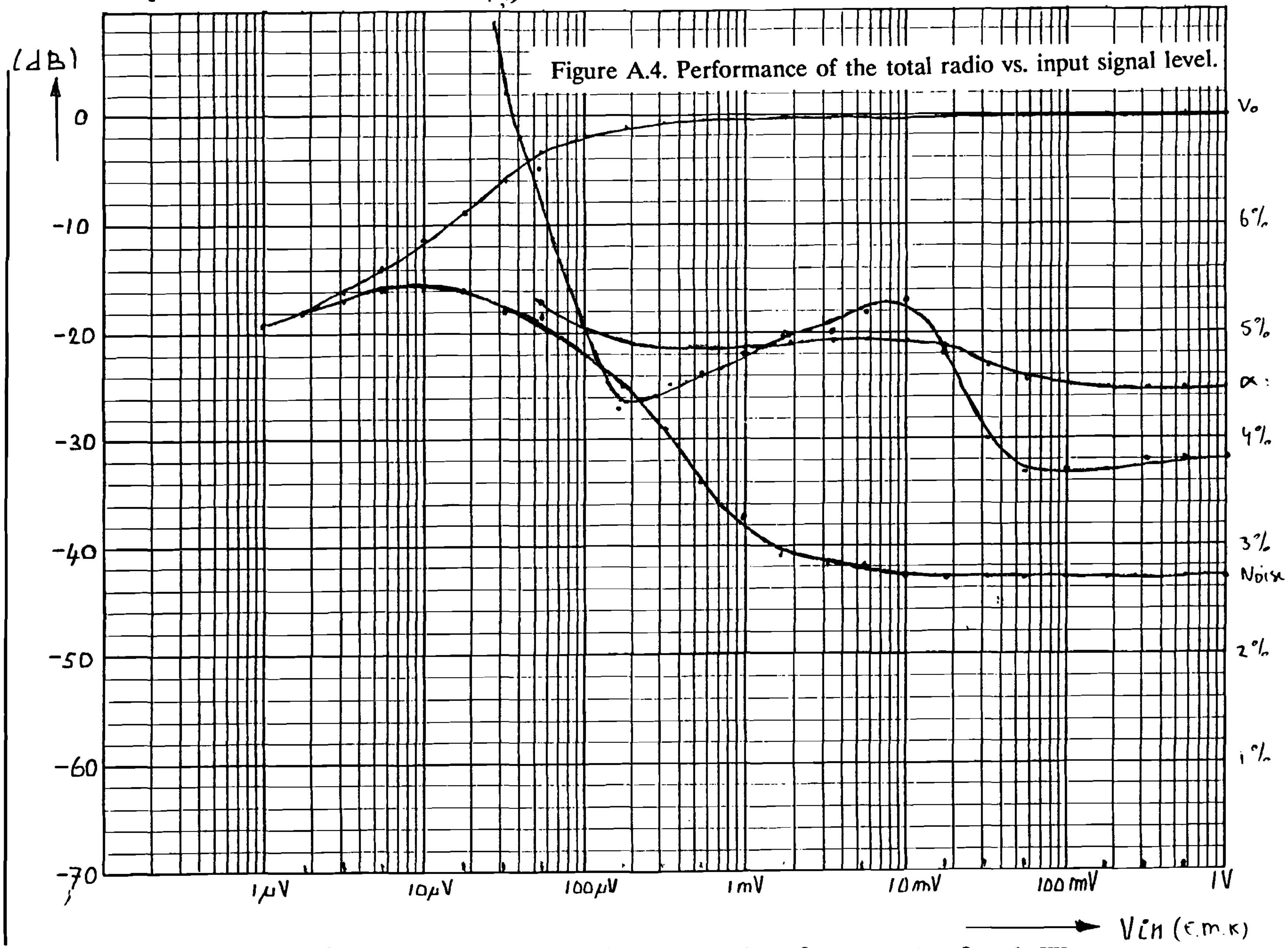


Figure A.4. Performance of the total radio vs. input signal level.

V_0: L=R, m=0.3, f_{MOD}=1 KHz, 1 V over 4Ω   THD: L-only, m=0.8, f_{MOD}=1 KHz
Noise: incl. dB(A)-filter   α: L-only, m=0.3, f_{MOD}=1 KHz
pilot=10%

APPENDIX B: The pilot detection circuit

As described in paragraph 4.6 the pilot detection circuit contains the blocks as in figure B.1.

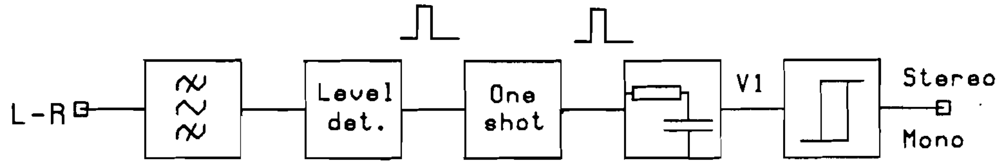


Figure B.1. Configuration of the pilot detection circuit.

In order to extract the pilot tone from the L-R information a band-pass filter is used. This can be a simple one stage filter, using an operational amplifier. With the filter in figure B.2 it is possible to obtain a maximum Q of about 10.

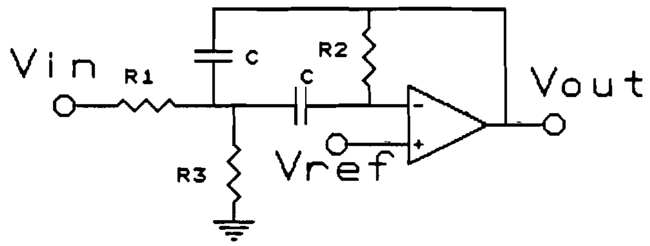


Figure B.2. A band-pass filter.

The transfer function of the filter is given by:

$$\frac{V_{OUT}}{V_{IN}} = \frac{-j\omega R_2 R_3 C}{R_1 + R_3 + j\omega R_1 R_3 C + R_1 R_2 R_3 C^2 (j\omega)^2} = \frac{A_0\, j\omega/\omega_0 Q}{1 + j\omega/\omega_0 Q + (j\omega/\omega_0)^2} \qquad \text{(B.1)}$$

where: A_0 = amplification at the resonant frequency ω_0
Q = quality of the band-pass filter

If we establish the values of A_0, ω_0 ($=2\pi 25$ Hz) and Q and choose a value for C, we can calculate the remaining components (see the data sheets on the Motorola decoder in appendix A.1).
The filter is followed by a comparator, which compares the output of the filter with a reference (V_{REF1} in figure B.5). The output of the comparator will be 0 when the output of the filter is lower than the reference voltage V_{REF1}. When the output of the filter is greater than V_{REF1}, the output of the level detector will be close to the plus. V_{REF1} must be chosen in such a way that at modulation levels lower than 4% for the pilot tone the output of the comparator will remain 0. This means that if the level of the pilot tone exceeds 4% and varies slightly, the comparator will produce a square wave of 25 Hz with a varying duty cycle. This square wave has to be converted into a square wave with a constant duty cycle. With such a square wave we can charge a capacitor. If the charge period is not constant it is not

possible to take the voltage over the capacitor as a measure for deciding between mono and stereo. A circuit which performs such a function is the monostable multivibrator (an example is shown in figure B.3).

