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Title: Integrated Services Digital Network basic rate access based on echo cancellation.

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CONTENTS

0. ABSTRACT

In the following decennium, a new telecommunication network called Integrated Services Digital Network will be introduced. The basic analog subscriber connection will then be replaced by a digital link called basic rate access. The basic rate access is to be implemented using the existing 2-wire subscriber loop for full-duplex digital transmission. The realization is based on echo cancellation and hybrid transformers.

Three line codes proposed for the basic rate access have been considered, namely, 2B1Q, 4B3T and Alternate Mark Inversion. These line codes have been evaluated in this work with respect to range and compatibility with existing subscriber services.

A near-end-crosstalk simulator and a 2B1Q simulator have been realized for measuring of possible disturbance from the basic rate access to existing subscriber services. The measurements showed that the basic rate access will interfere with a carrier frequency equipment (known as the 1+1 system) used for frequency multiplexing of two subscribers on one line. The line code 2B1Q was found to be the best performing line code with respect to both criterions.

Two echo cancellation configurations have also been evaluated. The two configurations both generate a digital echo replica which is subtracted from the incoming signal containing an unwanted echo component. The difference between the two configurations is whether to do the subtraction digital or analog. The digital configuration was found to be best suited.

In addition, this paper comprises a system study of the Network Termination 1 with emphasis on technical solutions and production considerations. The Network Termination 1 is basically a 4-to-2 wire converter (and vice versa) with a S-interface (4 wire) at one side and an U-interface (2 wire) at the other.

No CCITT recommendation exists for the U-interface, which has resulted in two different solutions, one specified by American National Standard Institution and the other by Deutsche Bundespost. A Network Termination 1 configuration with all U-interface circuitry on a daughter-board is proposed. This enables a change of U-interface without having to replace the whole Network Termination 1.

TABLE OF SYMBOLS

α	-	Iteration constant (p.46)
α_c	-	Cable-attenuation per km (p.28)
σ^2	-	Variance (noise power) (p.83)
σ_e^2	-	Quantization noise power (p.53)
σ_r^2	-	Mean square of the residual signal (p.45)
τ_{eq}	-	Integral of the squared pulse (p.84)
a(k)	-	Transmitted data from exchange (p.44)
$\mathbf{a}_k$	-	Exchange input vector (p.44)
a_n	-	Near-end-crosstalk attenuation (p.28)
A	-	Distance between amplitude levels AMI (p.22)
A_i	-	MMS43 alphabet (p.24)
b(k)	-	Transmitted data from subscriber (p.44)
b_n	-	Transmitted data symbol (p.24)
$\mathbf{b}_k$	-	Subscriber input vector (p.44)
B	-	Distance between amplitude levels 4B3T (p.25)
B_T	-	Transmission bandwidth (p.22)
$c_i(k)$	-	Filter coefficients (p.44)
$\mathbf{c}_k$	-	Filter coefficient vector (p.44)
cs(k)	-	Synchronization word correlation function (p.48)
C	-	Information capacity (p.21)
C(f)	-	Transfer function of the cable (p.28)
D	-	Distance between amplitude levels 2B1Q (p.26)
D(z)	-	Decimation filter transfer function (p.55)
D(f)	-	Amplitude transfer function of the decimation filter (p.55)
e(k)	-	Echo signal - time discrete (p.19)
e(t)	-	Echo signal - time continuous (p.37)

$\hat{e}(k)$	-	Residual echo signal - time discrete (p.19)
$\hat{e}(t)$	-	Residual echo signal - time continuous (p.37)
$E(z)$	-	z-transform of the delta-sigma-modulator quantization noise (p.53)
$\|E(f)\|^2$	-	Spectral density of the quantization noise (p.52)
f_b	-	Baseband (p.53)
f_l	-	Lower frequency limit (p.95)
f_s	-	Sampling frequency (p.51)
f_u	-	Upper frequency limit (p.95)
$f(k)$	-	Total received signal including the echo signal - time discrete, = s(k) + n(k) + e(k) (p.38)
$f(t)$	-	Total received signal - time continuous (p.37)
f_δ	-	Frequency deviation (p.32)
$g(i)$	-	Echo path coefficient (p.44)
g	-	Echo path vector (p.44)
$G(f)$	-	Power spectrum (p.22)
$h(k)$	-	Transmission path coefficient (p.44)
h	-	Transmission path vector (p.44)
H	-	Noise density function (p.84)
$H(s)$	-	Transfer function (p.51)
$H_e(f)$	-	Equalizer transfer function (p.81)
k	-	Number of discrete time interval t=kT (p.19)
k_s	-	Number of states in a pseudo-random sequence (p.67)
L	-	Transmission range limited by near-end-crosstalk (p.29)
M	-	Number of coefficients (p.44)
n	-	Efficiency (p.21)
$n(k)$	-	Additive noise - time discrete (p.19)

$n(t)$	-	Additive noise - time continuous (p.37)
N_0	-	Noise power inside the baseband (p.53)
$\|N(f)\|^2$	-	Spectral density of the delta-sigma-modulator output (p.52)
$\|N'(f)\|^2$	-	Spectral density of the decimation filter output (p.56)
NEXT(f)	-	Transfer function of the NEXT-path (p.28)
N_R	-	Noise power at the receiver (p.83)
N	-	Number of coefficients in a digital filter (p.44)
P_i	-	Symbol probability (p.21)
$p_y(y\|H_0)$	-	Conditional probability density function (p.83)
ph(k)	-	Synchronization word difference (p.48)
P_{be}	-	Bit error probability (p.28)
P_{ei}	-	Conditional error probability (p.83)
P_{se}	-	Symbol-error probability (p.83)
P(f)	-	Fourier transform of the transmitted pulse (p.22)
Q( )	-	Area under the gaussian tail (p.83)
r(k)	-	Residual received signal, (received signal plus noise) - time discrete (p.19)
r(t)	-	Residual received signal - time continuous (p.37)
r	-	Transmission rate (p.21)
R(f)	-	Transfer function of the receiver filter (p.28)
$R_a(n)$	-	Amplitude auto-correlation (p.22)
RDS(k)	-	Running digital sum (p.24)
s(k)	-	Received signal - time discrete (p.19)
s(t)	-	Received signal - time continuous (p.37)
sw(i)	-	Synchronization word (p.48)

S	-	Power (p.95)
S_u	-	Mean square value of received signal plus additive noise (p.45)
S(f)	-	Transfer function of the transmitter filter (p.28)
SNR_{req}	-	Minimum required signal-to-noise ratio (p.28)
S_R	-	Signal power at the receiver (p.84)
T	-	Sampling period (p.52)
u(k)	-	Received signal plus additive noise, s(k) + n(k) - time discrete (p.45)
V	-	Delta-sigma-modulator feedback level (p.51)
W	-	Basebandwidth (p.32)
X(z)	-	z-transform of the delta-sigma-modulator input (p.53)
Y(z)	-	z-transform of the delta-sigma-modulator output (p.53)
z	-	z-transform operator (p.53)

MATHEMATICAL NOTATION

$E\{\bullet\}$	-	Mathematical expectation (p.44)
$\delta/\delta c_i(k)$	-	Partial derivative with respect to a filter coefficients (p.46)
grad	-	Gradient (p.46)
(+)	-	Modulo two addition (p.49)
$sign(\bullet)$	-	Sign operator (p.47)
$(\bullet)^T$	-	Transposed (p.44)

TABLE OF ABBREVIATIONS

ADC	-	Analog-to-digital converter (p.36)
ADF	-	Adaptive digital filter (p.16)
AMI	-	Alternate Mark Inversion (p.20)
ANSI	-	American National Standard Institute (p.20)
CCITT	-	The international telegraph and telephone consultative committee (p.14)
cfp	-	Cable failure percentage (p.75)
DAC	-	Digital-to-analog converter (p.37)
DBP	-	Deutsche Bundespost (p.63)
DFE	-	Decision feedback equalizer (p.26)
DPLL	-	Digital phase-lock loop (p.48)
DSM	-	Delta-sigma-modulator (p.50)
DUT	-	Delft University of Technology (p.53)
EC	-	Echo canceller (p.36)
ET	-	Exchange termination (p.17)
HFSL	-	High frequency subscriber loop (p.32)
IC	-	Integrated circuit (p.33)
IDL	-	Interchip digital link (p.65)
IDN	-	Integrated digital network (p.14)
IOM	-	ISDN oriented modular interface (p.41)
ISDN	-	Integrated services digital network (p.14)
LAN	-	Local area network (p.16)
LT	-	Line termination (p.17)
MMS43	-	Modified monitored sum 4B3T (p.20)
MOS	-	Metal oxide semiconductor (p.50)
NEXT	-	Near-end-crosstalk (p.27)
NLG	-	Dutch guilder (p.11)
NOK	-	Norwegian kroner (p.11)
NTA	-	Norwegian Telecommunication Administration (p.63)
NTRL	-	Norwegian Telecom research lab. (p.30)
NT1/2	-	Network Termination 1/2 (p.16)
PABX	-	Private branch exchange (p.16)
PCM	-	Pulse code modulation (p.31)
PDF	-	Probability density function (p.83)
PDM	-	Pulse-density-modulation (p.40)
PRBS	-	Pseudo random bit sequence (p.67)
SIC	-	S-interface circuit (p.41)
SNR	-	Signal-to-noise ratio (p.19)
TA	-	Terminal adapter (p.17)
TE1/2	-	Terminal equipment 1/2 (p.16)
UIC	-	U-interface circuit (p.41)
USD	-	United States dollar (p.32) (1 USD = 6.60 NOK = NLG 2.10)
XOR	-	Exclusive-or gate (p.49)

1. INTRODUCTION

The Integrated Services Digital Network (ISDN) is about to be introduced in Norway. The introduction will start with a pilot project, followed by pilot services. ISDN requires the establishment of a digital link to the subscriber, which in its simplest form is called basic rate access. A solution using only 2 wires for full duplex transmission has been chosen, realized with echo cancellation and hybrid transformers.

Telettra, as one of the main suppliers of transmission equipment in Norway, has shown great interest in the basic rate access. In the period 1983 - 85, a 2-wire digital subscriber system was developed. This system applied analog echo cancellation and the line code used was Alternate Mark Inversion.

Later development has proved that the line codes 4B3T or 2B1Q may be better suited for the basic rate access. It has also been claimed that the echo cancellation must be done digitally. Accordingly, it was desirable with further research, and the following subjects were specified.

- Performance study of the line codes Alternate mark inversion, 4B3T and 2B1Q in a digital 2-wire system.

- Study of the two echo cancellation solutions with respect to performance and implementation.

- System study of the Network Termination 1 and production considerations of the (4B3T/2B1Q) Network Termination 1.

- Comparison of noise influence from Alternate Mark Inversion, 4B3T and 2B1Q transceivers on existing subscriber loop systems, with use of a near-end-crosstalk simulator, a 4B3T Network Termination 1 and a 2B1Q simulator.

- Realization of the near-end-crosstalk simulator and the 2B1Q hardware simulator.

First, in chapter 2, a description of the Integrated Services Digital Network is given with emphasis on the basic rate access. Chapter 3 provides a comparison of the three line codes while chapter 4 compares the two echo cancellation solutions. The system study of the network terminal can be found in chapter 5 and production considerations in chapter 6. The realization of the 2B1Q hardware simulator is described in chapter 7. Chapter 8 describes the rest of the simulator system and contains the results obtained from the measurements. Chapter 9 contains the conclusions.

2. INTEGRATED SERVICES DIGITAL NETWORK

This chapter will give an introduction to ISDN, the Integrated Services Digital Network.

Telephony and telex have been known as the major public telecommunication services for several decades. However, the development within electric information transfer through the last few decades has created several co-existing networks which carry different types of information. From the users point of view, this implies both expensive and inflexible communication solutions.

ISDN will play an important role in solving this situation. The expression "ISDN" was introduced by CCITT in 1979 and was defined as

> An ISDN is a network, in general evolving from a telephony IDN[1], that provides end-to-end digital connectivity to support a wide range of services, including voice and non-voice services, to which users have access by a limited set of standard multi-purpose user-network interfaces.
> (CCITT Red book, Fascicle III.5, p.3)

Generally, for the user, this means that ISDN will provide easy access to a multiplicity of services through a single network connection. New services will play an important role for the introduction of ISDN, and several telecommunication providers concentrate on creating new service concepts. Fig. 2.1. shows an example of present and future ISDN subscriber services.

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[1] Integrated Digital Network.

The first step towards ISDN has been the digitalization of the switching network by introducing digital transmission equipment and exchanges. ISDN can be introduced faster in those countries which have a high degree of digital exchanges. The ISDN carrier equipment will consist, in this early stage, of a combination of the existing digital networks. These may be packet networks, as well as public switched networks.

An overall majority of the subscriber loops uses copper wire with an expected lifetime of 30 - 70 years and will not be replaced for economical reasons within the near future.

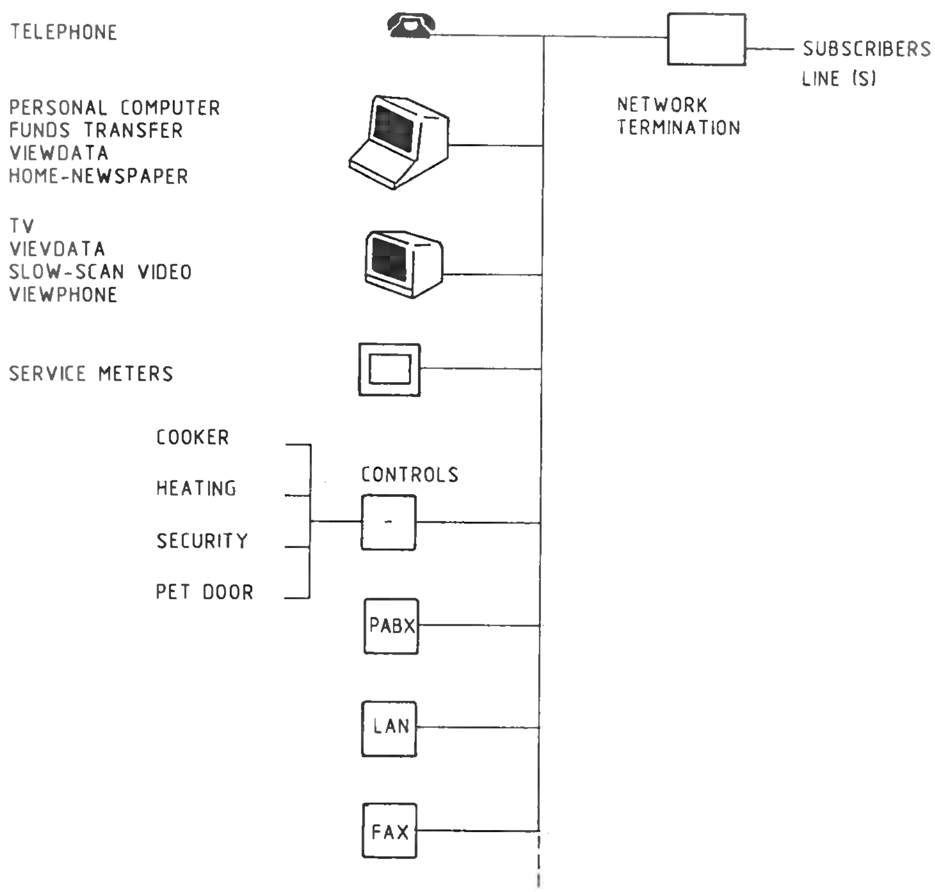


Fig. 2.1: Typical ISDN-subscriber services.

2.1. The reference model for the digital subscriber loop.

CCITT uses several reference models to describe ISDN. Each model consists of functional groups that contain functions naturally related to each other, and reference points that separate the groups. Fig. 2.1.1. shows the subscriber reference model [1].

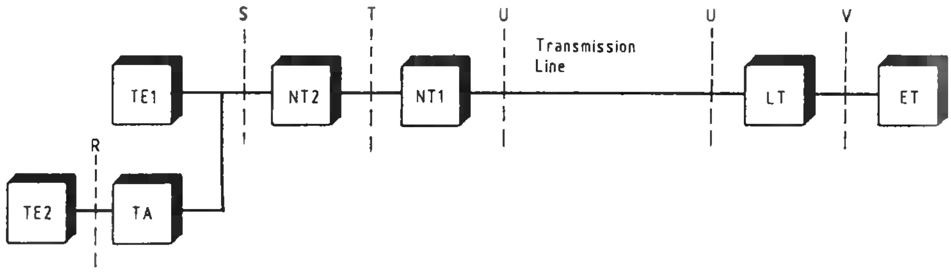


Fig 2.1.1: ISDN subscriber reference model.

The groups are defined as follows:

- NT1 (Network Termination 1) terminates the transmission line between the user and the network. It has functions for line maintenance, performance monitoring, timing,power transfer, multiplexing and activation/deactivation.

- NT2 (Network Termination 2) might consist of functions that comprise a Private Automatic Branch eXchange (PABX), a Local Area Network (LAN) or a terminal controller. Typical functions are protocol handling, multiplexing, switching, concentration and maintenance. For user-network interfaces without the necessity of these functions, NT2 is only a physical connection.

- TE1 (Terminal equipment type 1) consists of functions that comprise a terminal with ISDN interface i.e., a terminal capable of using ISDN directly.

- TE2 (Terminal equipment type 2) consists of functions that comprise a terminal without an ISDN interface i.e., a terminal that must use an adapter to connect to ISDN.

- TA (Terminal adapter) consists of functions that adapt TE2 to ISDN i.e., a device capable of converting between non-ISDN standards and ISDN.

- LT (Line termination) is similar to NT1 with the additional functions necessary at the exchange side of the subscriber loop.

- ET (Exchange termination) consists of functions related to the exchange i.e., protocol handling, concentrations and multiplexing.

The interfaces (or reference points) in the subscriber loop can briefly be described as follows:

- The R-interface is the interface between a non-ISDN compatible user equipment and the adaptor equipment. (An example of such an interface is RS-232-C.)

- The S-interface is the interface between the user terminal equipment and the network functions of the terminal.

- The T-interface is the interface which separates the network provider equipment from the user equipment. In the basic rate access, where NT2 is only a physical connection, the T- and S-interface are identical. This is the reason why the interface between NT1 and terminals often are indicated as the S/T-interface.

- The U-interface separates the terminal equipment from the exchange office equipment. At present this is a two-wire twisted pair, but in the future it may be replaced by an optic fiber.

- The V-interface is the interface between the line termination (LT) and the exchange termination in the public digital exchange.

2.2. Basic rate access

CCITT has recommended several transfer rates for the digital subscriber loop; the basic rate access and the primary rate access are the first being implemented.

The primary rate access consists of 30 (24 in North America and East Asia) user channels (called B-channels) and one signalling channel (called D-channel). Each channel has a capacity of 64 kbit/s.

The basic rate access consists of two user channels (2B) and one signalling channel (D). The B-channels have a capacity of 64 kbit/s and the D-channel of 16 kbit/s. The actual transmission-rate over the U-interface is 160 kbit/s (16 kbit/s for framing and housekeeping). This access rate is the minimum offered by ISDN. It replaces the analog telephone baseband and uses the same transmission media, the twisted pair. The basic rate access configuration consists of a NT1 for the twisted pair, no NT2 and allows 1 to 8 terminals (TE) to be connected to ISDN.

Three transmission techniques have been proposed for the U-interface to achieve full duplex digital transmission in the ISDN basic rate access.

- time division.
- frequency division.
- hybrid transformers and echo cancellation.

CCITT chose the latter technique. Full duplex two-wire data transmission can in principle be realized by hybrid transformers as a four-wire to two-wire interface. The problem with hybrid transformers is, however, that the transmitter leaks into its own receiver (see fig. 2.2.1.). The attenuation between transmitter and receiver can be as low as 10 dB, while the received signal from the other end may be attenuated 40 dB. In addition a part of the transmitted signal can be reflected due to imperfections in the two-wire circuit. These two "echoes" must therefore be cancelled to ensure a proper signal-to-noise ratio (SNR) at the receiver.