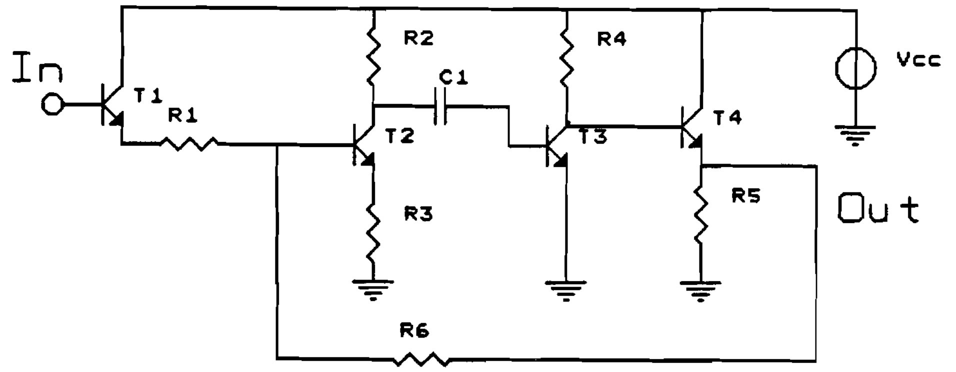


Figure B.3. A monostable multivibrator.

When T_1 is not conducting, the base of T_2 will be low. This means that the collector of T_2 is high. T_3 will be conductive and T_4 will closed.
When transistor T_1 is made conductive during a short period, transistor T_2 will immediately switch on and the voltage at the collector of T_2 will drop, making the base of T_3 negative. This means that T_4 will conduct, supplying T_2 with a positive bias. Then the capacitor will discharge itself over R_4, until T_3 starts to conduct again. In the mean time T_1 will be closed again, since it was only a short pulse that triggered T_1. As soon as T_3 starts to conduct, the emitter of T_4 will be low again and T_2 will be closed as well. The emitter voltage of T_4 has now remained high from the start of the incoming pulse until T_3 starts to conduct again. This time t_0 is only dependent on a timing constant of the capacitor in combination with a resistor. When we choose this time t_0 to be about half the period time of the pilot, t_0 will be independent of the modulation level of the pilot.
With the now gotten square wave we can charge a capacitor. The voltage will then rise according to an exponential function. During the time the capacitor is not charged, it will be discharged through the resistor parallel to the capacitor. The discharge time is larger than the charge time.
When the pilot is present during a certain number of cycles and has the correct level, the capacitor will be charged up to a value where it will pass a threshold. The decoder is then switched to stereo.
The voltage across the capacitor will rise to a maximum where the charging is balanced by the discharging. When the pilot disappears several times, the capacitor will be less charged than discharged. Thus the voltage across the capacitor will drop. When it is below a certain predetermined level, the following comparator will switch the decoder to mono.
When the decoder is in mono mode and a pilot tone is present during a certain number of cycles of 25 Hz, the capacitor will be charged sufficiently to switch the decoder to stereo. By short circuiting this capacitor it is possible to keep the decoder in the monaural mode.

The circuit with the two thresholds, which determines to switch to mono or stereo can be made with an opamp, used as a comparator.
A comparator with a lower and upper threshold is given in figure B.4.

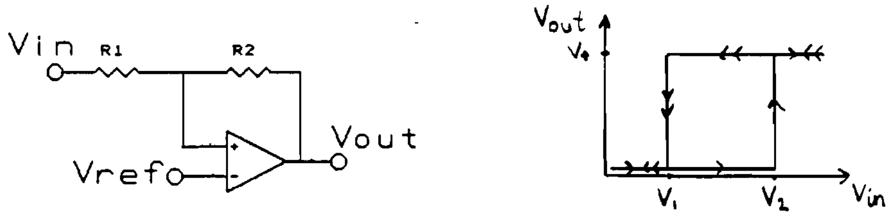


Figure B.4A. A comparator with two thresholds. B.4B. Input-output relation.

For an input voltage lower than V_1 the output will remain low. For an input voltage larger than V_2, the output will be high. In between the two voltages, the output will not change. In figure B.4B the output is sketched as function of the input. V_1 and V_2 are given by the following equations:

$$V_1 = V_{REF} - K\,(V_+ - V_{REF}) = (1+K)\,V_{REF} - K\,V_+ \tag{B.3}$$

$$V_2 = V_{REF} + K\,V_{REF} = (1+K)\,V_{REF} \tag{B.4}$$

where: $K = R_1/R_2$
V_+ = the maximum output voltage of the opamp.

The complete pilot detection circuit can be seen in figure B.5.

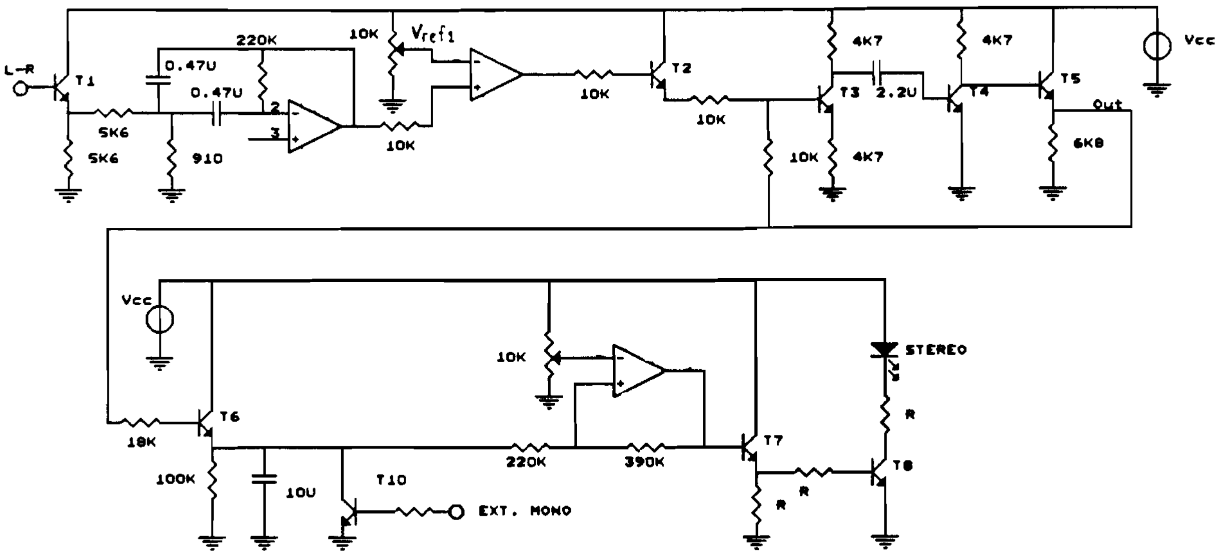
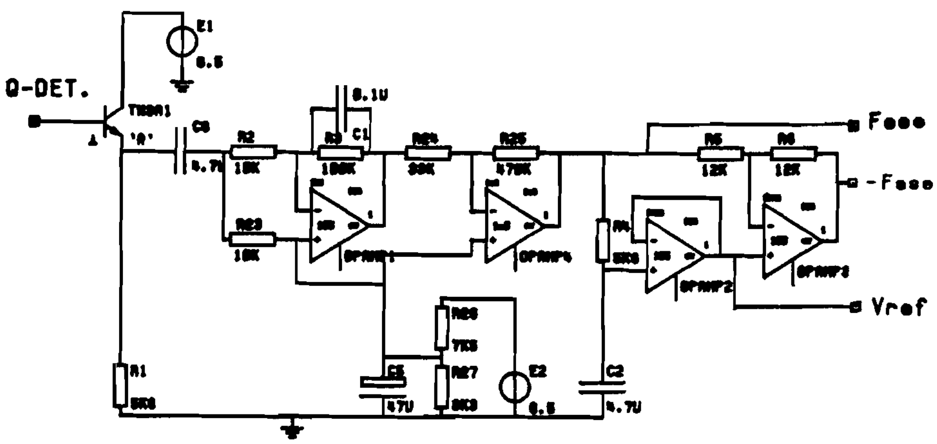
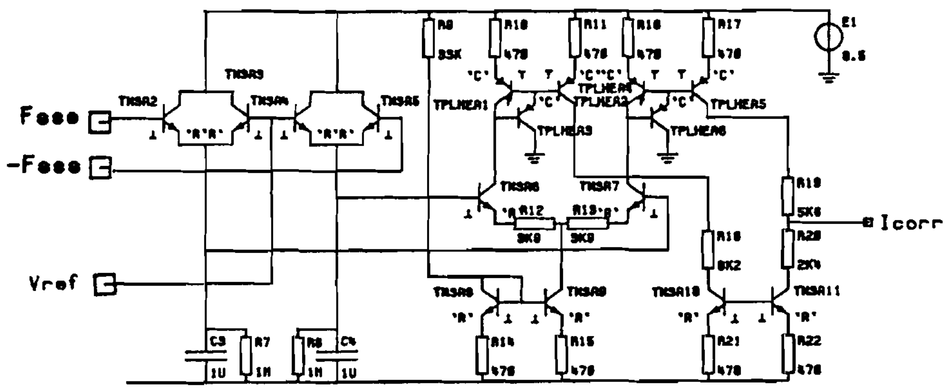


Figure B.5. The pilot detection circuit.

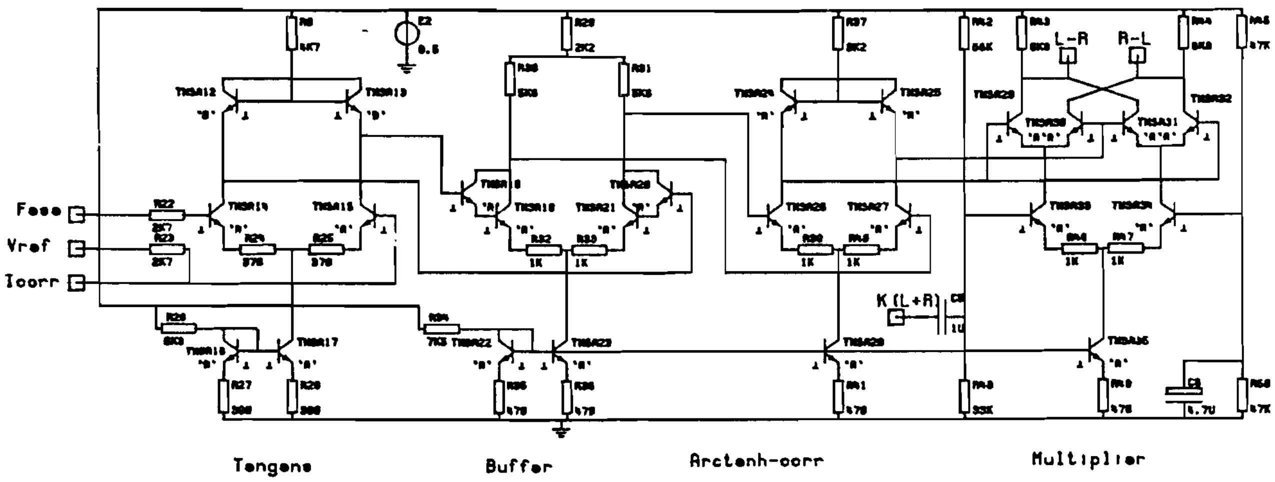
APPENDIX C: Diagrams of the alternative decoder.