Echo cancellation can be realized by using adaptive digital filters (ADF) which produce an estimate of the echo (see fig.2.2.1.). The echo estimation is subtracted from the incoming signal. Since the echo canceller is used with different two-wire loops it must be adaptive. In addition to this it must be able to adapt when the transmission path changes with temperature variations.

Fig. 2.2.1. shows one end (subscriber-end) of the two wire digital subscriber loop. The other end (exchange-end) is identical. The signal $r(k) = s(k) + n(k) + e(k) - \hat{e}(k)$ is the control signal used with the adaptive filter. $s(k)$ is the wanted signal, $n(k)$ is additive noise, $e(k)$ is the echo signal, and $\hat{e}(k)$ is the echo replica.

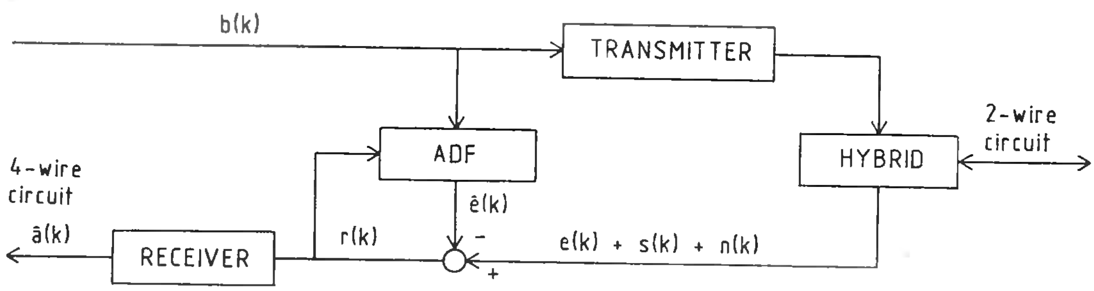


Fig. 2.2.1: Block diagram of ISDN basic rate access transceiver.

3. COMPARISON OF LINE CODES FOR BASIC RATE ACCESS

The ISDN basic rate access uses digital two-wire transmission with hybrid transformers and echo cancellation. This kind of digital full-duplex transmission requires the use of a line code to reach specified range (8 km) [19]. This chapter compares three different line codes proposed for the Norwegian network. The first line code proposed in 1983 was Alternate Mark Inversion (AMI) and the second was a 4B3T type called Modified Monitored Sum 4B3T (MMS43). MMS43 is used in the German pilot project today. The last arrived line code, 2B1Q, was proposed by British Telecom as a standard for the American National Standards Institute (ANSI). 2B1Q was chosen as a standard for US in 1986.

Section 3.1. gives an introduction to these line codes and a general analysis. Performance comparison is done in section 3.2., and section 3.2.1. discusses range limitation due to near-end-crosstalk. It has been stated that problems may arise from the coexistence of ISDN and the existing subscriber equipment. The Norwegian subscriber network contains a large number of carrier frequency equipment (the 1+1 system) used for frequency multiplexing of two subscribers on one line. Section 3.2.2. will give a general overview of the interference problem which may occur when introducing ISDN in the subscriber loop. More specific, section 3.2.3. discusses the possible near-end-crosstalk problem from ISDN to the 1+1 system, and an analysis is performed for the three line codes mentioned above.

3.1. Line code descriptions

Baseband digital transmission requires a reformatting of the digital signal, which is done by applying a line code. This section describes the pulse amplitude modulated line codes AMI, 4B3T and 2B1Q, and analyses are made with respect to the basic features below.

1. Transparency
2. Unique decodability
3. Efficiency
4. Favorable energy spectrum
5. Content of sufficient timing information

Transparency, unique decodability, and sufficient timing information are essential for digital two-wire transmission. High efficiency and favorable energy spectrum are desirable.

Efficiency is defined as the ratio of the information capacity of the message signal and that of the coded version. Information capacity [2] is given by

$$C = r \sum_i P_i \, {}^2\log \left(\frac{1}{P_i}\right) \quad \text{bit/s} \qquad (3.1.1.)$$

where r is the symbol rate and P_i is the probability of occurrence of the i-th symbol. The efficiency n is then

$$n = \frac{C_m}{C_c} \qquad (3.1.2.)$$

where C_m and C_c are the information capacity of the message signal and coded signal, respectively.

The transmission spectrum must be as narrow as possible, in order to reduce the influence of noise. It is also desirable to have a transmission spectrum with a minimum amount of low-frequency (lf) components and no dc component, due to cable characteristics and power feeding.

3.1.1. Alternate Mark Inversion (AMI).

Alternate Mark Inversion is a widely used line code. It is a three level linear code. The binary "zeros" are left uncoded, while the binary "ones" are encoded by a pulse with alternating polarity. (See example, table 3.1.1.)

Table 3.1.1: Binary - AMI conversion.

Binary data	0	1	1	0	0	1	0	1	0	0	0
AMI line code	0	+	-	0	0	+	0	-	0	0	0

(+ denotes a positive pulse A, while - denotes a negative pulse, -A, of the ternary signal.)

AMI is transparent, uniquely decodable and has no dc component. On the other hand the efficiency is low, η = 66,7%. The main disadvantage is the lack of timing information which occurs during long periods of "zeros". The minimum transmission bandwidth for ISDN basic rate access is 80 kHz ($B_T = 0.5\ r$, see also section 2.2).

Fig. 3.1.1. shows the power spectrum G(f) of the line code. G(f) is given by the following expression [3]

$$G(f) = r\ |P(f)|^2 \{R_a(0) + 2 \sum_{n=1}^{\infty} R_a(n) \cos(\frac{2\pi nf}{r})\} \qquad (3.1.3.)$$

where P(f) is the Fourier transform of the transmitted pulse and $R_a(n)$ is the amplitude auto-correlation function. AMI has the following amplitude correlation

$$R_a(n) = \begin{cases} A^2/2 & n = 0 \\ -A^2/4 & n = 1 \\ 0 & n \geq 2 \end{cases} \qquad (3.1.4.)$$

If a rectangular pulse is used, P(f) is given by

$$P(f) = \left(\frac{1}{r}\right) \operatorname{sinc}\left(\frac{f}{r}\right) \qquad (3.1.5.)$$

Substituting (3.1.4.) and (3.1.5.) into (3.1.3.) gives the following power spectrum for AMI

$$G(f) = \left(\frac{1}{r}\right) A^2 \operatorname{sinc}^2\left(\frac{f}{r}\right) \sin^2\left(\frac{\pi f}{r}\right) \qquad (3.1.6.)$$

The total power for AMI has been specified to be 10 dBm (see p. 35), which implies that A = 6.4 V.

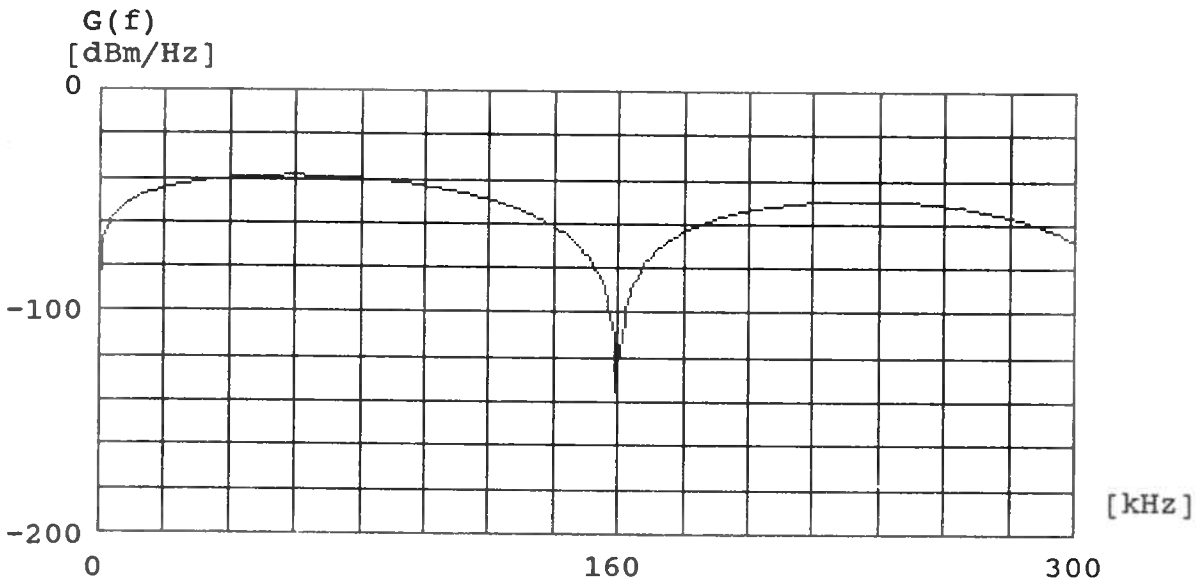


Fig. 3.1.1: Power spectrum of the line code AMI.

3.1.2. <u>4B3T</u>

4B3T is a three level block code. The coder converts blocks of four binary source digits into blocks of 3 ternary line digits. There are 16 possible binary blocks and 27 possible ternary blocks (2^4 resp. 3^3). Conversion can be done in several ways; this paper uses a code called MMS43. Table 3.1.2. shows the corresponding coding alphabets.

Table 3.1.2: The 4B3T code MMS43 [4].

Binary Word	Ternary Word A_{-1}	A_0	A_{+1}	A_{+2}	Word Digital Sum
0011	0 - +				0
0101	- 0 +				0
0110	- + 0				0
1110	+ - 0				0
1101	+ 0 -				0
1011	0 + -				0
1000	+ - +	+ - +	+ - +	- - -	+1;-3
1001	0 0 +	0 0 +	0 0 +	- - 0	+1;-2
1010	0 + 0	0 + 0	0 + 0	- 0 -	+1;-2
1100	+ 0 0	+ 0 0	+ 0 0	0 - -	+1;-2
0111	- + +	- + +	- - +	- - +	+1;-1
1111	+ + -	+ + -	+ - -	+ - -	+1;-1
0001	+ + 0	0 0 -	0 0 -	0 0 -	-1;+2
0010	+ 0 +	0 - 0	0 - 0	0 - 0	-1;+2
0100	0 + +	- 0 0	- 0 0	- 0 0	-1;+2
0000	+ + +	- + -	- + -	- + -	-1;+3

MMS43 has four different alphabets, A_{-1}, A_0, A_{+1}, and A_{+2}; which one to apply depends on the Running Digital Sum (RDS), defined [5] as

$$RDS(k) = \sum_{n=0}^{k} b_n + RDS(0) \; ; \; b_n = 1, 0, -1 \quad (3.1.7.)$$

associated to the k-th digit of the ternary sequence b_n. RDS(0) is a chosen constant. The alphabet index indicates the value of the corresponding RDS. (Clearly if RDS(k) = -1, then A_{-1} is chosen, RDS ϵ (-1, 0, 1, 2))

As well as AMI, MMS43 is also transparent and uniquely decodable. MMS43 is more efficient, η = 84,1 %. The minimum transmission bandwidth is smaller (60 kHz), due to the reduction of transmission rate (3/4 of the original rate). MMS43 is a balanced code, with almost no lf-components. (See fig. 3.1.2.)

Unlike AMI there is always enough timing information in a MMS43 signal, since there are never more than four equal consecutive pulses.

The amplitude correlation function for MMS43 is equal to

$$R_{a(n)} = \begin{cases} 2B^2/3 & n = 0 \\ -\,5B^2/32 & n = 1 \\ -\,3B^2/32 & n = 2 \\ 0 & n \geq 3 \end{cases} \qquad (3.1.8.)$$

where B is the amplitude of the output signal. If rectangular pulses are used, the following power spectrum can be found.

$$G(f) = \frac{1}{r} B^2 \mathrm{sinc}^2\left(\frac{f}{r}\right) \left[\frac{2}{3} - \frac{5}{16}\cos\left(\frac{2\pi f}{r}\right) - \frac{3}{16}\cos\left(\frac{4\pi f}{r}\right)\right] \qquad (3.1.9.)$$

The total power for 4B3T has been specified to be 9 dBm (see p. 35), which implies that B = 4.9 V.

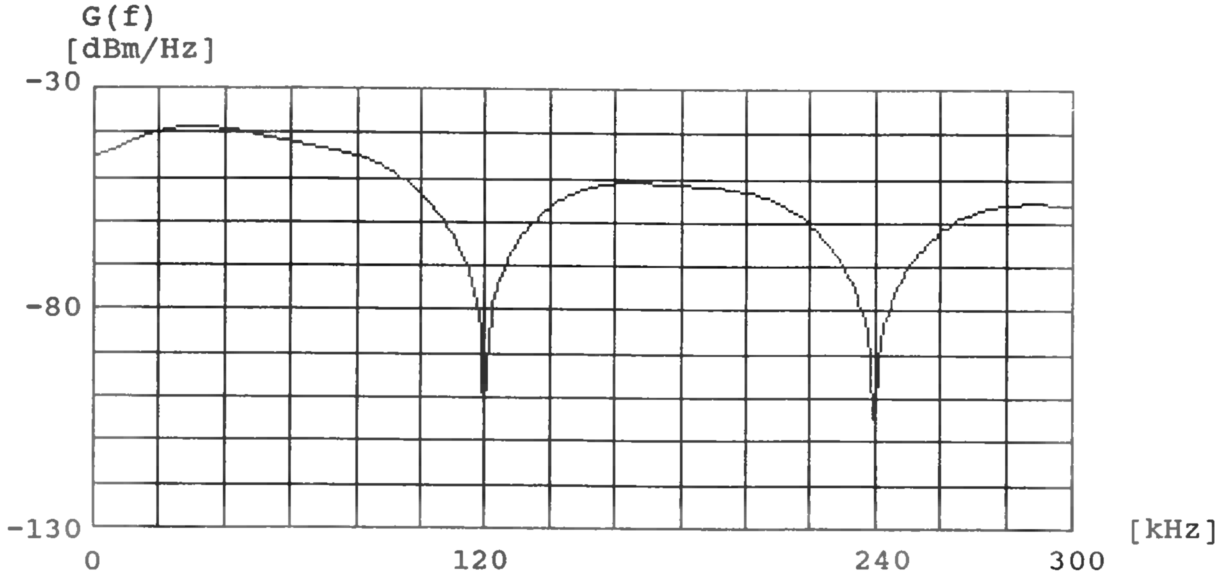


Fig. 3.1.2: Power spectrum of the line code MMS43.

3.1.3. 2B1Q

2B1Q is a four level line code, with no redundancy, n = 100 %. This line code converts two binary bits into a four level symbol. Table 3.1.3 shows the coding rules for 2B1Q.

Table 3.1.3: The binary-2B1Q conversion.

Binary data	Quaternary symbol
10	+3D/2
11	+ D/2
01	- D/2
00	-3D/2

The first bit is a sign bit, while the second determines the magnitude of the output symbol. D is the distance between the output levels.

The line code is transparent and uniquely decodable. The 2B1Q line code has two inconveniences; lack of timing information can occur and the power spectrum contains dc/lf components. This can be improved by introducing a scrambler before the line code converter and a high pass filter following it. A decision feedback equalizer (DFE) at the receiving-end must be used to recover the removed dc/lf components (see appendix-A).

The two disadvantages are largely compensated by the low transmission rate which is only half of the original rate. For the ISDN basic rate access, the minimum transmission bandwidth is only 40 kHz, which results in less cable-attenuation and near-end-crosstalk (NEXT).

The amplitude correlation for 2B1Q is given by

$$R_a(n) = \begin{cases} 5D^2/4 & n = 0 \\ 0 & n \geq 1 \end{cases} \qquad (3.1.10.)$$

resulting in the following power spectrum (the pulse shape is assumed to be rectangular).

$$G(f) = \frac{5}{4r} D^2 \operatorname{sinc}^2\left(\frac{f}{r}\right) \qquad (3.1.11.)$$

The total power for 2B1Q has been specified to be 10 dBm (see p. 35), which implies that A = 4.5 V.

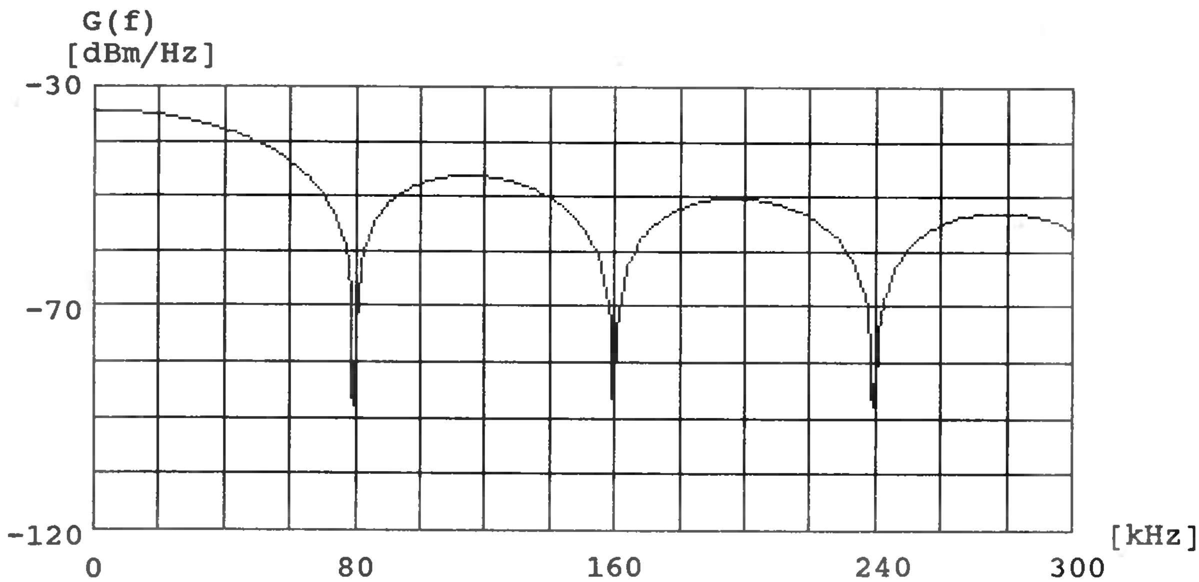


Fig. 3.1.3: Power spectrum of the line code 2B1Q.

3.2. Performance

This section contains a performance comparison of the three line codes with emphasis on the possible range and compatibility with existing subscriber systems. The main range limiting factor in digital subscriber loops is the NEXT. Section 3.2.1. discusses possible line ranges and section 3.2.2. possible interference from ISDN to existing subscriber services in general. Section 3.2.3. concentrates on interference with the 1+1 system.

3.2.1. Range limits due to near-end-crosstalk.

ISDN requires a bit error probability (P_{be}) of 10^{-7} [6] which can be translated into the following minimum SNR requirements (for calculations see appendix-B).

Table 3.2.1: SNR requirements at the receiver.

LINE CODE	SNR_{req} [dB]
2B1Q	21.1
MMS43	18.8
AMI	17.4

Fig. 3.2.1. shows the calculation model used to find the maximum subscriber line length.

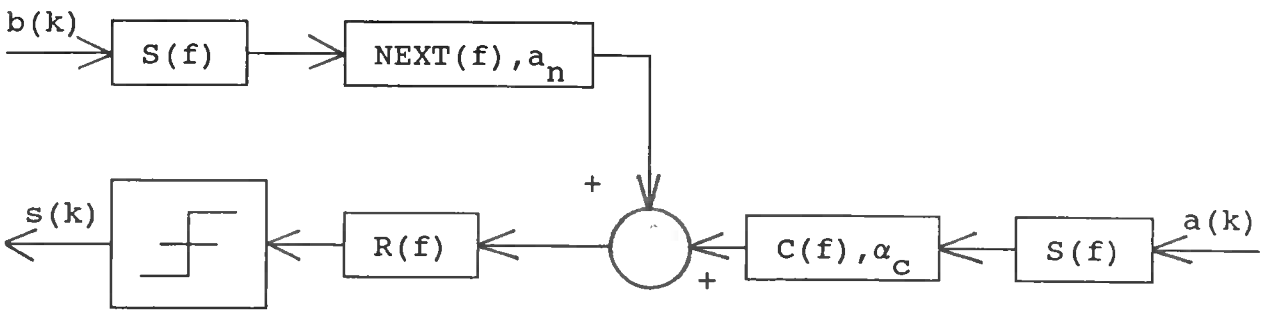


Fig. 3.2.1:Calculation model for maximum range limited by NEXT.