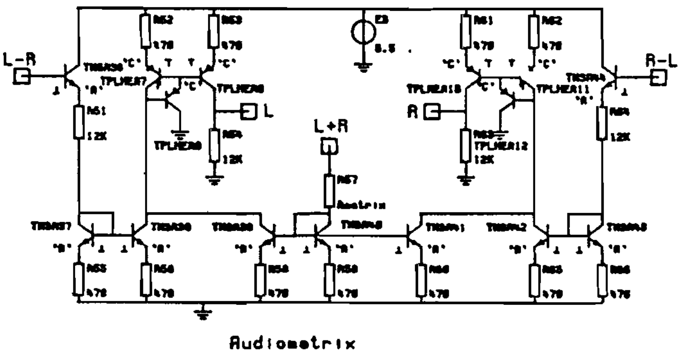


Integrator   Inverter



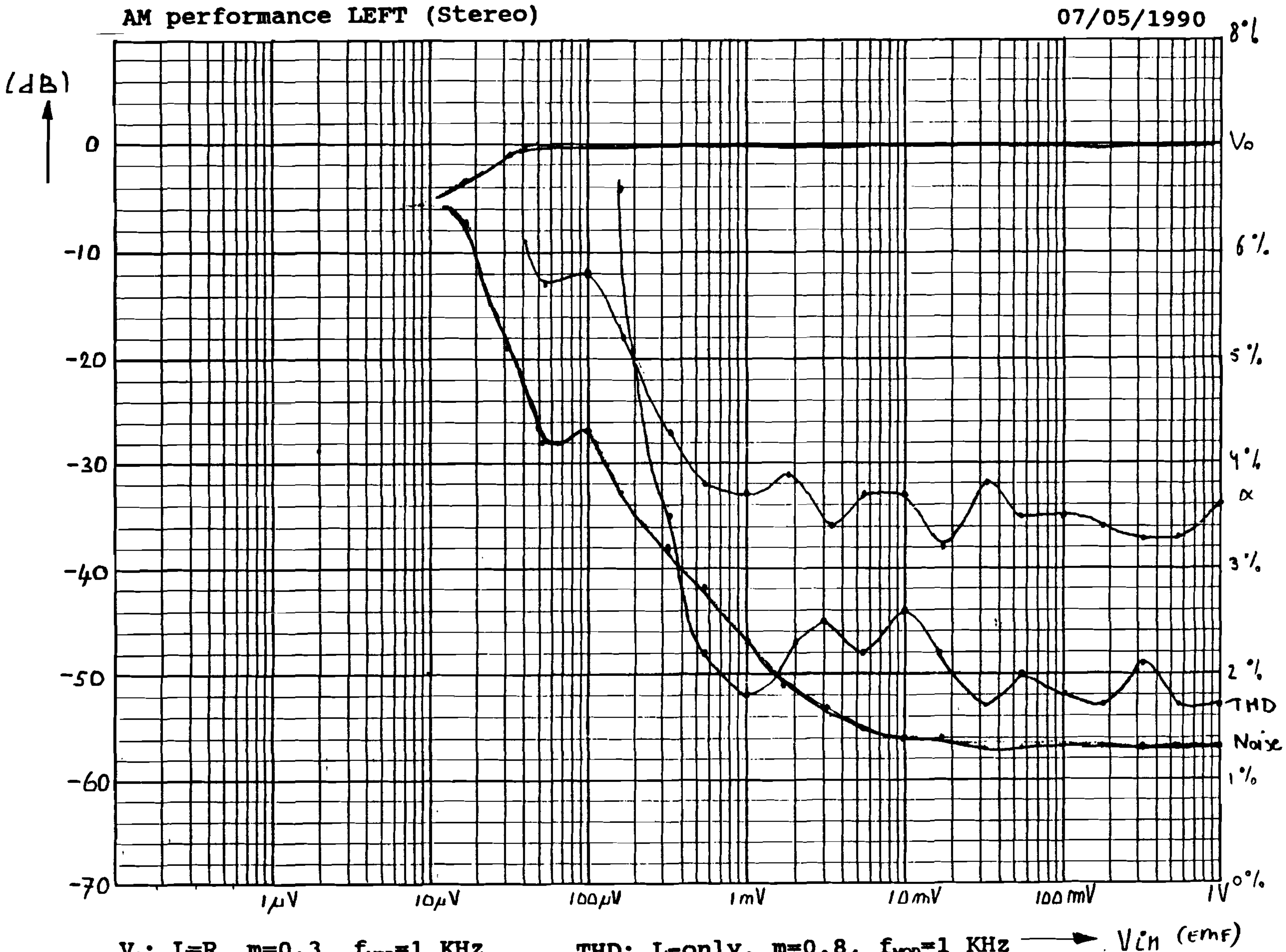
Leveldetector   Differential Amplifier


Fase
Vref
Icorr
L-R
R-L
K(L+R)
Tangens
Buffer
Arctanh-corr
Multiplier


L-R
R-L
L
R
L+R
Audiomatrix

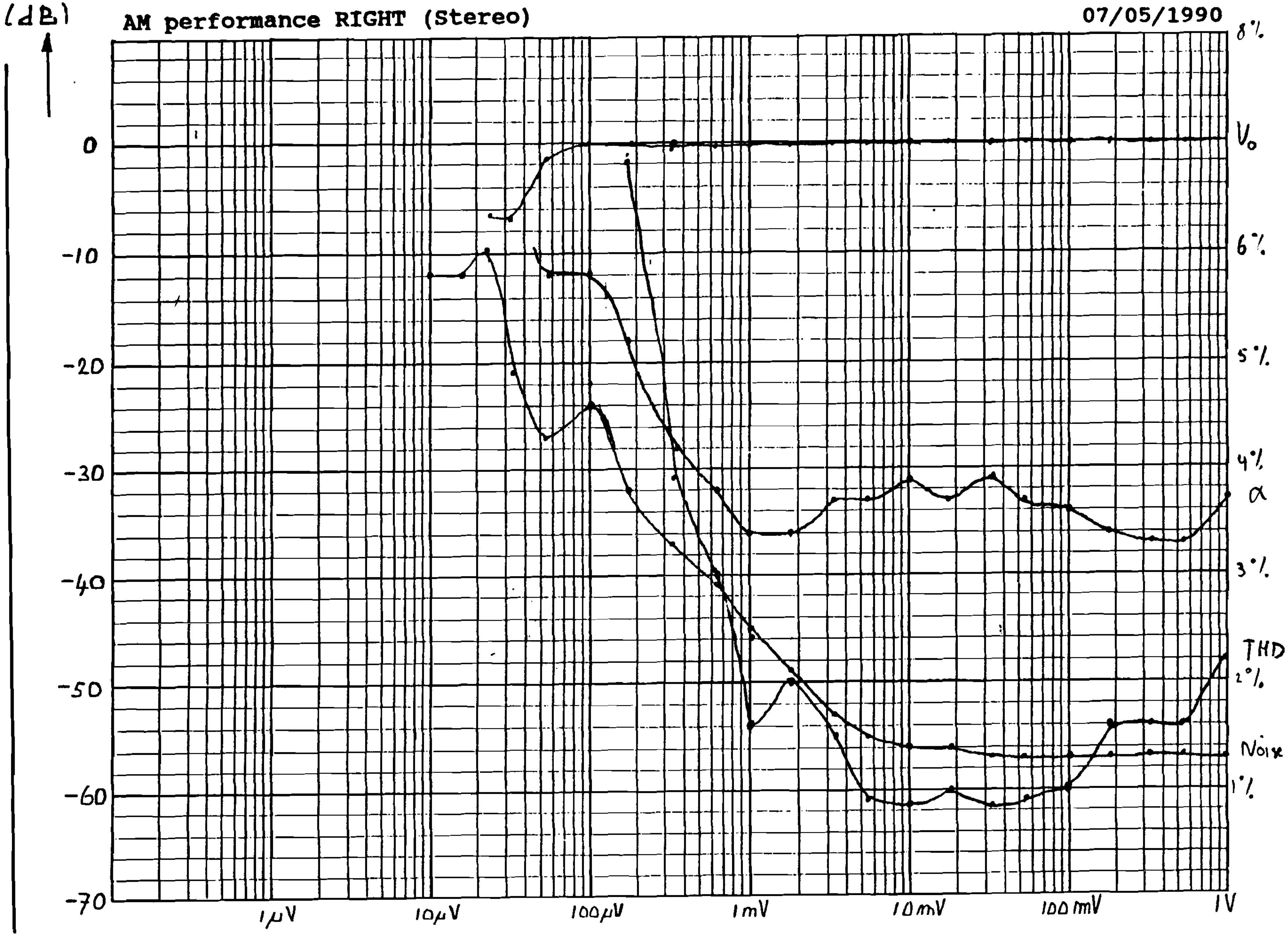
APPENDIX D: Measurements of the alternative decoder.

D.1. The breadboard model without IF filtering



V_0: L=R, m=0.3, f_{MOD}=1 KHz
Noise: incl. dB(A)-filter
THD: L-only, m=0.8, f_{MOD}=1 KHz
α: L-only, m=0.3, f_{MOD}=1 KHz

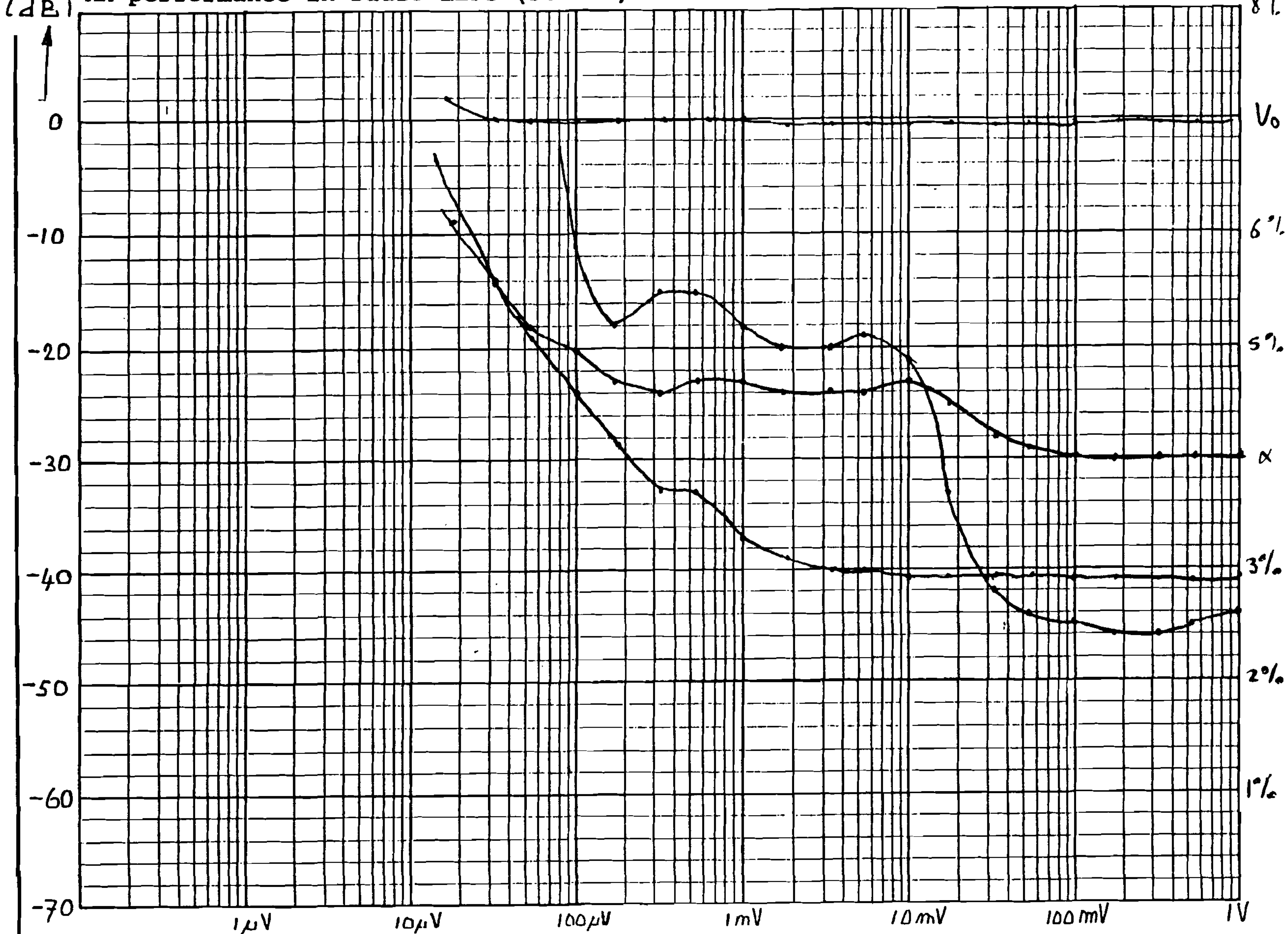
AM performance RIGHT (Stereo) 07/05/1990



V_0: L=R, m=0.3, f_{MOD}=1 KHz
Noise: incl. dB(A)-filter

THD: R-only, m=0.8, f_{MOD}=1 KHz
α: R-only, m=0.3, f_{MOD}=1 KHz

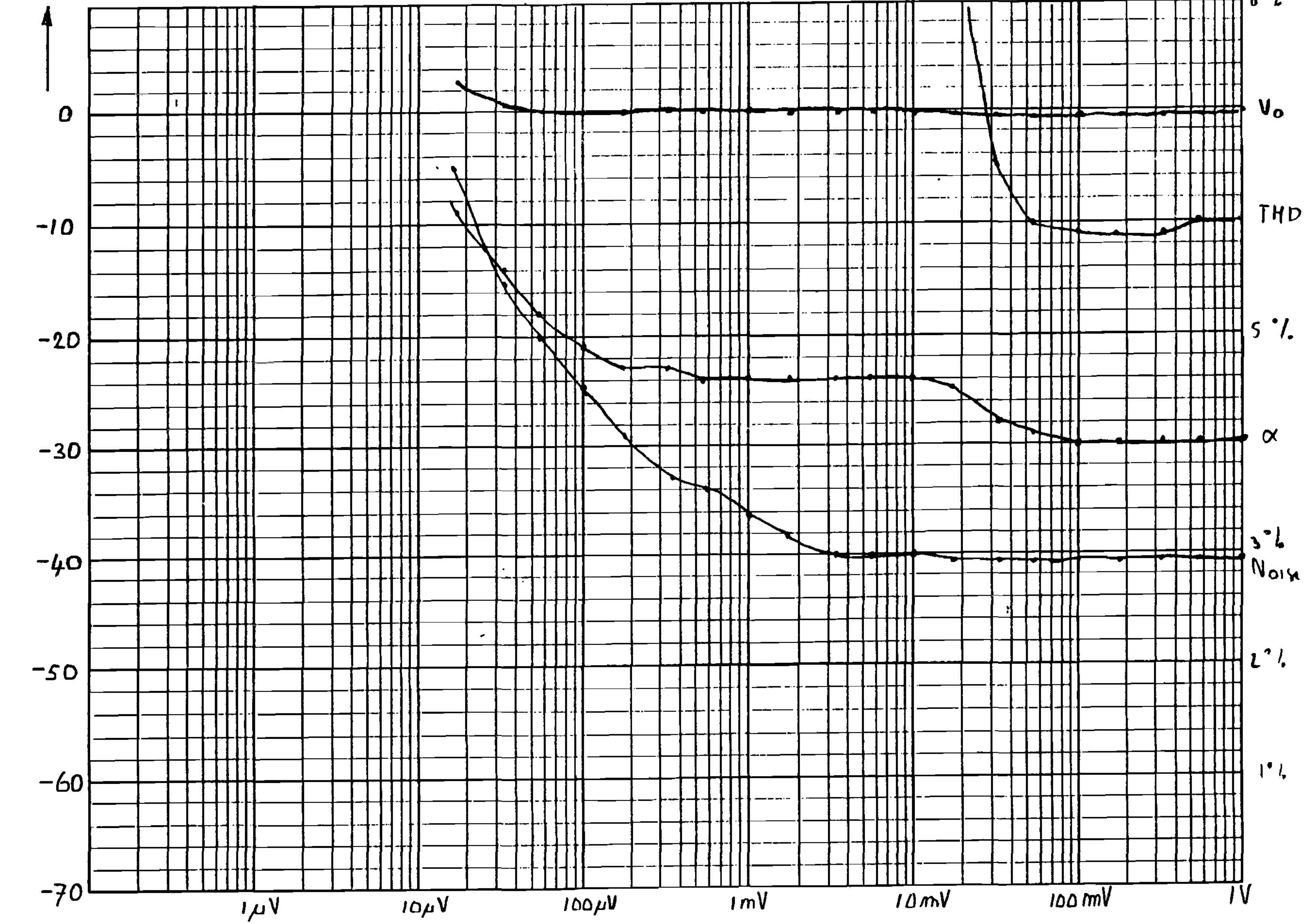
AM performance in radio LEFT (Stereo) 25/06/1990