S(f), C(f), R(f), and NEXT(f) are the transfer functions of the transmitter filter, cable, receiver filter, and the NEXT-path, respectively. a(k) and s(k) are the transmitted and received signals while b(k) is the signal transmitted from near-end which interferes through the NEXT-path with the transmission of a(k). a_n and α_c are NEXT-attenuation and cable-attenuation per km. The cable-attenuation α_c and the NEXT-attenuation a_n are both frequency dependent (a_n see fig. 3.2.2. [7]). The maximum range L (if limited only by NEXT-noise) is given by [8]

$$L = \frac{a_n - SNR_{req}}{\alpha_c} \quad . \qquad (3.2.1.)$$

The maximum range can now be calculated. The minimum required SNR are found in table 3.2.1. The cable- and NEXT-attenuation values used to obtain the maximum range are based on measurements done by the Norwegian telecom research lab. (NTRL) [7]. Table 3.2.2. show the measured values and table 3.2.3. the obtained maximum range limited by NEXT.

Table 3.2.2: Cable-attenuation per km and NEXT-attenuation.

Freq. [kHz]	α_c(0.4mm) [dB/km]	α_c(0.6mm) [dB/km]	a_n [dB]
40	8.01	3.79	62.4
60	8.83	4.13	59.2
80	9.37	4.51	57.5

Table 3.2.3: Estimated range in km for three line codes limited by NEXT.

Line code	Range(0.4mm) [km]	Range(0.6mm) [km]
2B1Q	5.2	10.9
4B3T/MMS43	4.6	9.8
AMI	4.3	8.9

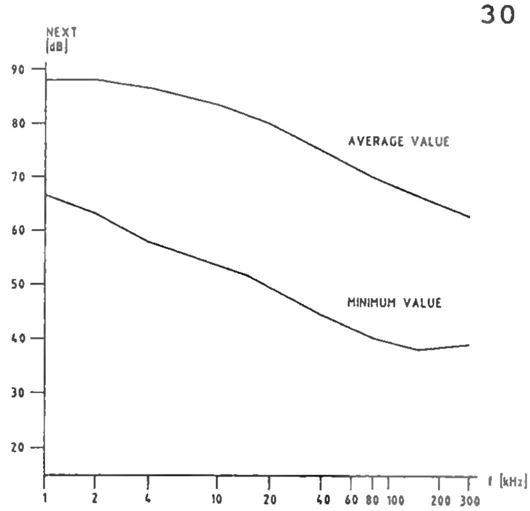


Fig. 3.2.2: NEXT-attenuation-frequency characteristic [7].

The values in table 3.2.3. are pessimistic values, since α_c increases and a_n decreases with increasing frequency, and the maximum frequency is assumed (=0.5 r). The NEXT values reported by NTRL are pair-to-pair attenuations. To obtain the total NEXT power in a 10 pair cable, 12.3 dB must be subtracted from these values [9].

There are also other noise sources in the system. The most decisive are inter-symbol interference, residual echo, impulse noise and implementation loss. The practical range will therefore be less than given in table 3.2.3. These effects can be included by adding an extra margin to the NEXT. In literature a margin of 10 to 15 dB is given [9], [10]. Table 3.2.4. includes a 10 dB margin.

Table 3.2.4: Estimated range for three line codes with a 10 dB noise margin included.

Line code	Range(0.4mm) [km]	Range(0.6mm) [km]
2B1Q	3.9	8.3
4B3T/MMS43	3.4	7.4
AMI	3.2	6.7

3.2.2. Interference to subscriber loop equipment

This section will consider interference from ISDN basic rate access to existing subscriber equipment. Reverse interference has not been considered, due to the high transmission level specified for the ISDN basic rate access (9 - 11 dBm) [17,22].

Table 3.2.5. shows existing services in Norway which use the subscriber loop as transmission media [20].

Table 3.2.5: Existing services in the subscriber loop.

Services	Frequency range [kHz]	Transmission output level [dBm]
Analog telephony	0 - 3.4	0
Alarm services	4.38 - 5.1	- 6
Telex (high level)		60V, 40 mA
Telex (low level)	0.98/1.18, 1.65/1.85	- 8
Facsimile	1.3 - 2.1	-10
Telefax	1.8	-10
Teledata	1.7 +/- 0.4	-10
Low speed data transmission (0.6 - 9.6 kbit/s)	0 - 4.0	- 6
PCM[2] 704/2048 kbit/s	0 - 2048	14
Tariff impulses	16	7
Fault location signals	24,75,84,91,100	-
Connections for local radio stations	0 - 15	6 - 9
Data transmission:		
48 kbit/s	0 - 80	0
64 kbit/s	0 - 128	7
1+1 system (frequency carrier equipment)	20 - 87	0

[2] Pulse-Code Modulation (PCM)

No interference problems are assumed for subscriber systems using other frequency bands than ISDN for transmission. These systems are: analog telephony, 704 kbit/s and 2048 kbit/s PCM, low speed data transmission, facsimile, telefax, teledata, telex and alarm services.

Tariff impulse equipment and fault location equipment are systems which are compatible with ISDN due to equal transmission level and the burst nature of the transmission.

A number of connections (less than 500) used by local radio stations exists in the subscriber loop. Possible interference with these systems is assumed to be acceptable due to the limited number of systems.

Low level data transmission 48/64 kbit/s will be disturbed by ISDN. However the number of these connections is limited, and will probably be replaced by an ISDN connection at a lower price. The price of an intercity (200-300 km) 64 kbit/s connection in Norway today is US Dollar (USD) 18.600 per year, while ISDN basic rate access (2 * 64 kbit/s) will cost approximately USD 450 plus normal telephony tariff. This leaves the 1+1 system which is discussed in next section.

3.2.3. Interference to carrier frequency equipment

The frequency modulated carrier equipment, 1+1 system [10,11], consists of a low frequency subscriber loop (300-3700 Hz) and a high frequency subscriber loop (HFSL). The HFSL contains two frequency modulated channels with center frequencies at 28 kHz (from subscriber to exchange) and 79 kHz (reverse direction), see fig. 3.2.3. The baseband, W, and the transmission bandwidth, B_T, are 3.4 kHz, and 16 kHz, respectively.

The transmit power of the frequency carrier is 0 dBm. Maximum loop attenuation is 40 dB at 79 kHz, which corresponds to a maximum range of 8 km (0.6 mm cable).

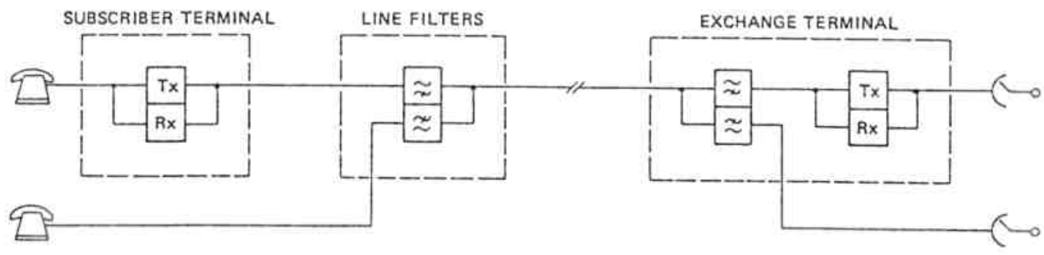


Fig. 3.2.3: Block diagram of a 1+1 telephone system [10].

The Norwegian network contains of more than 100.000 1+1 systems, representing a value of approximately USD 1 billion. It is therefore essential that the 1+1 system can exist in an ISDN environment, since it is neither technical nor economical desirable to replace these units when ISDN is introduced.

The digital telephone transmitter level based on echo cancellation must, according to Norwegian pre-ISDN specifications, [12] not exceed 0 dBm, while the manufacturers of ISDN ICs today seem to prefer 9-11 dBm.

Thus, it is important to analyze whether this increased transmission level will cause any problem. The Norwegian specification for the 1+1 system [11] requires noise-immunity against noise power less than - 47 dBm.

Fig. 3.2.4. and fig. 3.2.5. shows the theoretical average power density spectrum of AMI, MMS43, and 2B1Q in the 28 kHz and 79 kHz band, respectively.

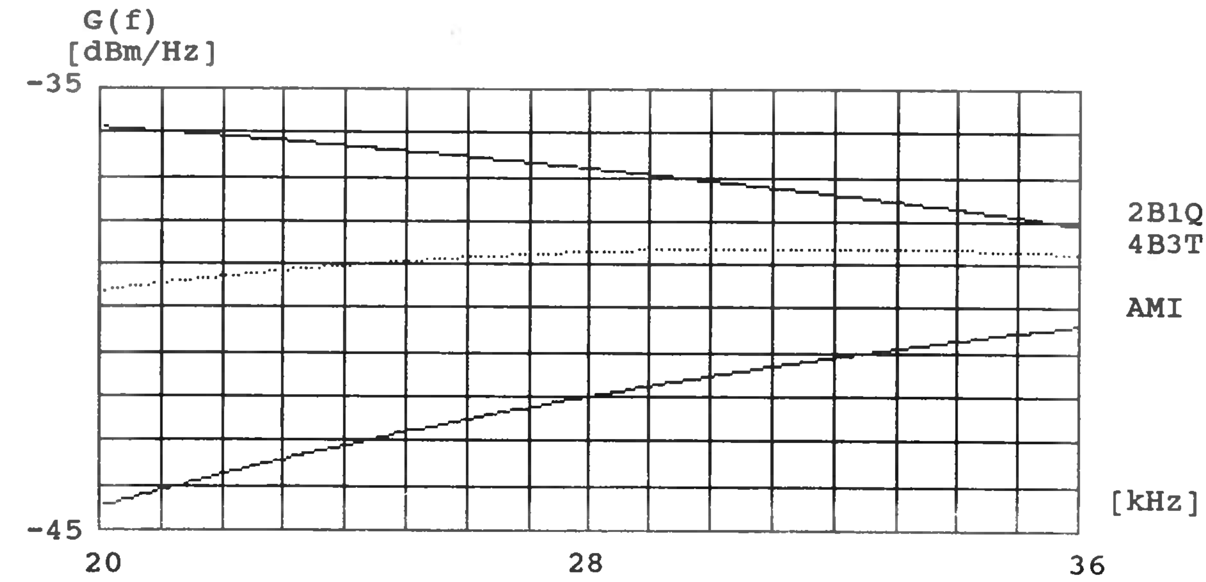


Fig. 3.2.4: Theoretical average power density spectrum of the three line codes in the 28 kHz band.

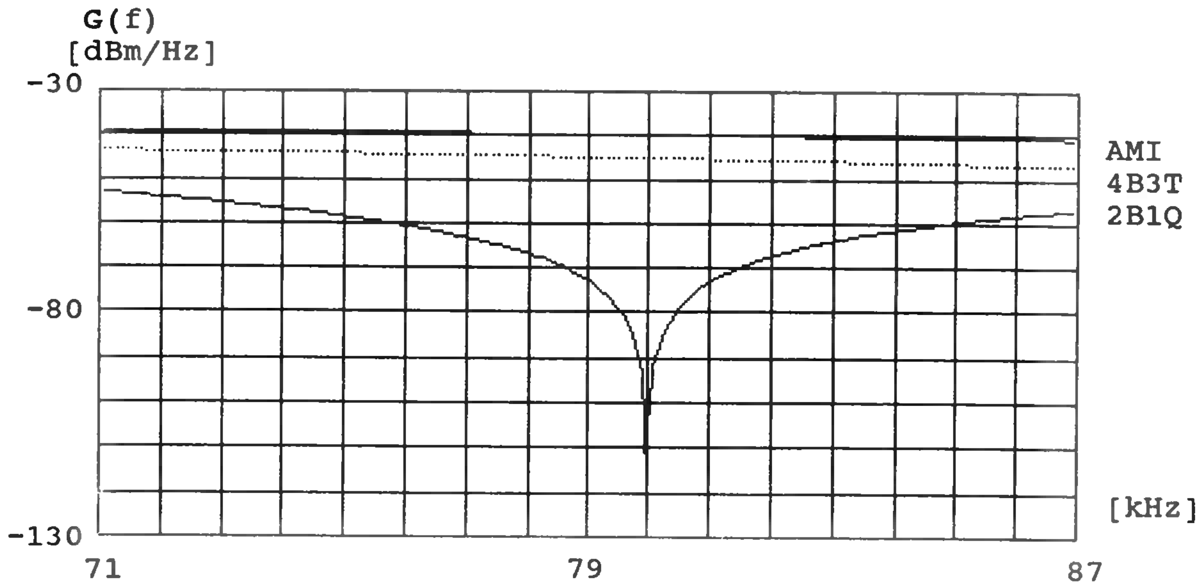


Fig. 3.2.5: Theoretical average power density spectrum of the three line codes in the 79 kHz band.

Table 3.2.6. shows the total power of the three line codes in the frequency-bands used by the 1+1 system (see appendix-D for calculations).

A total transmission level of 11 dBm has been specified for 2B1Q [17], while 9 dBm has been specified for MMS43 [22]. The transmission level specified for AMI in 1983 was 0 dBm, but 10 dBm is used in this thesis for comparison.

Table 3.2.6. contains the minimum required NEXT-attenuation with respect to the -47 dBm noise-power-limit. The minimum required NEXT-attenuation is given at 100 kHz, both for the subscriber-side and the exchange-side of the 1+1 system. (See table 3.2.2. for typical NEXT-attenuation.)

Table 3.2.6: Estimated average power in the frequency band used by the 1+1 system for the three line codes 2B1Q, MMS43, and AMI. Minimum required NEXT-attenuation, at 100 kHz, with respect to -47 dBm noise-power-limit.

LINE CODE	AV. POWER 28 kHz BAND [dBm]	AV. POWER 79 kHz BAND [dBm]	Min. req. NEXT-att. Subscr. end [dB]	Exchange end [dB]
2B1Q	5.2	-17.2	28.5	44.2
MMS43	3.2	- 2.8	42.7	42.2
AMI	0.1	2.3	47.8	39.1

The results in table 3.2.6. indicate that ISDN can be introduced in the subscriber loop without causing any problem for the existing 1+1 system, and that 4B3T is the best performing line code.

On the other hand it must be noted that noise outside the frequency band used for these calculations, will also contribute to the total noise in the demodulated channel. This is due to the fact that non-ideal filters and hybrid transformers are used in the 1+1 system. This is also the reason why measurements were done with an actual system. The results of the measurements and the system description can be found in chapter 8.

4. ECHO CANCELLATION IN THE BASIC RATE ACCESS

In chapter 2 a basic rate access transceiver has been described which uses a digital adaptive filter to produce a replica of the echo signal. This digital filter needs a digital control signal, while the input signal is analog. An analog-to-digital converter is therefore needed in the echo canceller. Two configurations have been proposed for the echo canceller. The difference between these two echo cancellers is the location of the analog-to-digital convertor. The most straight forward one has the analog-to-digital converter at the input, prior to any signal processing. This configuration is called digital echo canceller, since the detection of the incoming signal is done digitally. In the other configuration detection is performed from an analog signal, and only the signal needed to control the adaptive process of the echo canceller is converted into a digital representation. This configuration is called analog echo canceller.

This chapter will describe and compare the two echo canceller configurations for the ISDN basic rate access. Section 4.1. contains the description of the two configurations, section 4.2. considers implementation precautions of the echo canceller, and section 4.3. compares the two echo cancellation structures.

4.1. Signal processing structure

This section will describe the signal processing structure of two different echo canceller (EC) configurations. The difference between the two configurations is whether to do the subtraction in a digital or analog way. The key point will be the analog-to-digital converter (ADC) in the EC. Section 4.1.1. describes the analog configuration and section 4.1.2. the digital EC configuration.

4.1.1. Analog echo cancellation

The configuration of the analog EC is shown in fig. 4.1.1. The echo replica ê(t) is subtracted from the (lowpass filtered) input signal f(t) (= e(t) + s(t) + n(t)), producing a signal r(t) (see chapter 2). The r(t) signal is converted into the digital representation r(k), which is used for the updating of the filter coefficients in ADF. The ADF will produce ê(k) which must be converted into an analog signal by a digital-to-analog converter (DAC) and a lowpass filter. The lowpass filter must remove the periodic frequency components of ê(k) to ensure proper detection of r(t).

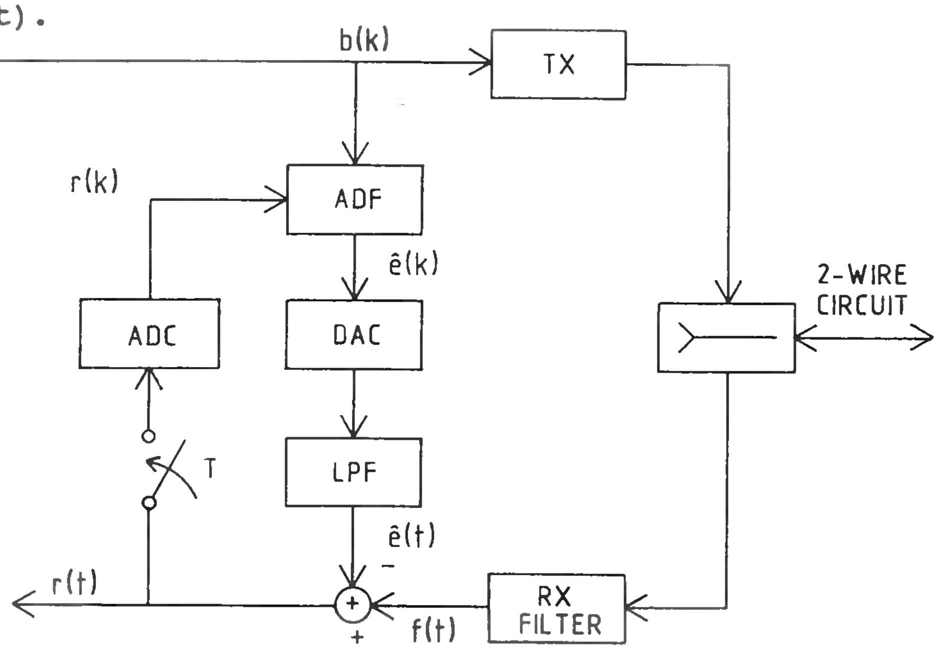


Fig. 4.1.1: Analog echo cancellation configuration.

The detection of r(t) must still be possible even when the strongest echo-signal occurs simultaneous with the weakest s(t). Which requires that the lowpass filter must be almost ideal, these severe filter requirements make this configuration untractable [13]. On the other hand only the echo cancellation process is affected by the finite precision of the ADC. The natural occurrence of dithering in this configuration [13] can be exploited so that the finite precision of the ADC doesn't negatively affect the echo cancellation process. s(t) + n(t) is used as dithering-signal. The result is that a simpler ADC can be used.

4.1.2. Digital echo cancellation

This section describes the digital EC (see fig. 4.1.2.). The lowpass filtered received signal f(t) is converted into the digital representation f(k). ê(k) is subtracted from f(k) to remove the echo component. The result r(k) is used to control the ADF directly. The ADF produces an echo replica ê(k) which also can be used directly. Detection of the data is done digitally since r(k) is a digital signal.