V_0: L=R, m=0.3, f_{MOD}=1 KHz
Noise: incl. dB(A)-filter

THD: L-only, m=0.8, f_{MOD}=1 KHz
α: L-only, m=0.3, f_{MOD}=1 KHz

AM performance in radio RIGHT (Stereo) 25/06/1990



V$_0$: L=R, m=0.3, f$_{MOD}$=1 KHz
Noise: incl. dB(A)-filter

THD: R-only, m=0.8, f$_{MOD}$=1 KHz ⟶ Vin (emf)
α: R-only, m=0.3, f$_{MOD}$=1 KHz

APPENDIX E: Simulations of system errors

As described in paragraph 5.1.1 a few errors can best be simulated in order to get an idea of their importance to the system performance.
In paragraph E.1 the non-linear errors are analysed and in paragraph E.2 the phasing error between $\phi(t)$ and L+R is analysed.
In the simulations the left channel is modulated with a sine wave and the right channel is not modulated.

E.1. Errors in amplitude

Here below a listing is given of the used program. The following errors have been simulated:

- the amplitude of the phase is not correct ($K_P \neq 1$, figures E.1 and E.2)
- the amplitude of 1 compared to L+R is not correct ($K_M \neq 1$, figures E.3 and E.4).

The modulation index for the left channel is 0.8. This means that the resulting amplitude for the left channel should be 1.6.
The results show that both errors have approximately the same influence but an error in K_P has more influence on the performance. The maximum cross talk can be as high as -15 dB (0.3 compared to 1.6 for K_P = 1.05), which cannot be overlooked. The result is that both errors should be much smaller than 5%.

```
Program listing:

circuit$
  Kp=.95 $
  Km=1 $
C:
  Left channel
$
  Ha=0.8*sin(2*pi*1000*t) $
  E1 (1,0) $
  R1 (1,0) 22k $
  C1 (1,0) 1p $
  V1 (1,0) $
C:
  Right channel
$
  Hb=0 $
C:
  Functions
$
  Sum=Ha+Hb $
  Dif=Ha-Hb $

  phase=arctan(Dif/(1+Sum)) $
  Tanphase=tan(Kp*phase) $

  Lacc=Sum+(Km+Sum)*Tanphase $
  Racc=Sum-(Km+Sum)*Tanphase $
end $
TR $
  T=AN(0,2ML,1000) $
  file:Lacc,Racc;(format=ifp) $
end$
run $
```

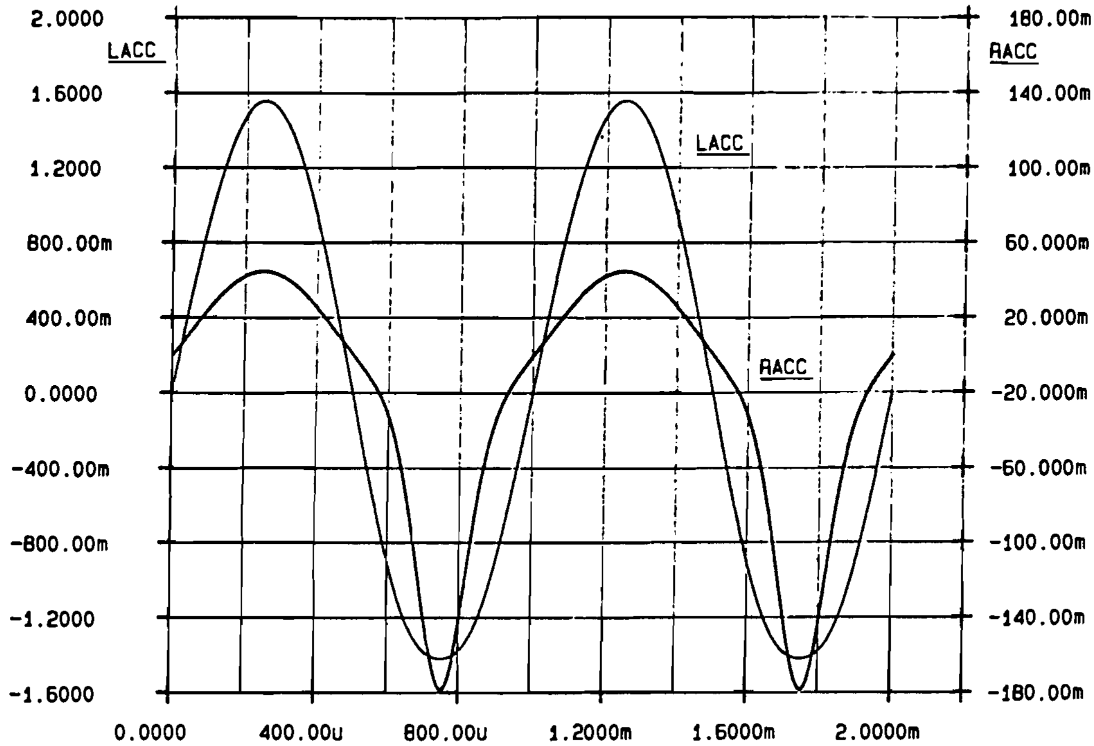


Figure E.1. Left and right channel after decoding with error K_r = 0.95.