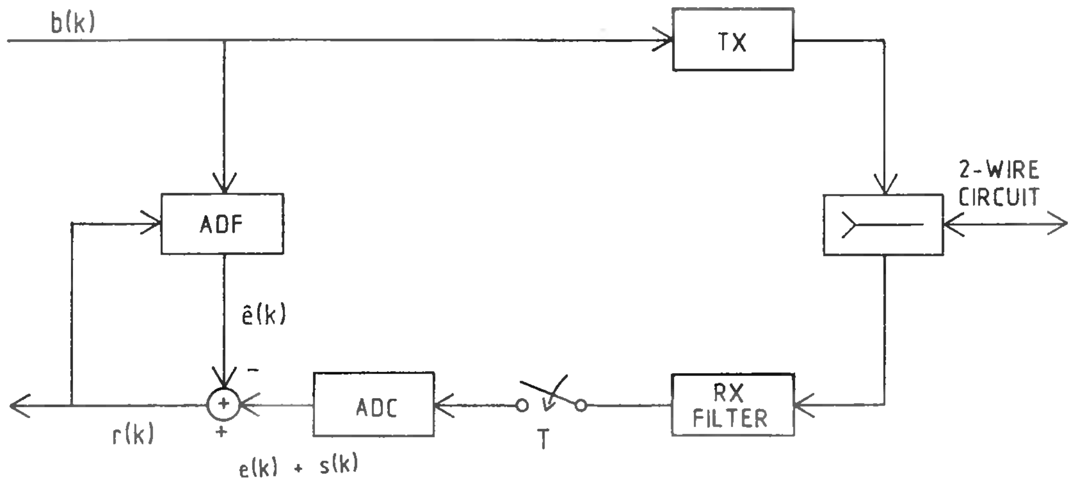


Fig. 4.1.2: Digital echo canceller.

The ADC must be designed so that neither the detection of transmitted data nor the echo cancellation are affected by the finite precision of the ADC. The result is a more complex ADC (more bits necessary). The advantages of the digital EC are clear; no DAC or complex lowpass filter is needed. The periodic frequency components of ê(k) are compensated by the periodic frequency components of e(k) [13].

4.2. Implementation considerations

Integrated circuits (ICs) are sensitive to the physical arrangement of their components and interconnections. Analog ICs are more sensitive than digital ICs. In digital systems noise does not accumulate as it would in a comparable analog system. It is therefore necessary to take certain precautions when integrating analog circuitry in a digital IC.

First of all, analog and digital power lines must be separated, since the digital logic circuits will produce large noise spikes due to switching. Secondly clock lines must be shielded from their environment in order to prevent them from picking up or injecting noise. A possible shielding arrangement is shown in fig. 4.2.1. The lines used are metal lines. Such shielding can also be used to separate analog and digital circuits.

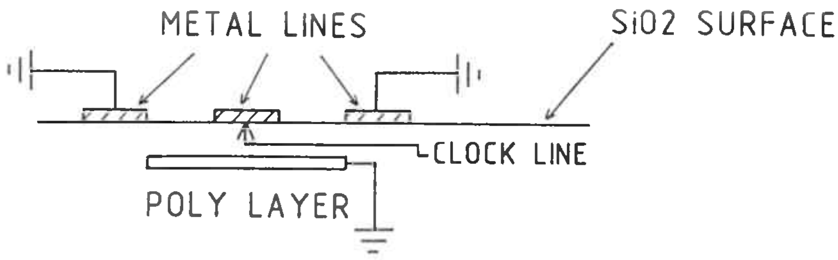


Fig. 4.2.1: Shielding arrangement for clock lines.

Another noise source in the IC is due to coupling through the substrate. This effect can be reduced by minimizing the noise into and out of the substrate, which can be achieved by using clean power supply to bias the substrate and by shielding the substrate from noisy lines.

In addition, one must consider: nonlinear distortions, accuracy of matched elements, op-amp offset voltages and sensitivity to process variations. Summarily one can conclude that analog circuits in a digital environment cause extra design problems. The amount of analog circuitry should therefore be minimized in the echo canceller.

4.3. Comparison

The ADC of the analog configuration can be optimized with respect to the echo canceller. The digital configuration must be dimensioned to meet the requirements both from the echo canceller as well as from the detector. This means that the noise due to residual echo and quantization in the digital EC must be in the same order as residual echo noise in the analog EC.

The digital EC requires an ADC with dynamic range and linearity corresponding to 12 bit accuracy [14]. In [6] an ADC is described which meets these requirements for 60 kHz baseband transmission (line code 4B3T). The ADC is realized using pulse-density-modulation (PDM) technique, described in section 5.6. This type of ADCs is also suited for implementation in a digital environment.

The basic rate access transceivers usually apply decision feedback equalizers (DFE) (see appendix-A). The structure of these equalizers is similar to the ADF. In a digital EC, the DFE and the ADF can be combined for simpler implementation.

In an analog EC, the penalties for simpler ADC are an extra DAC, complex lowpass filtering of both the echo replica and the input f(t), and severe implementation requirements. The analog EC configuration was designed to reduce the complexity of the ADC. The introduction of PDM makes this configuration superfluous, and the digital EC canceller is therefore the best alternative.

5. THE NETWORK TERMINATION 1 (NT1).

This chapter will describe the structure of the present Network Termination 1 based on the most common 4B3T transceiver. Section 5.1. will give a general system overview, followed by a more specific description of the functional blocks.

5.1. System overview.

The NT1 can be divided into two parts, the U-interface circuit (UIC) and the S-interface circuit (SIC). A block diagram of the former is shown in fig. 5.1.1. and the latter in fig. 5.1.2. The interface between these two sections will be called system interface, and is specified by the manufacturer of the circuits. Even if this interface has not been specified by CCITT, an industry standard has developed. That is the ISDN Oriented Modular interface (IOM) which was first introduced by Siemens. IOM interface is a standard 4-wire local interface for the interconnection of ISDN devices within the ISDN basic rate access. It consists of a receive and a transmit data line, an 8 kHz frame signal and a 512 kHz to 8.192 MHz data clock. The clocks are generated in the UIC.

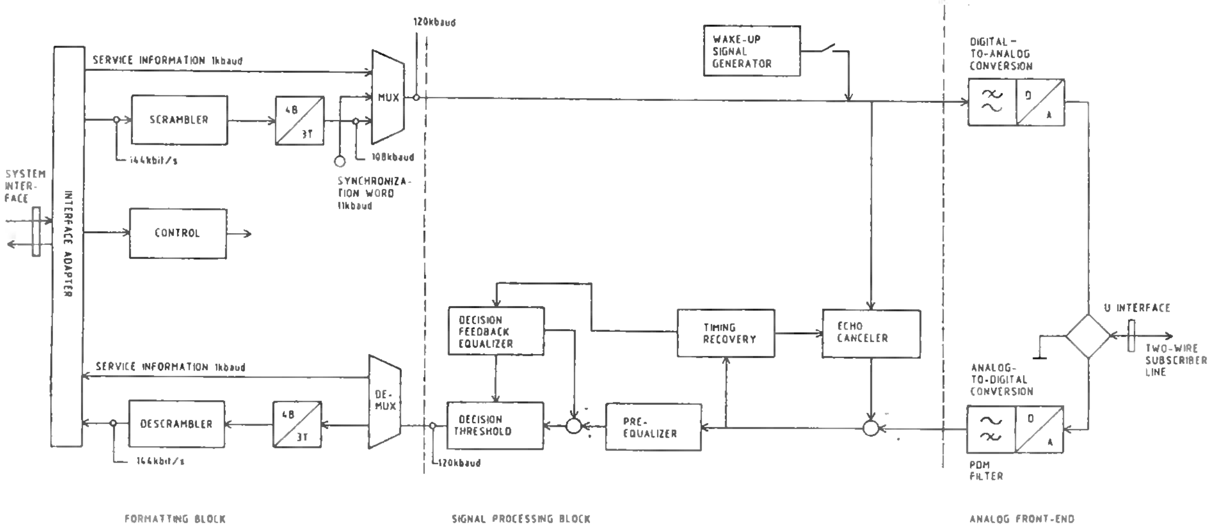


Fig. 5.1.1: Block diagram of the U-interface circuitry [22].

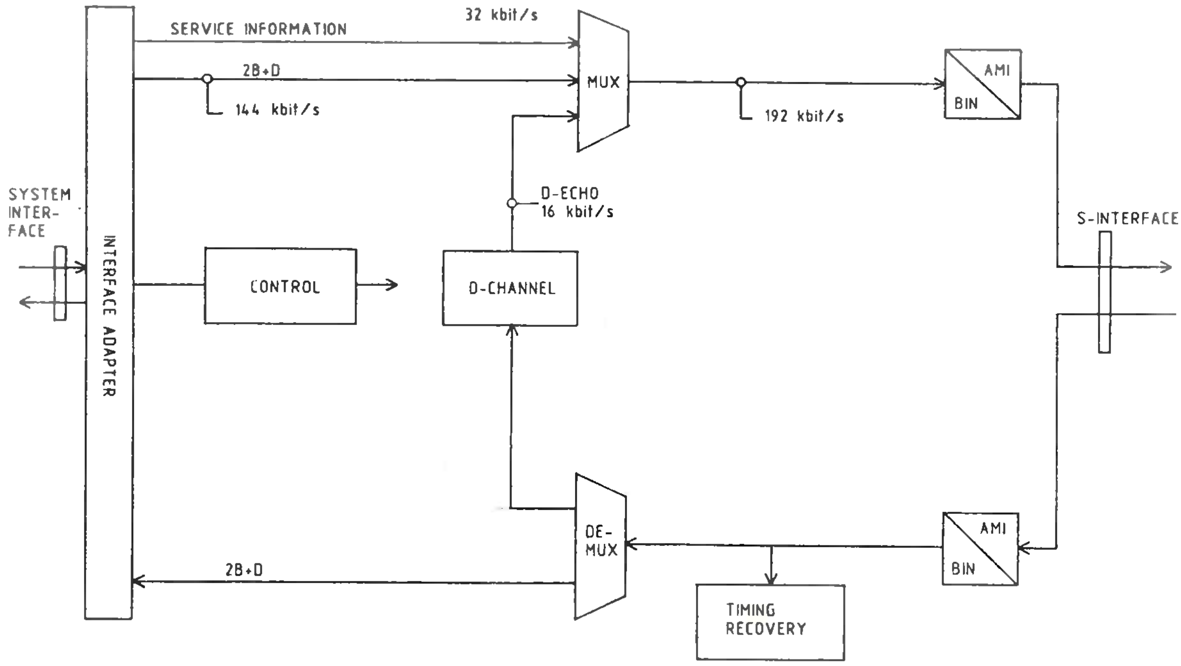


Fig. 5.1.2: Block diagram of the S-interface circuitry.

The task of the UIC is to transport 144 kbit/s of data (2B + D) over the subscriber loop intended for 4 kHz analog transmission. The transmission must be full duplex. As mentioned in chapter 2, the UIC functions are realized by hybrid transformers and adaptive echo cancellation. The UIC can be divided into three sections. These are the analog front-end, the digital signal processing block and the formatting block.

The analog front-end consists of a DAC and an ADC. Its task is to provide conversion between the subscriber loop and the digital signal processor. Section 5.6. will describe the ADC.

The digital signal processor consists of an echo canceller, a timing recovery unit, an equalizer and a decision threshold detector. The echo canceller is built up by an adaptive digital filter which produces a replica of the echo. This echo replica is subtracted from the incoming signal. Section 5.2. describes the adaptive digital filter and section 5.3. algorithms which can be used to update the filter coefficients. The timing recovery unit uses correlation technique to provide both frame- and bit-timing (see section 5.4.).

The function of the equalizer can be compared to that of the echo canceller. The transmission characteristic of the subscriber loop generates inter-symbol interference when data are transmitted at a high rate (for 4B3T 120 kbaud). The detection of which symbol that has been received must not be influenced by data received before or after the decision point. Any such information must be cancelled to ensure that the right signal is detected (see appendix-A).

The functions of the formatting block are multiplexing, scrambling and encoding. The outgoing data signal (144 kbit/s) is scrambled and encoded before it is multiplexed with 1 kbaud service information and 11 kbaud synchronization information. (The synchronization words are used for timing recovery.) The incoming data stream is subject to inverse operations. More details about scrambling/descrambling, frame structure and synchronization can be found in section 5.4., 5.5. and 5.7.

The SIC provides the four-wire S/T-interface used to link the terminals to an ISDN. 144 kbit/s of data are multiplexed with 32 kbit/s of service information and the echo channel (16 kbit/s). The echo channel is used to retransmit the incoming D-channel. This is done to provide D-channel access control when more terminals are connected to the S-interface. The line code used is inverse AMI.

5.2. The adaptive digital filter

As shown in previous chapters, the echo canceller consists of an adaptive digital filter and a subtraction point. The adaptive digital filter needs a control signal to optimize to echo replica $\hat{e}(k)$. The optimum is reached when $\hat{e}(k) = e(k)$, where $e(k)$ is the echo signal (see also fig. 5.2.1.).

This section will prove that the gradient of E $\{r(k)^2\}$ with respect to c_k can be used as a control signal [13]. Here E{•} is the mathematical expectation, c_k the coefficient vector of the digital adaptive filter and r(k) the residual signal. $r(k) = s(k) + n(k) + e(k) - \hat{e}(k)$, where s(k) is the transferred signal and n(k) the additive uncancellable noise.

Let us assume the following: the transmission path of the echoes can be described by the finite duration impulse response g(n) with $g(n) = 0$ for $n < 0$ and $n \geq N$. A transversal digital filter with N variable coefficients $c_0(k)$, $c_1(k)$, ... $c_{N-1}(k)$ is used to produce an echo replica $\hat{e}(k)$. The transmission path between the transceivers can be described as an impulse response h(n) with $h(n) = 0$ for $n < 0$ and $n \geq M$. The input data to the subscriber transmitter are b(k), while the input data to the exchange transmitter are a(k).

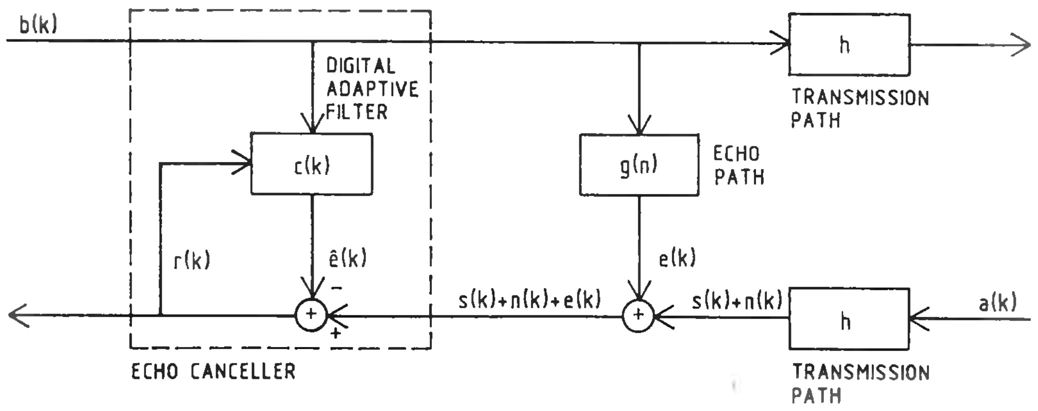


Fig. 5.2.1: Mathematical model of the echo canceller.

The following vectors can be defined:

- The exchange input vector: $a_k^T = [a(k), a(k-1) \ldots . a(k-N+1)]$
- The subscriber input vector: $b_k^T = [b(k), b(k+1) \ldots . b(k-N+1)]$
- The echo path vector: $g^T = [g(0), g(1) \ldots . g(N-1)]$
- The coefficient vector: $c_k^T = [c_0(k), c_1(k) \ldots . c_{N-1}(k)]$
- The transmission path vector: $h^T = [h(0), h(1) \ldots . h(M-1)]$

where $(\bullet)^T$ denotes the transposed vector. The echo signal, the echo replica and the received signal s(k) (subscriber) can now be expressed as the vector quantities

$$e(k) = \mathbf{b}_k^T \, g \qquad (5.2.1.)$$

$$\hat{e}(k) = \mathbf{b}_k^T \, c_k \qquad (5.2.2.)$$

$$s(k) = \mathbf{a}_k^T \, \mathbf{h}. \qquad (5.2.3.)$$

The residual signal after subtraction can be rewritten as

$$r(k) = \mathbf{b}_k^T \, (g - \mathbf{c}_k) + u(k) \qquad (5.2.4.)$$

where u(k) = s(k) + n(k). The mean square value of the residual signal is given by

$$\sigma_r^2 \triangleq E\{r(k)^2\}$$

$$= E\{[b_k^T(g - c_k) + u(k)]^2\} \qquad (5.2.5.)$$

Let $\mathbf{b}_k$ and $\mathbf{c}_k$ be statistically independent. (This is not true, but the dependence is very weak and can be neglected, see also (5.2.11.).) If, in addition, a(k), b(k), and n(k) are statistically mutually independent and the data inputs $b_k = \pm 1$ have the same probability of appearance, then (5.2.5.) can be rewritten as

$$\sigma_r^2 = E[(g - c_k)^T b_k b_k^T (g - c_k)] + S_u \qquad (5.2.6.)$$

$$= (g - c_k)^T \, (g - c_k) + S_u \qquad (5.2.7.)$$

where $S_u = E\{u(k)^2\}$. σ_r^2 has its absolute minimum, equal to S_u, for $\mathbf{c}_k = \mathbf{g}$, which implies that $e(k) = \hat{e}(k)$ and $r(k) = u(k)$ (no residual echo!). The gradient of σ_r^2 with respect to c_k must therefore be considered to find this minimum. The gradient is defined as

$$\textbf{grad}\ \sigma_r^2 = [\frac{\delta\sigma_r^2}{\delta c_0(k)} \ \cdots \ \frac{\delta\sigma_r^2}{\delta c_{N-1}(k)}]^T \qquad (5.2.8.)$$

and by substituting (5.2.7.) into (5.2.8.) it becomes

$$\text{grad}\ \sigma_r^2 = -2\ (g - \mathbf{c}_k) \qquad (5.2.9.)$$

or alternatively if (5.2.4.) and (5.2.5.) are substituted

$$\text{grad}\ \sigma_r^2 = -2E[\ r(k)\ \mathbf{b}_k\ ] \qquad (5.2.10.)$$

The gradient is a direct measure of the difference between the actual and the optimum value of the adaptive filter coefficients (5.2.9.). On the other hand **grad** σ_r^2 is a direct function of the signal $r(k)$ and b_k, which are available within the echo canceller (5.2.10.). This can also be seen as a measure for the correlation between $r(k)$ and $\mathbf{b}_k$. If **grad** σ_r^2 is zero, then $r(k)$ and b_k are uncorrelated, which means that $r(k)$ has no average echo component.

The adaptive digital filter can then be controlled by the gradient of σ_r^2. The updating of the filter coefficients are done iteratively according to

$$\mathbf{c}_{k+1} = \mathbf{c}_k - \alpha\ \textbf{grad}\ \sigma_r^2 \qquad (5.2.11.)$$

(The constant α must be small in order to ensure weak dependence between $\mathbf{b}_k$ and $\mathbf{c}_k$.) By applying (5.2.11.) the value of σ_r^2 is minimized, which implies minimum residual echo.

5.3. Algorithms

For practical applications an estimate of the gradient of σ_r^2 is used. This section will describe three algorithms for estimation of grad σ_r^2. The first algorithm gives an approximation of the gradient by a finite time average over K data intervals. The updating is performed once per K data intervals. This algorithm is called the correlation algorithm [13] and is given by

$$\text{grad } \sigma_r^2 \approx - \frac{2}{K} \sum_{l=k+1-K}^{k} r(l)\ b_l. \qquad (5.3.1.)$$

The second algorithm gives a rougher approximation. K is here set to one, deleting the time averaging. This algorithm is called the stochastic iteration algorithm. Updating is done each data interval, the gradient is given by [13]

$$\text{grad } \sigma_r^2 \approx - 2\ r(k) b_k. \qquad (5.3.2.)$$

The last algorithm is called sign algorithm, and gives the roughest approximation of the gradient. Only the sign of r(k) is used in combination with b_k [13].

$$\text{grad } \sigma_r^2 \approx - 2\ \text{sign}(r(k))\ b_k \qquad (5.3.3.)$$

This algorithm gives the simplest updating procedure for the filter coefficients.