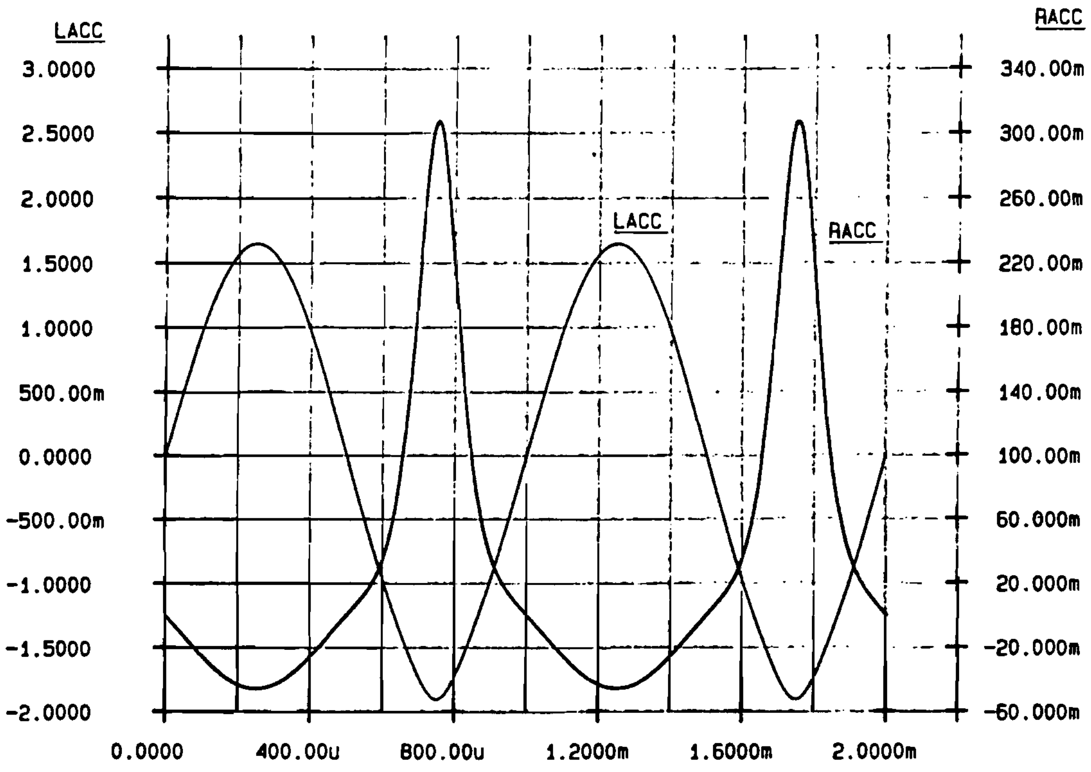


Figure E.2. Left and right channel after decoding with error K_r = 1.05.

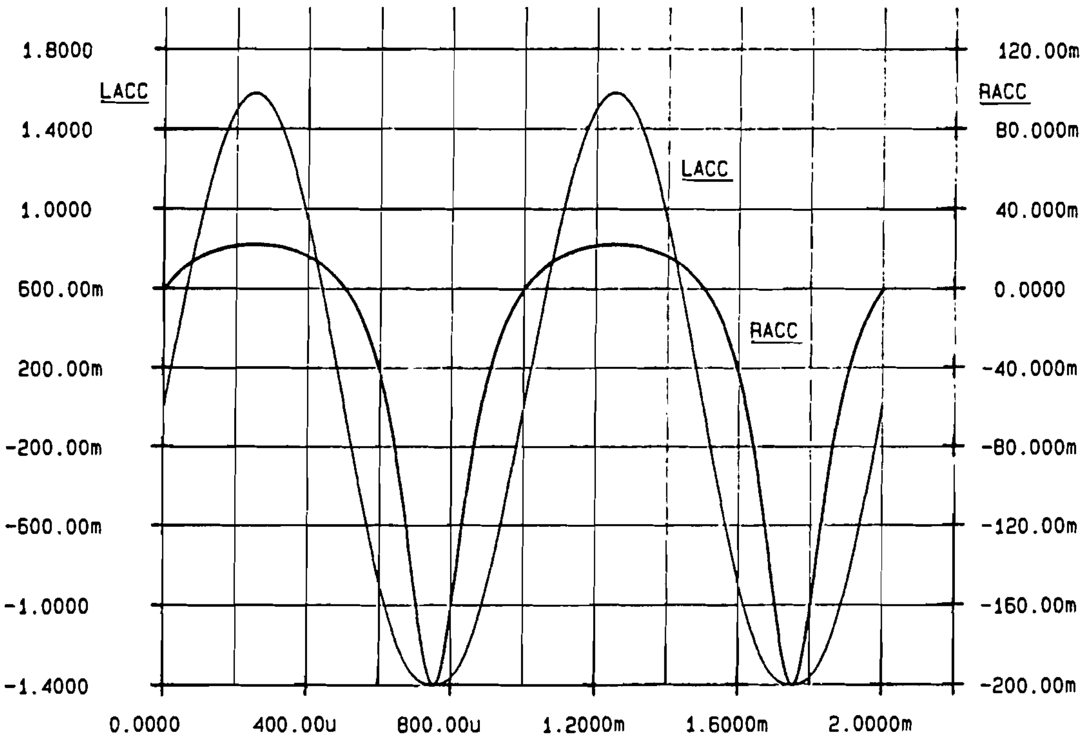


Figure E.3. Left and right channel after decoding with error K_M = 0.95.

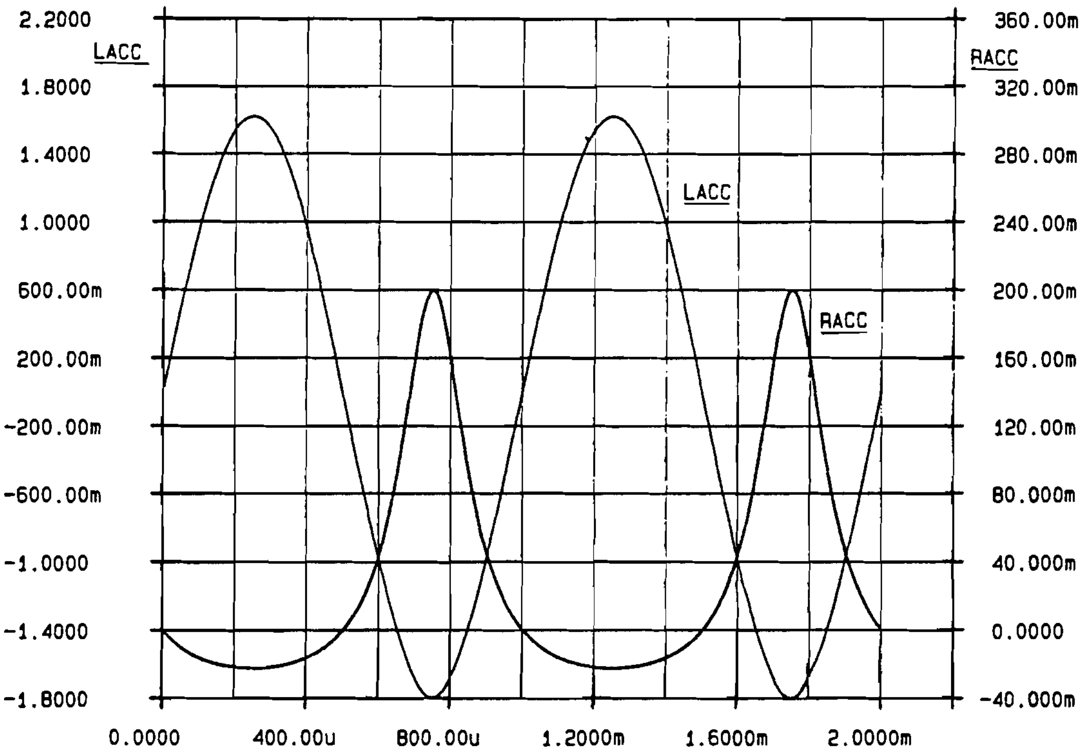


Figure E.4. Left and right channel after decoding with error K_M = 1.05.