5.4. Timing recovery.

For easy VLSI-realization the timing recovery at the U-side of the NT1 is done by a correlation digital phase-lock loop (DPLL). The principle of the DPLL is based on the use of an 11 bit synchronization word which is transmitted in each 1 ms frame. The received signal is continuously correlated with the synchronization word. The maximum of the synchronization word correlation function can be used for frame synchronization. The synchronization word correlation function is given by

$$cs(k) = \sum_{i=0}^{10} r(k - i) \quad sw(i + 1) \qquad (5.4.1.)$$

where r(k) is the received signal after echo cancellation and sw(i) the synchronization word. The correlation function approximately reproduces the shape of the receive step impulse response. The optimum sampling position is assumed to be 1/4 symbol period in advance of the maximum [14]. This is achieved by using the synchronization word correlation difference

$$ph(k) = cs(k_0 + 1) - cs(k_0 - 1). \qquad (5.4.2.)$$

If the levels are not the same, the feedbacked difference ph(k) can be used to adjust the sampling instant. The DPLL increases or decreases the 120 kHz symbol period (4B3T) by one 15.36 MHz clock cycle until the difference is zero.

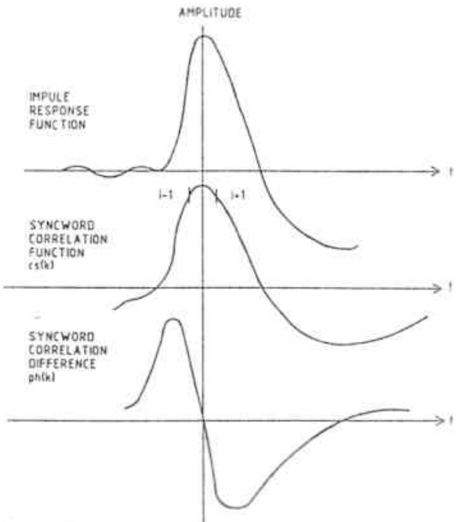


Fig. 5.4.1: Principle of the digital phase-lock loop.

The synchronization words to and from the NT1 are different, so that an uncancelled echo cannot disturb the DPLL (see section 5.7.). The part of the frame containing data may of course possibly contain a bit sequence equal to the synchronization word. This will have no effect on the DPLL, since resynchronization will only occur after the synchronization word has been found 64 times in another position than the one expected [22].

5.5. Scrambler/descrambler.

The scrambler function in the UIC is implemented to remove auto-correlation and cross-correlation of the transmitted and received signal. The exchange-end (LT) scrambler polynomial is

$$1 \ (+) \ Z^{-5} \ (+) \ Z^{-23}$$

and the subscriber-end (NT1) scrambler polynomial is

$$1 \ (+) \ Z^{-18} \ (+) \ Z^{-23}.$$

The descrambler polynomials are, for the exchange-end (LT)

$$1 \ (+) \ Z^{-18} \ (+) \ Z^{-23}$$

and for the subscriber-end (NT1)

$$1 \ (+) \ Z^{-5} \ (+) \ Z^{-23}$$

where (+) denotes the logical XOR function.

5.6. Analog-to-digital conversion with the use of pulse density modulation.

The ADC at the analog front-end showed in fig. 5.1.1. (UIC) must meet severe requirements. First it must have the dynamic range and linearity corresponding to 12 bit accuracy at a rate of 120 kHz (line code 4B3T). Secondly it must be easy to implement in digital MOS-technology. Conventional ADC techniques, employing mainly analog circuits, will not be able to satisfy these demands.

A technique which has shown to be successful is pulse-density-modulation. Pulse-density-modulation technique consists of a simple analog unit followed by a complex digital processing unit. The ADC is based on an analog 1-bit-coder which converts the analog input to an oversampled digital 1-bit stream. The 1-bit-code is then subject to digital processing, which results in linear PCM.

The design requirement for NT1 dictates a minimum signal-to-noise ratio of 72 dB for a signal-bandwidth of 60 kHz [6]. This can be achieved by using a second order delta-sigma-modulator (DSM) with an oversampling factor of 128. The low precision required of the analog components used in the DSM makes it suitable for integrated circuits implementation. Fig. 5.6.1. shows a block diagram of an ADC applying pulse-density-modulation.

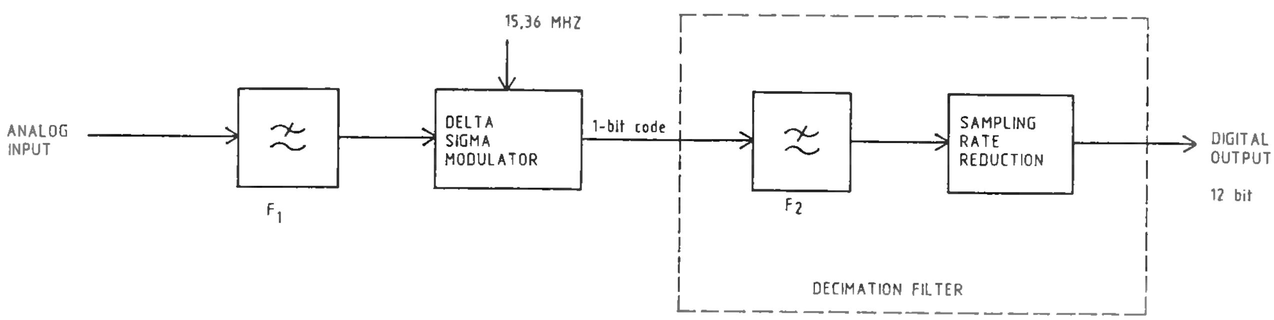


Fig. 5.6.1: Block diagram of an ADC using pulse-density-modulation.

The lowpassed analog input is converted into a digital 1-bit stream by the DSM. The sampling rate f_s is 15.36 MHz. The lowpass filter F_1 must have flat frequency characteristic below 60 kHz and reject frequencies above 7.68 MHz ($f_s/2$). The input filter can therefore be made quite simple in contrast to the input filter of a conventional ADC. The signal processing is done by a decimation filter. The decimation filter converts the oversampled 1-bit-code into a 12-bit word at the Nyquist rate (120 kHz). The frequency components above 60 kHz must therefore be removed (F_2) in order to avoid aliasing when the sampling rate is reduced.

5.6.1. Delta-sigma-modulation.

The basic DSM consists of an analog filter H(s), a quantizer and a digital-to-analog converter in a feedback loop, see fig. 5.6.2. The quantizer in the 1-bit DSM is realized by a comparator followed by a flip-flop. The task of the DAC is to convert the digital 1-bit code into an analog representation, typical two levels + V and - V, which is subtracted from the analog input. The filter H(s) must have a lowpass characteristic.

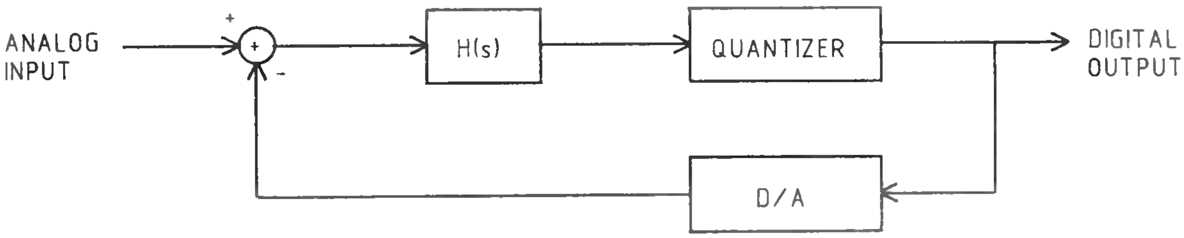


Fig. 5.6.2: Basic DSM configuration.

The dominant noise in the DSM is due to quantization, introduced in the comparator. The noise transfer function will have a high-pass characteristic, while the signal transfer function is lowpass.

The noise transfer function will shape the quantization noise spectrum, which is assumed to be white noise (uncorrelated), at the expense of amplifying the total noise power. Fig. 5.6.3. shows the shape of a typical DSM noise spectrum. If oversampling is applied, the noise can be moved away from the baseband. The signal transfer function must be flat in the baseband region.

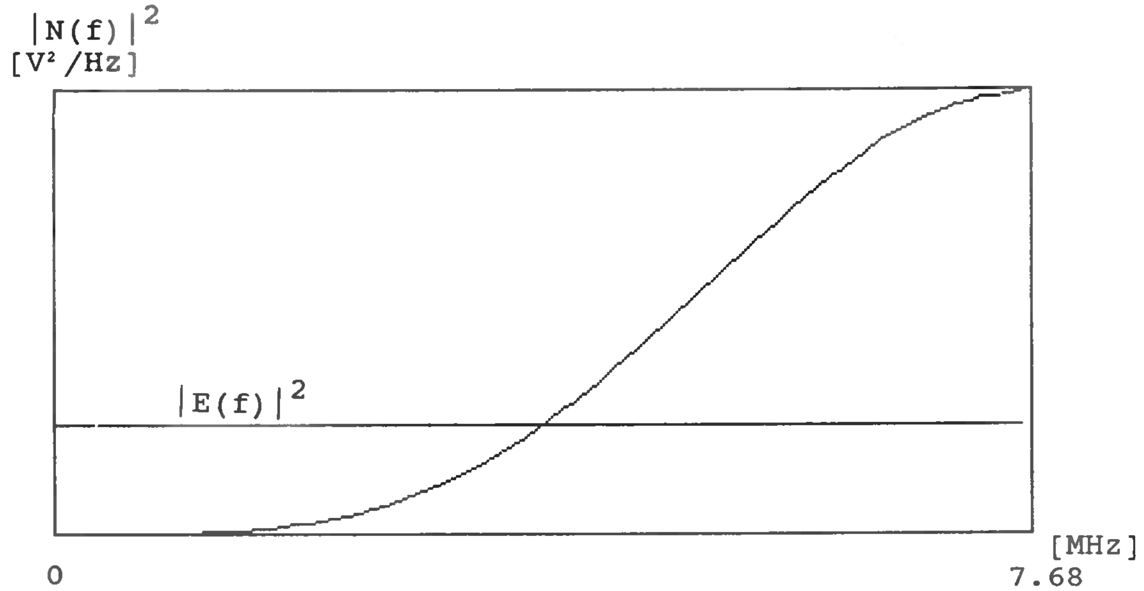


Fig. 5.6.3: Typical DSM noise-spectrum relative to additive noise density spectrum $|E(f)|^2$ (not to scale).

In UIC the DSM is implemented using switch-capacitor technique [6]. For switch-capacitor implementations a discrete-time model would be appropriate, since only the values of the input signal at the sampling instants are of interest (assuming that the input signal is constant over the sampling interval $T = 1/f_s$). Fig. 5.6.4. shows a digital model of the basic DSM.

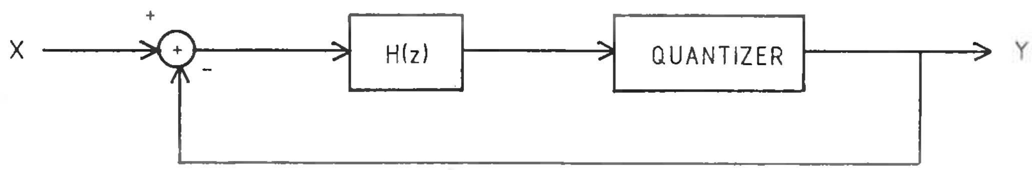


Fig. 5.6.4: Basic digital DSM.

The DSM may have more feedback loops, resulting in higher order transfer functions. More than two is however useless, since stability is not guaranteed.

The signal-to-noise ratio of the DSM is directly dependent of the oversampling factor. The second-order DSM provides a 15 dB/octave increase in SNR [15]. This is a theoretical value, measurements done at the Delft University of Technology (DUT) with a second-order DSM have shown that 11.3 dB/octave can be achieved. The 72 dB SNR requirement can be met with a second-order DSM applying an oversampling of 128.

Fig. 5.6.5. shows a mathematical model of a digital second-order DSM; blocks used to guarantee stability are left out. The quantization noise introduced by the comparator is modelled as addition of white noise, E. The z-transform of the output signal is then

$$Y(z) = z^{-1} X(z) + z^{-1} (1-z^{-1})^2 E(z) \qquad (5.6.1.)$$

The added noise is assumed to be white with spectral density $|E(f)|^2$ and σ_e^2 the power. The spectral density of the noise at the output is then

$$|N(f)|^2 = |E(f)|^2 \; | H(f) |^2 \qquad (5.6.2.)$$

where $H(f) = z^{-1} (1-z^{-1})^2 \, |_{z=e^{2\pi jfT}}$, the noise transfer function, thus

$$|N(f)|^2 = 16 \; |E(f)|^2 \; \sin^4 (\pi fT). \qquad (5.6.3.)$$

The noise power in the baseband is then given by [15]

$$N_0 = 2\, T \int_0^{f_b} |N(f)|^2 \; df. \qquad (5.6.4.)$$

If a three-term expansion for sine is used and $f_b\ T << 1$, then the noise power is approximately

$$N_0 = 24\ \pi^4\ \sigma_e^2\ (2\ f_b\ T)^5. \qquad (5.6.5.)$$

Equation (5.6.5.) indicates that the noise in the baseband decreases by 15 dB if the sampling rate f_s is increased one octave.

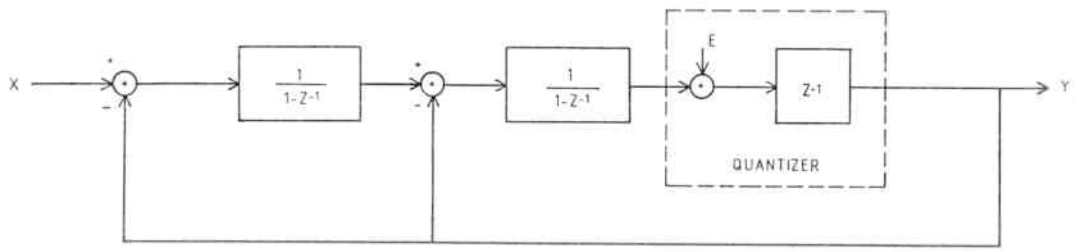


Fig. 5.6.5: Digital model of the second order DSM.

5.6.2. Digital decimation (lowpass) filter

As shown above, the DSM will produce a digital representation of the analog input, and the quantization noise will be shaped so that the major part is moved to high frequencies outside the baseband. The high rate 1-bit code is however unsuitable for signal processing such as echo cancellation and equalizing. A decimation filter is used to reduce the sampling rate. The filter must suppress high frequency noise such that only a negligible amount is aliased back into the baseband. The reduction of the sampling rate must produce a greater definition, that is, multibit linear PCM.

The total number of bits to be processed will be reduced since the PCM SNR increases 6 dB for each bit extended onto a word, while DSM SNR only increases approximately 11 dB for every doubling of the sample rate. In the case of the UIC, 72 dB SNR must be satisfied which corresponds to 12 bit PCM.

A digital filter which has been reported [16] to achieve this kind of sampling rate reduction is given by the following transfer function

$$D(z) = \frac{(1 - z^{-N})^2}{[(1 - z^{-1})N]^2} \qquad (5.6.6.)$$

which has a triangular impulse response. The amplitude transfer function of this filter is given by

$$| D(z=e^{2\pi jfT}) | = \frac{\sin^2(\pi NfT)}{\sin^2(\pi fT)} \qquad (5.6.7.)$$

For UIC applying a 4B3T line code, the baseband will be 60 kHz (see chapter 3). The DSM used operates at 15.36 MHz [14] (T=1/15.36 MHz) which correspond to an oversampling of 128 ($=f_s/2f_b$). A 256-th order (N=128) digital filter can be used [14], and the filter has attenuation zeros at multiplies of 120 kHz. Fig. 5.6.6. shows the amplitude transfer function D(f).

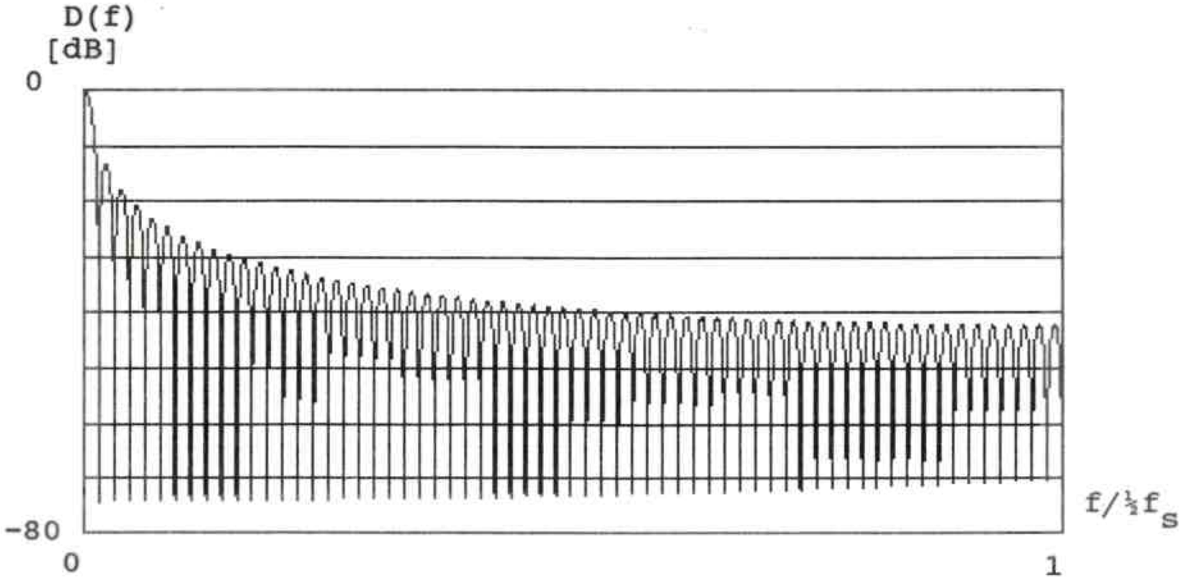


Fig. 5.6.6: Amplitude transfer function of the decimation filter.

The spectral density of the noise at the filter output, before sampling reduction can then be found by equation (5.6.3.) and (5.6.7.)

$$|N'(f)|^2 = |N(f)|^2 \quad |D(f)|^2 \qquad (5.6.8.)$$

Fig. 5.6.7. shows the noise spectral density before and after filtering.

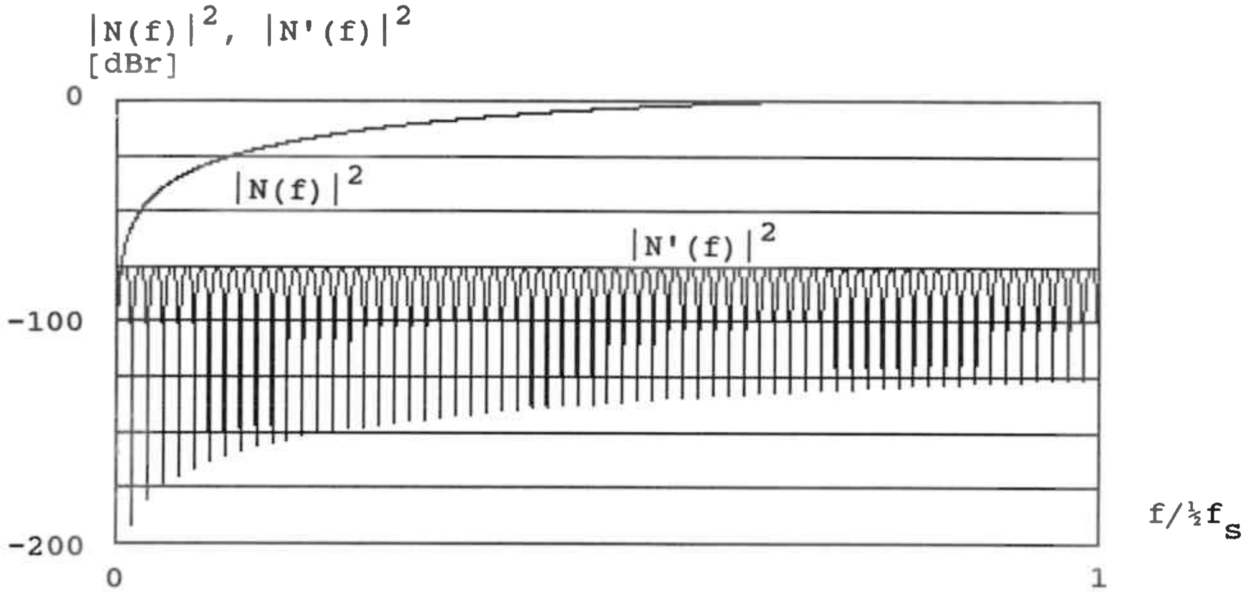


Fig. 5.6.7: Noise spectral density before and after filtering.

After filtering, the sampling rate can be reduced, and linear PCM is produced by omitting all samples between those at the wanted rate. Realization alternatives can be found in [16], [23] and [24].

5.7. Frame structure in NT1.

Frame structure at the S-side of NT1:

The transmission rate at the S-side of NT1 is 192 kbit/s [1]. The frame used consists of 48 bits with a duration of 250 μs. The frame structures are different for each direction of transmission. Fig. 5.7.1. shows the frame structure at the reference point S/T [1].

The bits in the frame are:

F	-	framing bit
L	-	DC balancing bit
D	-	D-channel bit
E	-	D-echo-channel bit
F_A	-	auxiliary framing bit
N	-	bit set to binary value $N = \overline{F}_A$
B1	-	B-channel one bit
B2	-	B-channel two bit
A	-	activation bit
S1,S2	-	reserved for future standardization.

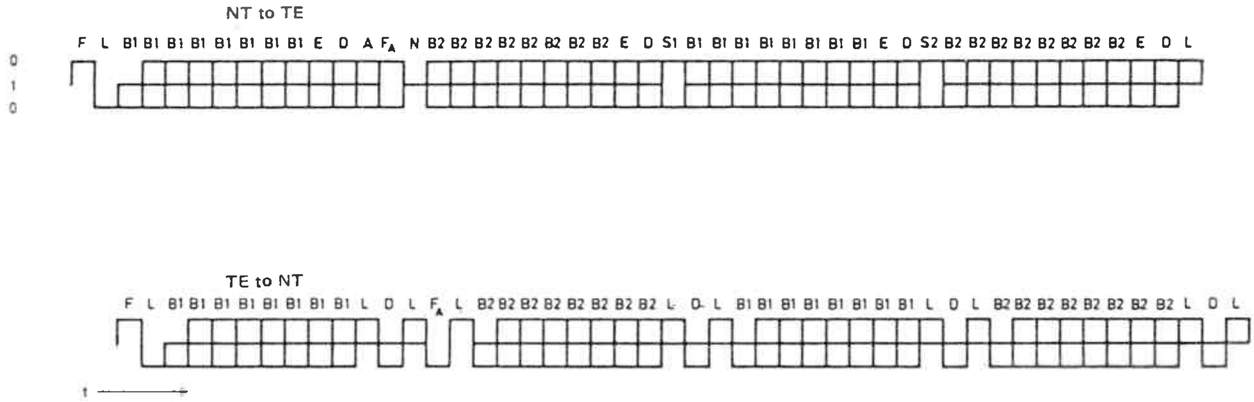


Fig. 5.7.1: Framing structure at the S-side of NT1.

The line code used is inverse AMI, i.e. binary zeros are coded in an alternating way with the voltage levels +A, -A, while binary ones correspond to zero voltage output-level.

The D-echo-channel bit is used for D-channel access control when a number of terminals are connected to the S-interface. The NT1 must echo the incoming D-channel bit. The terminals can then use the D-echo-channel bit to control whether D-channel access is granted.

The F_A bit is used to guarantee secure framing.

Frame structure at the U-side of NT1:

The transmission rate at the U-side is 160 kbit/s, resulting in 120 kbaud if the 4B3T line code is used. The 1 ms frame used at the U-side consists of a 11 symbol synchronization word, 108 symbol data and 1 symbol maintenance information.

The synchronization words are different in each direction of transmission. The synchronization words are [22]:

From LT to NT1:	+ + + - - - + - - + -
From NT1 to LT:	- + - - + - - - + + +

where + and - denotes a positive and a negative 4B3T symbol, respectively.

5.8. Activation/deactivation.

Activation can be initiated from both TE and LT, while deactivation only from the LT. A set of info signals has been defined for both the S-interface and the U-interface. These info signals are given in table 5.8.1. [1] and table 5.8.2. [19].

Table 5.8.1: Definition of info signals over the S-interface.

Signals from NT1 to TE	Signals from TE to NT1
Info S0 - no signal	Info S0 - no signal
Info S2 - Frames with all bits of B, D and D-echo channels set to binary zero. Bit A is set to zero.	Info S1 - A continuous signal with the following binary pattern: 0011111100 is coded as inverse AMI.
Info S4 - Synchronized frames with operational data on B, D and D-echo channels. Bit A is set to one.	Info S3 - Synchronized frames with operational data on B and D channels.

Table 5.8.2: Definition of signals over the U-interface.

Signals from NT1 to LT	Signals from LT to NT1
Info U1W - a ternary sequence: ++++++++-------- is repeated 16 times, resulting in a 7.5 kHz signal with duration 2.13 msec.	Info U2W - equals U1W.
Info U1A - binary continuous zeros, which are scrambled and encoded. No frame, ternary zeros instead of synchronization word.	Info U2A - equals U1A (used by repeaters).
Info U1 - equals U1A with synchronization word instead of ternary zeros.	Info U2 - equals U1.
Info U3 - equals U1 with binary ones instead of zeros before scrambling.	Info U4H - equals U3.
Info U5 - Binary data from the digital interface. Frame.	Info U4 - equals U5.
Info U0 - Ternary continuous zeros, no frame, no signal level.	Info U0 -

Activation from TE.

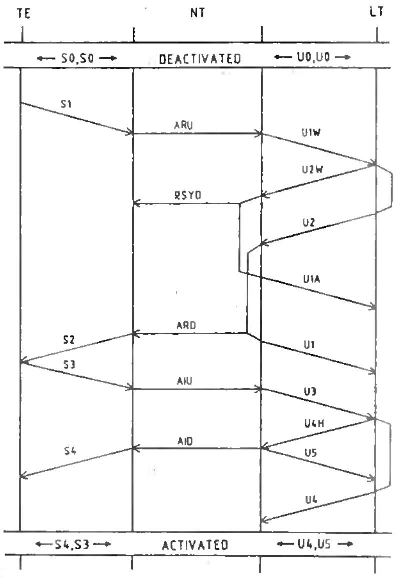


Fig. 5.8.1: Activation from TE.

A typical activation sequence is shown in fig. 5.8.1. TE starts activation by sending info S1, the NT1 responds by sending the awake signal U1W, and LT acknowledges with a similar signal U2W and U2 (delayed). U2W and U2 result in info U1A and U1 from the NT1, both delayed.

LT and NT1 are considered synchronized after U1 and NT1 starts sending empty frames (S2) to TE, which responds by transmitting operational data.

TE and NT1 are now considered synchronized. NT1 transmits an activation indication (U3) which is acknowledged by a similar signal U4H from the LT. The NT1 and the LT start transmitting data (U5, U4 and S4) and the link becomes transparent.

Activation from LT

Activation from LT is done in a similar way, as showed in fig. 5.8.2.

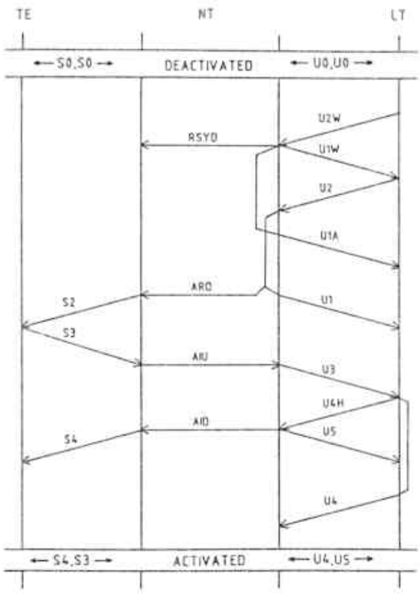


Fig. 5.8.2: Activation from LT.

Deactivation is shown in fig. 5.8.3., LT initiates deactivation by sending info U0, i.e. removing the line signal. NT1 responds by doing the same (U0) and sending a deactivation request (S0) to the TE, which is acknowledged (S0). Info S0 is equal to no signal, the deactivation is complete.

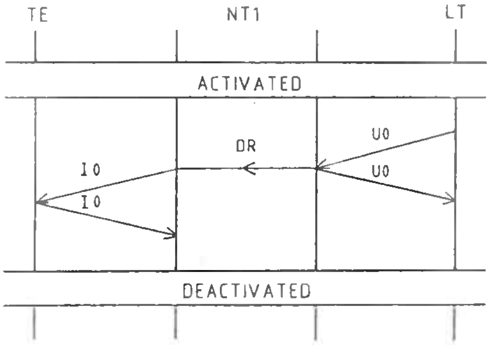


Fig. 5.8.3: Deactivation.

5.9. Power

In addition to the transmission unit described in the previous sections, the NT1 consists of a power unit. This power unit supplies both the NT1 and the connected terminals [19].

Phantom feeding over the S-connection is used for the terminals. The power unit derives its power from two sources, i.e. the (220 volt) main power supply and the exchange over the U-connection (maximum 96 volt DC).

Two operation modes are defined [19], emergency and normal operation. Table 5.9.1. shows the power unit specifications.

Table 5.9.1: Specifications for NT1 power unit.

	Normal operation		Emergency operation	
	idle [mW]	active [mW]	idle [mW]	active [mW]
Power to be supplied to S-interface.	≥ 4500	≥ 4500	≥ 45	≥ 410
Power consumption at U-interface.	≤ 30	≤ 350	≤ 90	≤ 800

As can be seen from table 5.9.1., a "power-down" mode is required during the deactivated state of NT1.

The NT1 must go into emergency state when the main supply drops below 85 % of its normal value. The polarity of the voltage (40 volt) fed over the S-connection will then be reversed to alarm the terminals of the emergency condition. At least one terminal must be powered up in emergency state to ensure the basic functions of a telephone terminal, i.e. call set-up, release, ringing function and audio functions. Terminals which derive their power from the main supply are not affected by the emergency condition.

6. PRODUCTION CONSIDERATIONS OF THE NETWORK TERMINATION 1.

This chapter will consider requirements for industrial production of NT1 and available ICs which can be implemented to satisfy these demands. Section 6.1. will discuss the technical specifications and economical requirements, while section 6.2. will consider available components for NT1 implementation.

6.1. Technical specifications and economical requirements for NT1

The NT1 is generally speaking a converter between S- and U-interface. The S-interface has been specified in CCITT recommendation I.430 [1]. For the U-interface however no CCITT recommendation has been specified due to the discussion over the line code, and the world has been divided in two. ANSI has chosen the line code 2B1Q for North America while the driving force in Europe, Deutsche Bundespost (DBP), has specified 4B3T for the pilot project in Germany. The NTA has also specified the ISDN basic rate access, the Norwegian specification is a translation of the DBP specification with two changes. NTA requires protection against higher over-voltage and stricter demands concerning the NT1 power. (A NT1 design according to the NTA specifications will also satisfy the German requirements.)

In addition to the technical specifications the NT1 must satisfy economical demands. The basic rate access which includes the NT1, will replace the analog telephone for the subscriber. The ISDN basic rate access must therefore cost approximately the same as the analog telephone. The price of basic ISDN installation in Norway has been set to twice the cost of installing an analog telephone, i.e. USD 450 (exclusive telephone receiver). The NT1 represents only a small part of the cost. In Norway it has been indicated that the price of NT1 in the middle of 1990s must be less than USD 150.

This is the framework established for the production of NT1. The following section will discuss realization possibilities for NT1.

6.2. Components for ISDN basic rate access

ISDN transceivers for the basic rate access were developed and implemented in a NT1 and LT by Telettra in 1985. It was however decided to use external components for realization of NT1 in 1988. This section will give an overview of the available components.

The U-interface circuit.

The pilot projects in different European countries today use line code 4B3T, while the North American standard is 2B1Q. It is possible that Europe will follow ANSI and specify 2B1Q as a standard after the pilot projects, but no decision has been taken yet. One must therefore consider the possibilities of two NT1 solutions.

None of the manufacturers has made the 2B1Q UIC pin-compatible with the 4B3T UIC. It is however possible to produce a PCB containing the basic functions equal for both NT1 solutions, while the UIC, the hybrid transformer and the line transformers are located on a daughter-board (see fig. 6.2.1.). The line code can now be selected by installing the right daughter-board. The daughter-board solution makes it possible to change the U-interface without having to replace the whole NT1.

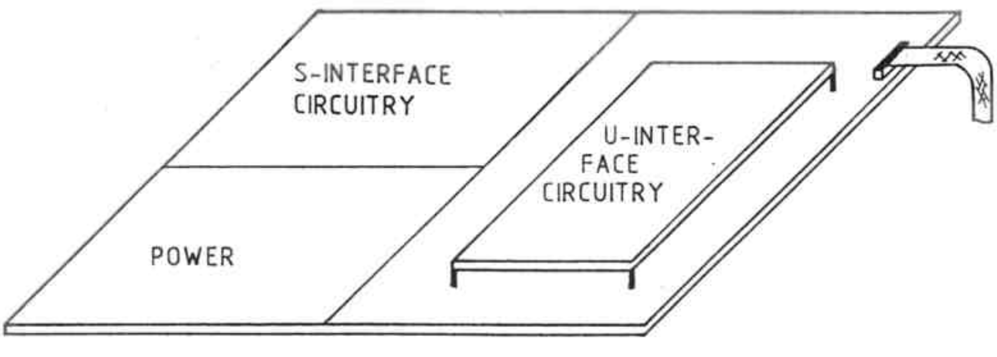


Fig. 6.2.1. Module realization of the NT1.

4B3T UICs are produced by 6 manufacturers (see appendix-E), two of these produce only for internal use. The price of a 4B3T UIC today varies from USD 60 to USD 240. An estimation given by one of the manufacturers, Mietec, indicates that the price in 1994 will be approximately USD 10.

2B1Q UICs are available in 1990 from 5 manufacturers (see appendix-E). 1990 prices vary from USD 35 to USD 100 while it is indicated that UICs will be available in 1995 for USD 20 - 27.

The S-interface circuit.

The SIC is produced by 7 different manufacturers, all according to the CCITT recommendation I.430 [1] (see also appendix-E). No specific problems are assumed to arise, the price and the system interface (discussed below) will be decisive for choosing the SIC. The SIC is available from USD 9 to 15 USD, a 50 % price reduction is expected within the next 5 years.

System interfaces.

As mentioned in chapter 5, there is an additional (non-CCITT specified) interface in the basic rate access, the system interface. This interface is only a concern for the manufacturers of ISDN-equipment since it is an internal interface. Three system interfaces have been introduced, i.e. IOM by Siemens, ST-bus by Mitel and Interchip Digital Link (IDL) by Motorola. The IOM-interface seems to have become the industrial standard, and has been adopted by 7 manufacturers. (See appendix-E)

Power.

The power specifications given in section 5.9. [19] show that 800 mW may be consumed by the NT1 over the U-interface in emergency operation. 410 mW must be supplied to TE over S-interface which leaves 390 mW for the NT1. The transmission section is specified to consume 310 mW [21], [22], measurements show 295 mW.

This implies that the efficiency of the power must be higher than 88 %. The isolation between U- and S-interface must exceed 1500 volt [19].

The emergency power supply may be realized as a switch-mode power supply. There is only one controller for switch-mode supplies available today which satisfies the power requirements, i.e. from Siliconix (Si 9111). Similar controllers will be available from Exar and Harris within the near future. A discrete controller has been realized by Telettra which shows better performance.

7. 2B1Q HARDWARE SIMULATOR

In chapter 3 a theoretical comparison between the three different line codes has been given with respect to ISDN noise injections. To be able to evaluate the result, it was desirable to measure the effect that ISDN noise would have on the existing subscriber loop systems. Since no 2B1Q transceiver is available yet, a hardware simulator had to be built. Power spectrum limits specified by ANSI [17] were used to design the simulator. This chapter will describe the simulator and chapter 8 will describe the rest of the measuring system and give the results. Circuit diagrams can be found in appendix-D.

The simulator can be divided into four as shown in fig. 7.1. A pseudo random bit sequence (PRBS) is generated at 160 kHz, and converted from serial (160 kbit/s) to two bits in parallel (80 kbit/s) to fit into the 2B1Q coder. The output pulse of the 2B1Q coder must be filtered to meet the power spectrum specifications for the 2B1Q transceiver [17].

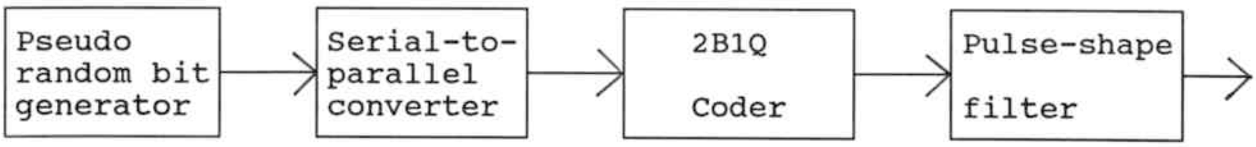


Fig. 7.1: Block diagram 2B1Q hardware simulator.

The pseudo random bit generator is a feedbacked shift register as shown in fig. 7.2. A shift register of length m bits or greater is clocked at a fixed rate of 160 kHz. An exclusive-OR (XOR) gate generates the serial input signal from the combination of the n-th bit and the m-th bit of the shift register. The circuit goes through a set of states, defined by the set of bits in the shift register. The sequence repeats itself after k_s clock periods.

The maximum number of states is equal to 2^m-1 (the state with all zeros is forbidden since the XOR will only regenerate zeros in that state). The criterion for maximum cycle length, $k_s = 2^m - 1$, is that the polynomial x^m (+) x^n (+) 1 must be irreducible and a prime of 1 (+) x^{k_s}.

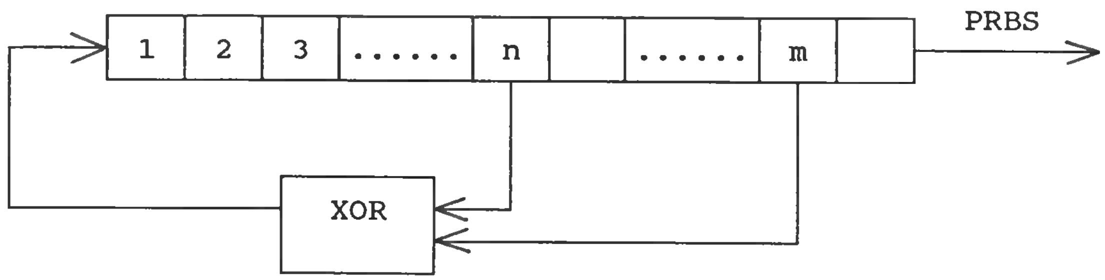


Fig. 7.2: Pseudo-random bit sequence generator.

31 was chosen for m and 28 for n, which implies a cycle time of approximately three hours and three quarters at 160 kHz.

The serial-to-parallel converter was realized by clocking the two last bits of the PRBS-shift register into a buffer at 80 kHz. The result of this operation is that the bit string is divided into sections of two bits, where a bit is only used once.