E.2. An error in phase

Here a phase shift between $\phi(t)$ and L+R is simulated. A listing of the program is shown at the bottom of this page. From figures E.5 and E.6 (see following pages) it can be concluded that the cross talk is proportional to the phase error. An error of only 1° gives a cross talk figure of -40 dB (about 7.6×10^{-3} compared to 0.8). An error of 5 ° gives a cross talk figure of -27 dB (36×10^{-3} compared to 0.8). The phase error should therefore be 1° or smaller. Otherwise the remaining signal on the channel that is not modulated will be too strong.

```
Program listing:

circuit $
C:
  fi = phase error in degrees
$
  fi=1 $

  L=0.4*sin(2*pi*1000*t) $
  Lacc=0.4*sin(2*pi*(1000*t+fi/360)) $

  el (1,0) L $$
  R1 (1,0) 22k $
  C1 (1,0) 1p $

  Tanphase=Lacc/(1+Lacc) $

  Left=L+Tanphase*(1+L) $
  Right=L-Tanphas*(1+L) $
end $
TR $
  T=AN(0,2ML,1000) $
  file:left,right $
end $
run $
```

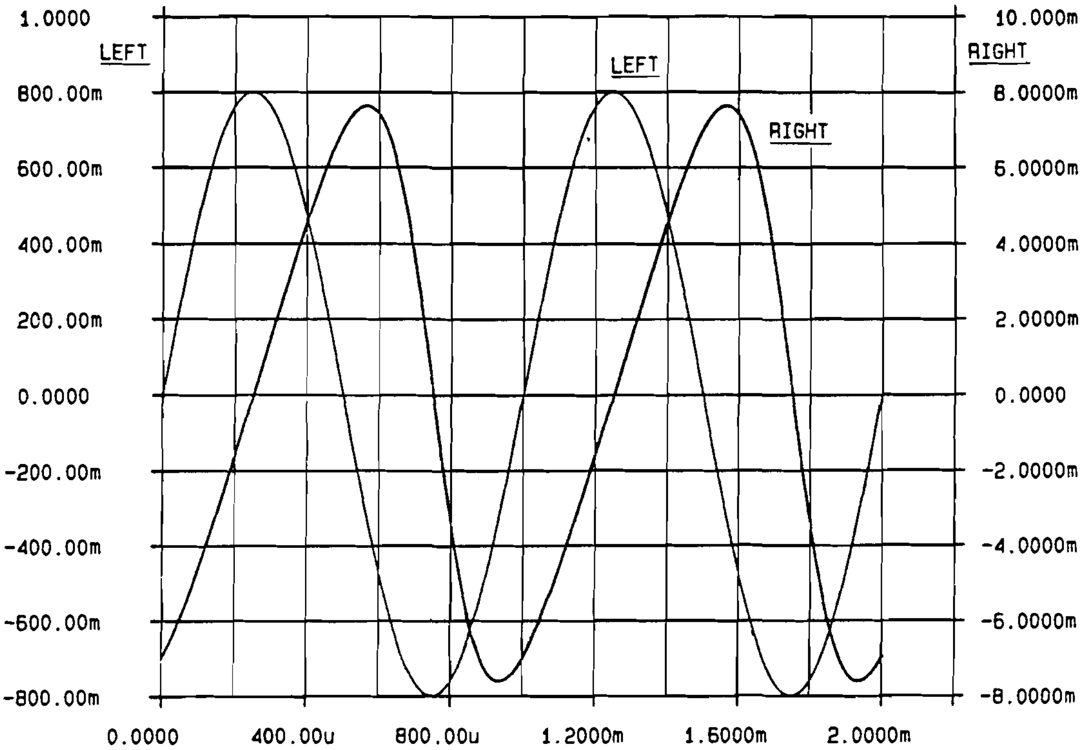


Figure E.5. Left and right channel after decoding with a phase error of 1°.

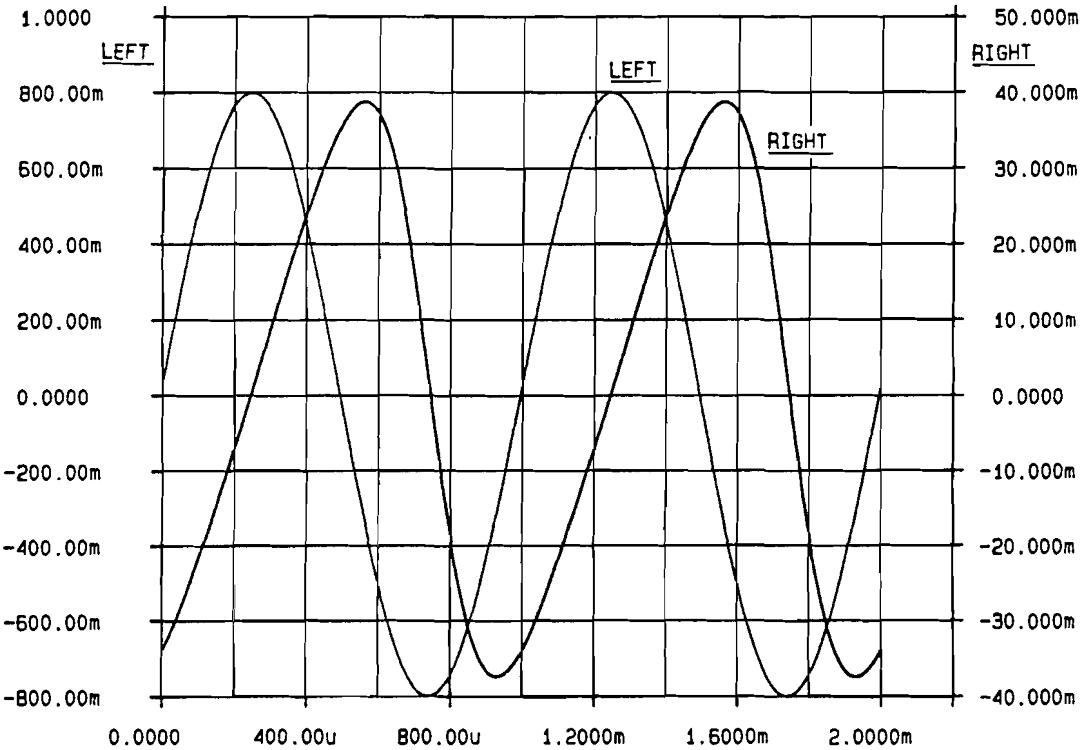


Figure E.6. Left and right channel after decoding with a phase error of 5°.