The 2B1Q coder was realized using a DAC where the digital inputs were encoded. The first bit, in the bit section, is used as a sign bit, while the second determines the magnitude of the four-level output. (See appendix-C and section 3.1.3.)

The pulse-shape filter is a lowpass filter with -3 dB point at 80 kHz and a falling slope of 40 dB/decade. (See fig. 7.3.)

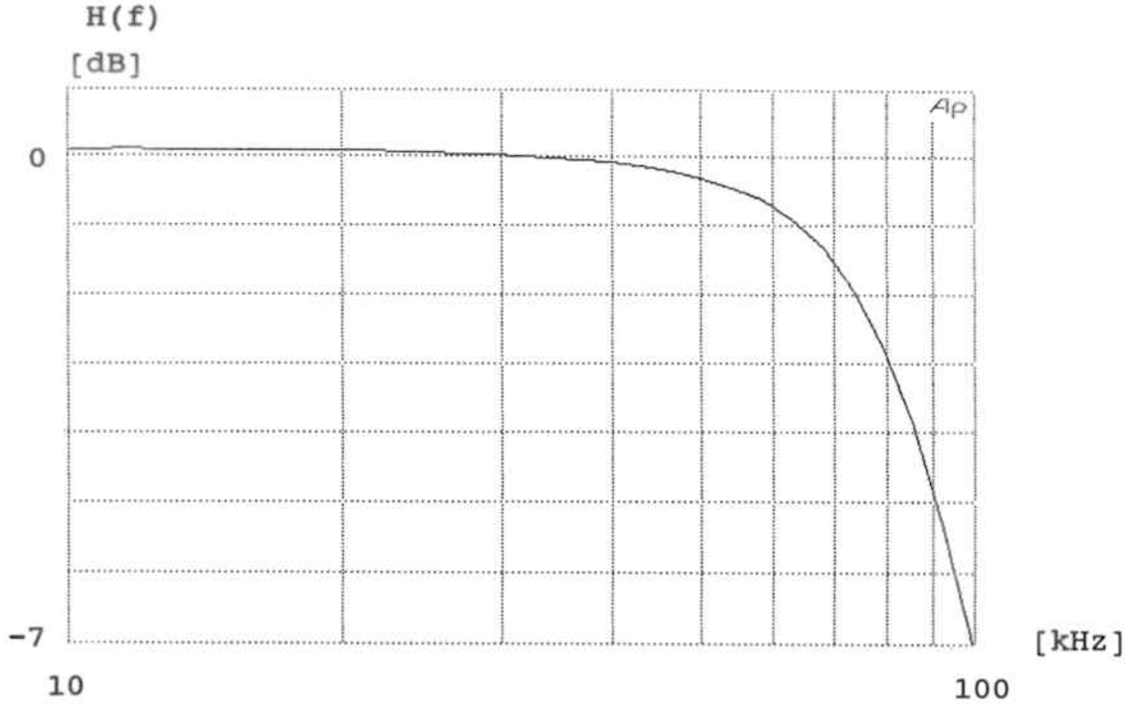


Fig. 7.3: Transfer function of the pulse-shape filter.

8. MEASUREMENTS OF ISDN NOISE INJECTION

In section 3.2.3. a theoretical discussion of ISDN noise injection (NEXT) to the 1+1 system has been given. This chapter will describe the system used to measure the minimum required NEXT-attenuation, and give the obtained results. Section 8.1. will explain the measuring system, while the results can be found in section 8.2. The chapter will end with a short discussion of the results.

8.1. System description.

The measurement system can be divided into three parts, the generators, the NEXT-path, and the subscriber loop. (See fig. 8.1.1.) The purpose of the measuring is to find the minimum required NEXT-attenuation which gives the maximum allowable noise. According to the Norwegian specifications for subscriber equipment, the maximum noise power in the demodulated channel must be less than -55 dBmp (psophometric). The generator will inject noise through the NEXT-path into the subscriber-loop. The NEXT-path-attenuation can be set to the value corresponding to -55 dBmp demodulated noise, which gives the minimum required NEXT-attenuation.

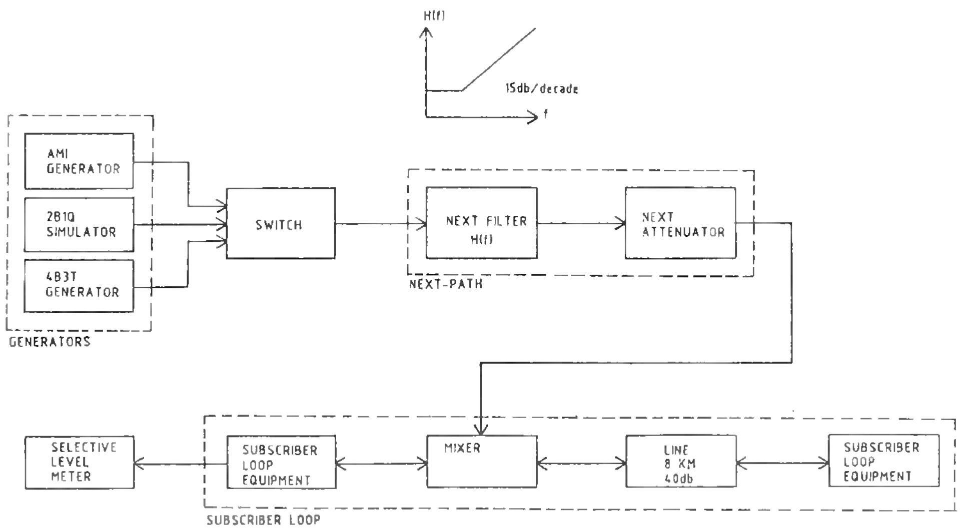


Fig. 8.1.1: System overview, ISDN noise injector.

There are three possible generators, generating an AMI, a 4B3T (MMS43), or a 2B1Q power spectrum, respectively. The 2B1Q generator has been described in the previous chapter. AMI was generated by a Hewlett Packard pattern generator (HP 3780A) followed by an attenuator, while an ISDN Network Termination (NT1) from Telettra was used as a 4B3T generator. The output level of the 2B1Q generator is 11 dBm and the 4B3T output level is 9 dBm, as specified in the preliminary specifications for the ISDN transceivers. 10 dBm was used as AMI output level. Fig. 8.1.2., fig. 8.1.3. and fig. 8.1.4. show the power spectra of the three generators.

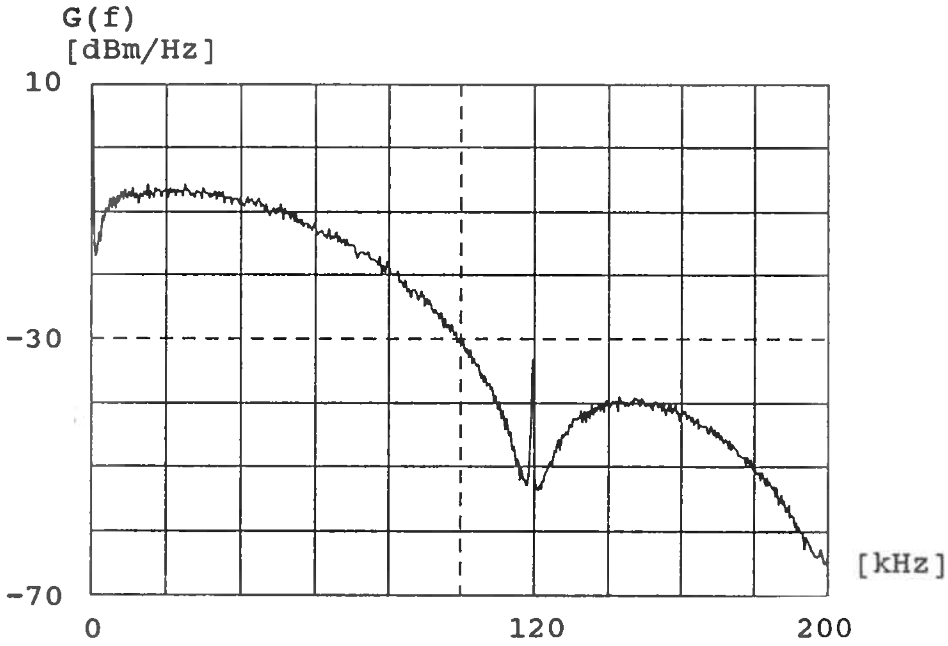


Fig. 8.1.2: Measured power density spectrum of the 4B3T generator.

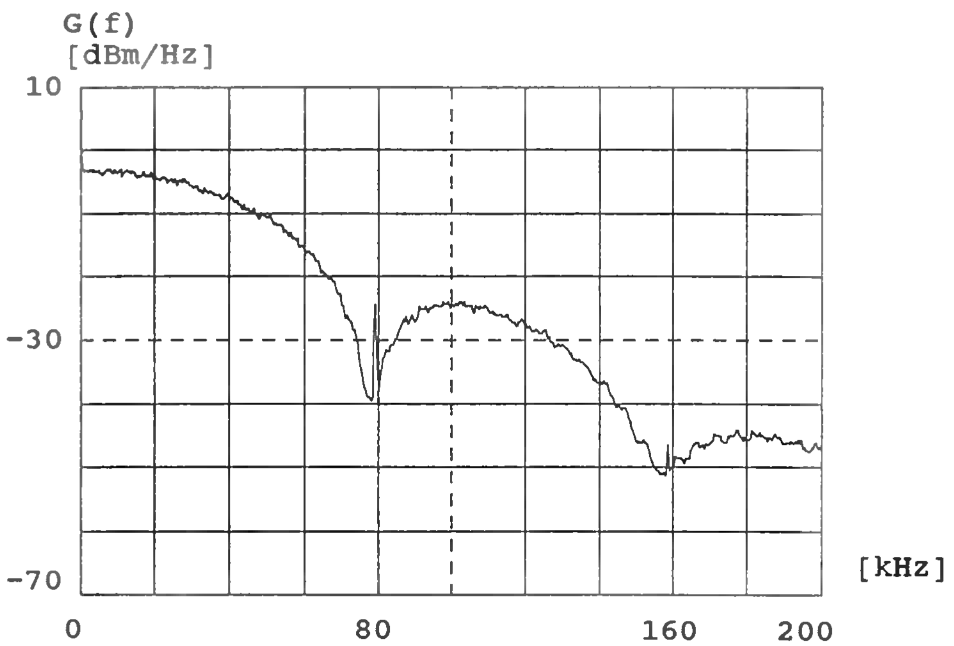


Fig. 8.1.3: Measured power density spectrum of the 2B1Q-simulator.

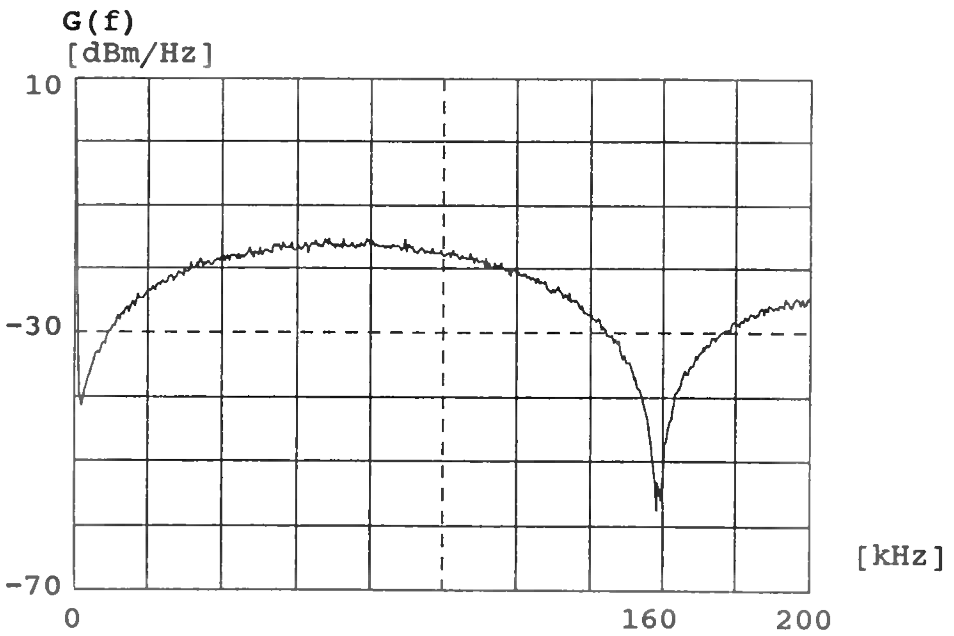


Fig. 8.1.4: Measured power density spectrum of the AMI generator.

The generators are followed by the NEXT-path, which consists of a filter and an attenuator. The transfer function of NEXT increases 15 dB per decade [7,8]. The NEXT-filter was designed to simulate this frequency characteristic, see fig. 8.1.5. The frequency range is 10 to 100 kHz, which covers the frequency-band used by the 1+1 system. The attenuation of the filter at 100 kHz is 0 dB.

The attenuator is used to set the proper simulated NEXT-path level. The attenuator consists of a variable attenuator with a range from 0 to 45 dB, in steps of 1.5 dB, and a constant 40 dB-attenuator. The 40 dB-attenuator is used when attenuation between 45 and 85 dB is desirable. Fig. 8.1.6. shows the first 10 steps of the NEXT-attenuator in combination with the filter.

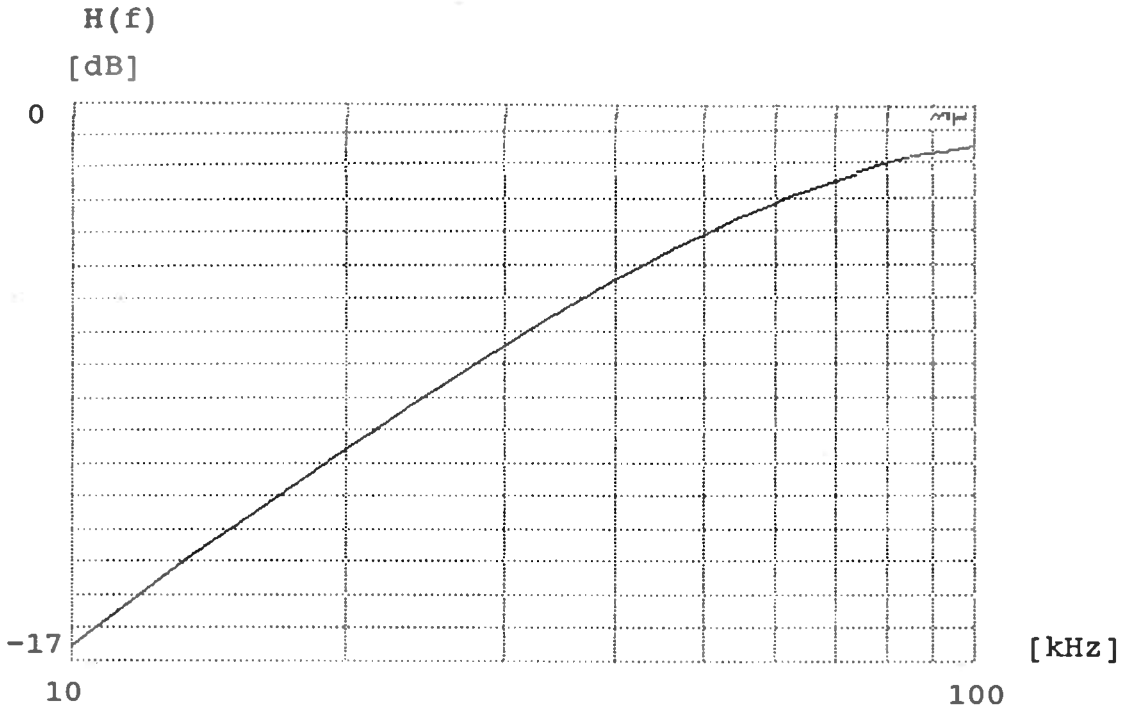


Fig. 8.1.5: NEXT-filter characteristic.

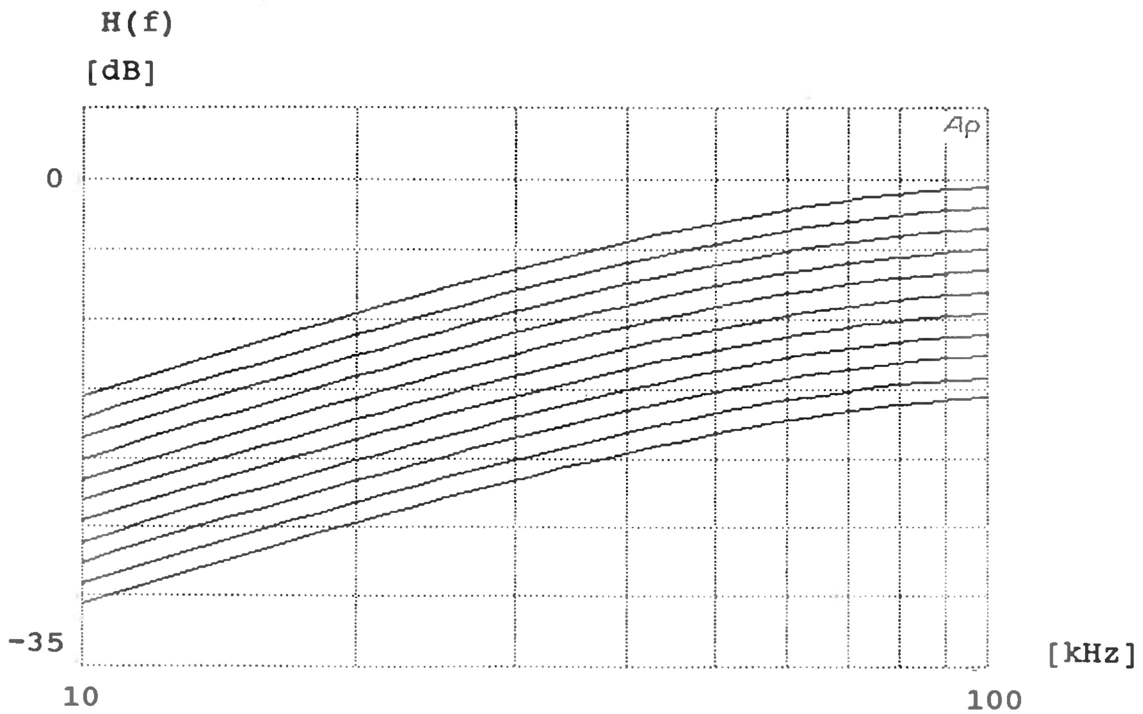


Fig. 8.1.6: Transfer characteristic of the NEXT-path, 1.5 dB/step.

The noise was injected into the subscriber-loop over a transformer and two 1500 ohm resistances (see appendix-C). This was done to eliminate any loss from the subscriber loop. (The input impedance of the 1+1 system is 150 ohm.) An amplifier prior to the transformer was built to compensate for the loss over the 1500 ohm resistances. The level meter used to measure demodulated noise was a Hewlett-Packard transmission test set (HP 3552A).

8.2. Measured minimum required near-end-crosstalk attenuation.

Minimum required NEXT-attenuation was measured, as described above, using maximum specified cable-attenuation in the subscriber loop. The 1+1 system transmission range is specified as 40 dB attenuation at 79 kHz (from exchange to subscriber) and 32 dB at 28 kHz (reverse direction). A real 0.6 mm cable was used. (40 dB attenuation at 79 kHz was equal to 8.3 km while 32 dB attenuation at 28 kHz was equal to 8.1 km.) Noise was injected both at the subscriber-end and the exchange-end of the 1+1 system. The results of the measurements are given in table 8.2.1. (minimum required NEXT-attenuation at 100 kHz).

Table 8.2.1: Minimum required NEXT-attenuation at 100 kHz for three different line codes, given that the demodulated noise must be less than - 55 dBmp.

Min.req. NEXT-att. 2B1Q		Min.req. NEXT-att. 4B3T		Min.req. NEXT-att. AMI	
Subscr. end (79 kHz) [dB]	Exchange end (28 kHz) [dB]	Subscr. end (79 kHz) [dB]	Exchange end (28 kHz) [dB]	Subscr. end (79 kHz) [dB]	Exchange end (28 kHz) [dB]
43.5	59.0	61.5	54.0	69.5	49.5

The performance of the line codes can be measured in minimum required NEXT-attenuation, but cable-failure-percentages (cfp) can also be used. These are the percentage of cables in the subscriber-loop-plant which fail to meet the minimum required NEXT-attenuation.

Table 8.2.2. gives the results in cfp, for two types of cables used in the Norwegian subscriber-loop-plant. It is the older 4-pair cable and the newer 10-pair cable. Table 8.2.2. is based on measurements done by NTRL [7], 3712 combinations of 10-pairs cable and only 195 combinations of 4-pairs cable were measured. The cfp values in table 8.2.2. are therefore not necessarily representative for the Norwegian subscriber-loop-plant, but are included to give an indication of the cfp (in lack of anything better).

Table 8.2.2: Cable-failure-percentages given for three different line codes, and demodulated noise less than -55 dBmp.

Cable-type # of pairs	Cable-failure-percentages 2B1Q Subscr. end [%]	Cable-failure-percentages 2B1Q Exchange end [%]	Cable-failure-percentages 4B3T Subscr. end [%]	Cable-failure-percentages 4B3T Exchange end [%]	Cable-failure-percentages AMI Subscr. end [%]	Cable-failure-percentages AMI Exchange end [%]
10	0.0	1.5	3.5	0.0	39.5	0.0
4	5.5	32.5	45.5	22.0	69.5	13.5

8.3 Comparison and discussion

The cfp-value given above is based on data concerning a pair-to-pair NEXT-attenuation, however the noise contribution usually comes from more pairs. The worst case in a 10-pair cable, is when 9 pairs are used for ISDN basic rate access while the last one is used for the 1+1 system. The total noise contribution from nine pairs is 12.3 dB higher than from a single pair [9], assuming the traffic to be 0.1 Erlang. The result is that the minimum required NEXT-attenuation must be increased by the same amount. The contribution from pairs outside the 10-pair cable is not taken into consideration since this value is negligible.

The NEXT-noise at the subscriber-end depends on the distance from the connection box, where the 10/4-pairs cables are splitted into single pairs, to the subscriber. The cable-attenuation over this distance will reduce the noise. However, in many cases this distance is not large enough to prevent NEXT. NEXT at the subscriber-end must therefore be considered as a possible problem.

All in all, looking at the results in table 8.2.1. and 8.2.2. one can conclude that problems will arise if ISDN basic rate access is introduced without any precautions. Care must be taken when connecting an ISDN-subscriber, so that the ISDN-equipment is not connected adjacent to a 1+1 system pair. Another solution is to reduce the output-level of the transceiver by adding an artificial 1.5 km line at each end of the 2-wire loop. The maximum range will then be reduced to 5 km [22]. This is however enough to cover approximately 98 % of the Norwegian subscriber loops, while the last 2 % is cover without the artificial line.

If table 8.2.1. is compared to the results in section 3.2.3., one will see the same internal ranking for the subscriber-end as well as for the exchange-end. The differences are higher required attenuation-level in general and the higher value measured at the subscriber-end. The result of these changes is that 2B1Q is the best performing line code, in contrast to the results found in section 3.2.3. where 4B3T was found to perform best.

The calculations in chapter 3 apply a rectangular pulse-shape while the actual pulses are more cosine-shaped. The use of rectangular pulses implies that more power is found in the higher frequency regions and less in the lower frequency regions. In addition, the effect of non-ideal filters and hybrid transformers is not included in the calculation model in chapter 3.

9. CONCLUSIONS

1. The analysis of the line codes in chapter 3 showed that all three can be used for the basic rate access. However 2B1Q showed best performance, it has the longest range (chapter 3) and least influence on the critical 1+1 system (chapter 8).

2. Chapter 8 showed that care must be taken when introducing ISDN in the Norwegian subscriber loop. The measurements indicated that the basic rate access transceivers can interfere with 1+1 systems if these systems are used in the same cable. The other subscriber systems are considered compatible with ISDN. The theoretical calculations in chapter 3 are in disagreement with the measurements in chapter 8, this is due to the limited model and the rectangular pulses assumed in chapter 3. Non-ideal effects in the filters and hybrid transformers of the 1+1 system are not taken into consideration.

3. The study of the two echo cancellation solutions showed that the digital solution should be preferred for the following reasons:
   - No digital-to-analog converter or complex analog filters are required.
   - Realization of the analog-to-digital converter using pulse-density-modulation technique is considered suitable for implementation in digital MOS-technique.

4. Two interfaces in the basic rate access have not been specified by the CCITT, both concerning the network termination 1, i.e the U-interface and the internal system interface. For production of the Network Termination 1 IOM is proposed as system interface since it is assumed to become the industry standard. For the U-interface in Network termination 1, a daughter-board comprising the U-interface circuitry is proposed since the U-interface can then be changed without having to replace the whole Network termination 1.

REFERENCES

[1] -
CCITT Recommendation I.430.
"Basic User-Network Interface-Layer 1 specification."
CCITT Red Book, Vol. III, Fascicle III.5, 1984.

[2] F.F.E. Owen,
PCM and digital transmission systems, first edition
New York: McGraw-Hill, New York, 1982.

[3] A.B. Carlson,
Communication Systems, third edition
Tokyo: McGraw-Hill, Singapore, 1986.

[4] G. Morgenstern and H.W. Wellhausen,
"Drei- und mehrstufige Leitungscodes für die Digitalsignal-übertragung.",
Fernmelde-Ing. (Germany), Vol. 41, No. 3,
pp. 1-47, March 1987.

[5] J.B. Buchner,
"Ternary line codes"
Philips Telecomm. Review, Vol. 34, No. 2, p. 72-86,
June 1976.

[6] D. Sallaerts et. al.,
"A single chip U-interface transceiver for ISDN."
IEEE J. of Solid-State Circuits, Vol. SC-22, No. 6,
pp. 1011, Dec. 1987.

[7] P. Klepsland,
"Bakgrunn for planleggings-regler i abonnentnettet. Transmisjonsmålinger på kabler.
TF-report No. 41/87, Teledirektoratets forskningsavdeling, Norwegian telecomm. Administration, July 1987.

[8] H. Ekkelenkamp,
Transmissie-aspecten van digitale communicatiesystemen, first edition.
PTT Dr. Neher Laboratorium, Leidschendam, The Netherlands.
Published by Secretarie B-CD, Dutch Telecomm. Administration, 1984.

[9] N.Holte,
"Line codes for digital subscriber lines."
ELAB report STF44 A833138, NTH, May 1983.

[10] L. Bellato et. al.,
"144 kbit/s digital transmission on the subscriber loop for ISDN user access."
Telettra Review, No. 36, pp. 51-86, Dec. 1984.

[11] Teledirektoratet, TLB,
Technical specification for 2-wire subscriber carrier frequency equipment, Sp. nr. 21, Utg. 7, Aug. 1983.

[12] P. Pedersen, B. Justnes,
"Field trails with digital transmission in the subscriber loop plant", (in norwegian),
Telektronikk, no. 4, p. 331-338, 1981

[13] N.A.M. Verhoeckx et. al.,
"Digital echo cancellation for baseband data transmission."
IEEE Trans. on acoustics, speech, and sign. processing,
Vol. ASSP-27, No. 6, pp. 768-781, December 1979.

[14] K. Széchényi et. al.,
"Integrated full-duplex U-interface circuit for ISDN subscriber loops."
IEEE Selected areas in communications, Vol. SAC-4, No. 8, Nov. 1986.

[15] B. P. Agrawal, K. Shenoi
"Design methodology for SDM."
IEEE Trans. on comm., Vol. COM-31, No. 3, pp. 360, March 1983.

[16] J.C. Candy,
"Decimation for sigma delta modulation."
IEEE Trans. on comm., Vol. COM-34, No. 1, pp. 72, Jan. 1986.

[17] American National Standard-ISDN,
Basic-access interface for use on metallic loops for application on the network side of NT layer 1 specifications, ANSI T1.601-1988, to be published in Dec. 1988.

[18] A.P. Clark,
Equalizers for digital modems, first edition
London: Pentech Press, London,1985.

[19] Norwegian Telecommunications Administration,
"Power feeding concept for the basic access."
Specification and technical requirements for the Norwegian ISDN pilot project., first edition, Jan. 1988.

[20] P. Pedersen, B. Justnes,
"Digital transmission field-trails in the subscriber loop."
Telektronikk, No. 4, April 1981 (In Norwegian)

[21] -
"S-bus interface circuit."
Technical manual, Rev. 2.1, Siemens AG. June 1987.

[22] -
"ISDN echocancellation circuit."
Target specification, Rev.2.1, Siemens AG. Dec. 1987.

[23] J.J. van der Kam,
"Een digitaal 'decimerend' filter voor analog/digital-omzetting van 'hifi'-audio."
Philips Techn J., T.42, nr. 6/7, Oct 1984.

[24] J.D. Everard,
"A single-channel PCM codec."
IEEE Trans. on comm., Vol. COM-27, No. 2, Feb. 1979.

APPENDIX-A: DECISION FEEDBACK EQUALIZER.

There are two types of equalizers which can be used in ISDN basic rate access, linear equalizer and decision feedback equalizer. Fig. A.1. shows the linear equalizer arrangement and fig. A.2. the decision feedback equalizer.

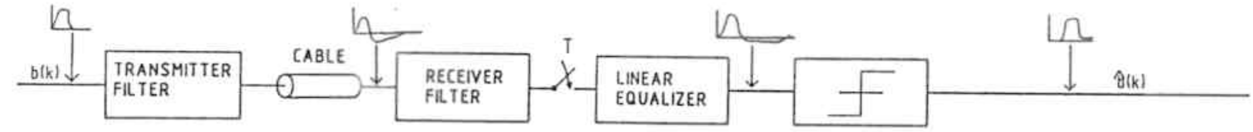


Fig. A.1: Linear equalizer.

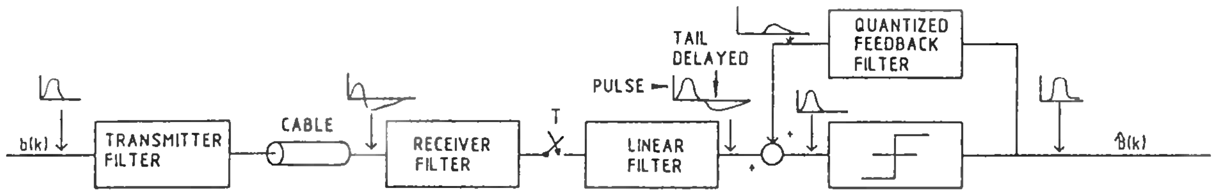


Fig. A.2: Decision feedback equalizer.

The equalizer output is distortionless (perfect equalization) if its transfer function is equal to [3]

$$H_e(f) = \frac{K\, e^{-j2\pi f t_d}}{C(f)}, \qquad \text{(A.1.)}$$

where C(f) is the cable transfer function, K and t_d are constants. This transfer function is however seldom possible to realize, so an approximation must be used.

Linear equalizers can be realized by transversal filters with transfer function given by [3]

$$H_e(f) = e^{-j2\pi fNT} \left(\sum_{m=-N}^{N} c_m\, e^{-j2\pi fmT} \right) \qquad \text{(A.2.)}$$

where c_m, T and 2N + 1 are the adjustable tap gains, the delay of one filter element, and number of taps, respectively. This transfer function can approximate equation (A.1.). The result is that the tail of the incoming pulse is forced to zero at the sampling instants adjacent to its own, and the inter-symbol interference is brought to a minimum.

For practical realizations the equalizer must be made adaptive, since the cable transfer function vary with time. Linear equalizers can not be used with unbalanced line codes, since dc/lf restoration is impossible.

The decision feedback equalizer (see fig. A.2.) consists of a linear filter (pre-equalizer) and a quantized feedback filter. The linear filter partly equalizes the channel. All components of the sampled cable impulse response preceding that of the largest magnitude, are set to zero (see fig. A.2.). This is done without changing the relative values of the remaining components. The quantized feedback filter completes the equalization. DFE is a nonlinear process, capable of restoring the removed dc and lf components. DFE outperforms the linear equalizer and is therefore also used with balanced line codes.

Decision feedback equalization has two inconveniences, error extension and possible instability. Normally error extension is not serious due to the high SNR. Practice shows that instability can be avoided by a proper design. DFE noise-tolerance is 3-5 dB better than that of a linear equalizer [18].

APPENDIX-B: SNR REQUIREMENT FOR BIT ERROR PROBABILITY = 10^{-7}.

To find the relation between SNR and P_{be}, the error probability for each level of the code must be found. This appendix starts with finding the P_{be}/SNR relation for 2B1Q. Fig. B.1. shows the four levels of 2B1Q with white gaussian noise, which is a pessimistic approximation. The levels are taken to be

$$b(k) = +/- \frac{3D}{2} \ , +/- \frac{D}{2} \qquad \text{(A.1.)}$$

assuming equal probability for zeros and ones in the binary signal, which for 2B1Q gives equal probability for all levels. This means that the probability of a symbol-error is equal to

$$P_{se} = \frac{1}{4} (P_{e0} + P_{e1} + P_{e2} + P_{e3}) \qquad \text{(B.2.)}$$

where P_{e0} is the probability of the conditional PDF, $p_Y(y|H_0)$, being less than - D. Since the noise is gaussian [3],

$$P_{e0} = Q \left(\frac{D}{2\sigma}\right) \qquad \text{(B.3.)}$$

where Q (k) = 0.5 erfc ($k/\sqrt{2}$) and σ^2 is the noise power at the receiver, N_R.

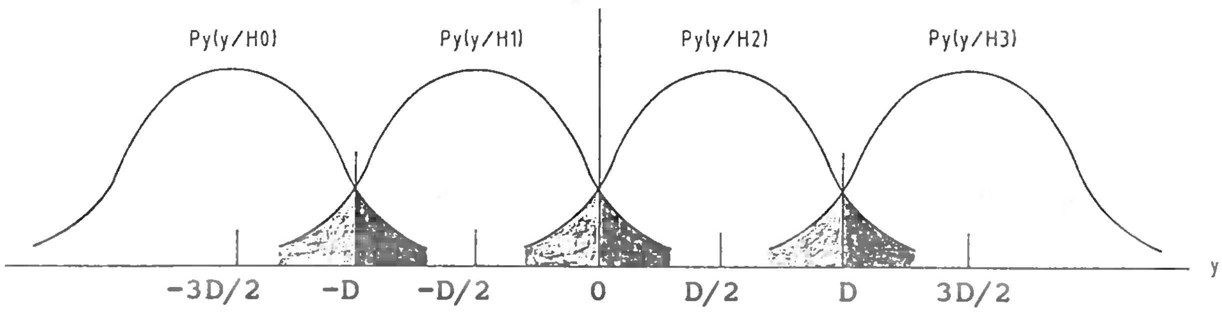


Fig. B.1: Conditional PDFs for 2B1Q signal with noise.

This is also true for P_{e3}, while

$$P_{e1} = P_{e2} = 2\ Q\ (\frac{D}{2\ \sigma}) \qquad (B.4.)$$

because both positive and negative noise samples produce an error. The total symbol error probability is then

$$P_{se} = 1.5\ Q\ (\frac{D}{2\ \sigma}) \qquad (B.5.)$$

and if only a transition to an adjacent level is considered, then

$$P_{be} = \frac{P_{se}}{2} = 0.75\ Q\ (\frac{D}{2\ \sigma}) \qquad (B.6.)$$

since only one bit is different between two adjacent levels (see table 2.3.1.). According to [3]

$$S_R = <(b(k))^2>\ r\ \tau_{eq} \qquad (B.7.)$$

where τ_{eq} is the integral of the square pulse, and r is the transmission rate. The average power is equal to

$$<(b(k))^2> = \frac{5}{4}\ D^2 \qquad (B.8.)$$

The signal power at the receiver is therefore equal to

$$S_R = \frac{5}{4}\ D^2\ r\ \tau_{eq} \qquad (B.9.)$$

so

$$D^2 = \frac{5}{4}\ \frac{S_R}{r\ \tau_{eq}} \qquad (B.10.)$$

Noise power N_R is equal to

$$N_R = \sigma^2 = H\ B_T \qquad (B.11.)$$

where H is white noise density and B_T is the transmission bandwidth.

The noise power N_R has a minimum value, which can be expressed in noise density and τ_{eq}

$$N_R = \sigma^2 \geq \frac{H}{2\ \tau_{eq}} \qquad (B.12.)$$

where N_R is equal to the minimum value if matched filtering is used [3]. The minimum transmission bandwidth is limited by the transmission rate according to the following relation

$$B_T \geq \frac{r}{2} \qquad (B.13.)$$

At last we are able to find the relation between $D/2\sigma$ and the SNR_R, using (B.10.) and (B.12.)

$$\frac{D^2}{4\ \sigma^2} \leq \frac{2\ S_R}{r\ H} \qquad (B.14.)$$

By substituting (B.13.) into (B.14.) the following relation can be found

$$\frac{D^2}{4\ \sigma^2} \leq \frac{S_R}{5\ N_R} \qquad (B.15.)$$

So the minimum required value of SNR_R can be found by substituting (B.15.) into (B.6.)

$$P_{be2B1Q} \geq \frac{3}{4}\ Q\ [\frac{\sqrt{SNR_R}}{\sqrt{5}}] \qquad (B.16.)$$

Similar calculations can be done for AMI and MMS43 (see fig. B.2.), the sub-results are:

$$\text{AMI}: \quad P_{se} = 1.5\ Q\ (A/2\sigma) \tag{B.17.}$$

$$\text{MMS43}: \quad P_{se} = (43/32)\ Q\ (B/2\sigma) \tag{B.18.}$$

MMS43: $p_r(+B) = p_r(-B) = 63/192$, $p_r(0) = 66/192$
AMI : $p_r(+A) = p_r(-A) = 0.25$, $p_r(0) = 0.5$

$$\text{AMI}: \quad P_{be} = P_{se} \tag{B.19.}$$

$$\text{MMS43}: \quad P_{be} = 1.8\ P_{se} \tag{B.20.}$$

(A MMS43 symbol-error produces in average 1.8375 bit-errors, while an AMI symbol-error produces one bit-error.)

$$\text{AMI}: \quad <(b(k))^2> = (1/2)\ A^2 \tag{B.21.}$$

$$\text{MMS43}: \quad <(b(k))^2> = (2/3)\ B^2 \tag{B.22.}$$

At last the relation between P_{be} and SNR can be found

$$P_{beAMI} \geq 1.5\ Q\ [\sqrt{(SNR_R/2)}] \tag{B.23.}$$

$$P_{beMMS43} \geq 2.5\ Q\ [\sqrt{(3SNR_R/8)}] \tag{B.24.}$$

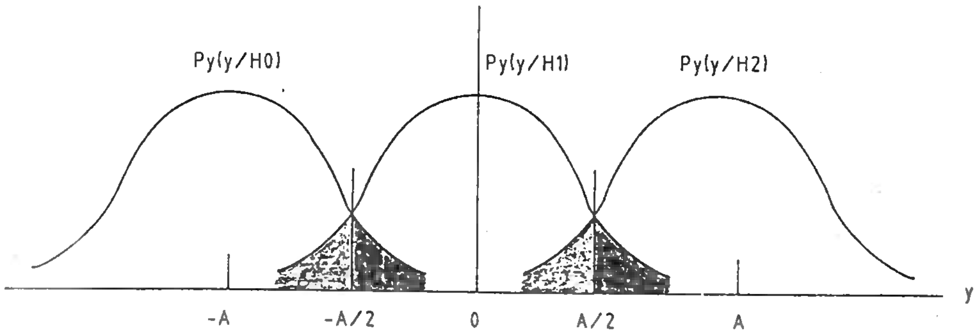


Fig. B.2: Conditional PDFs for AMI/MMS43 signal with noise.

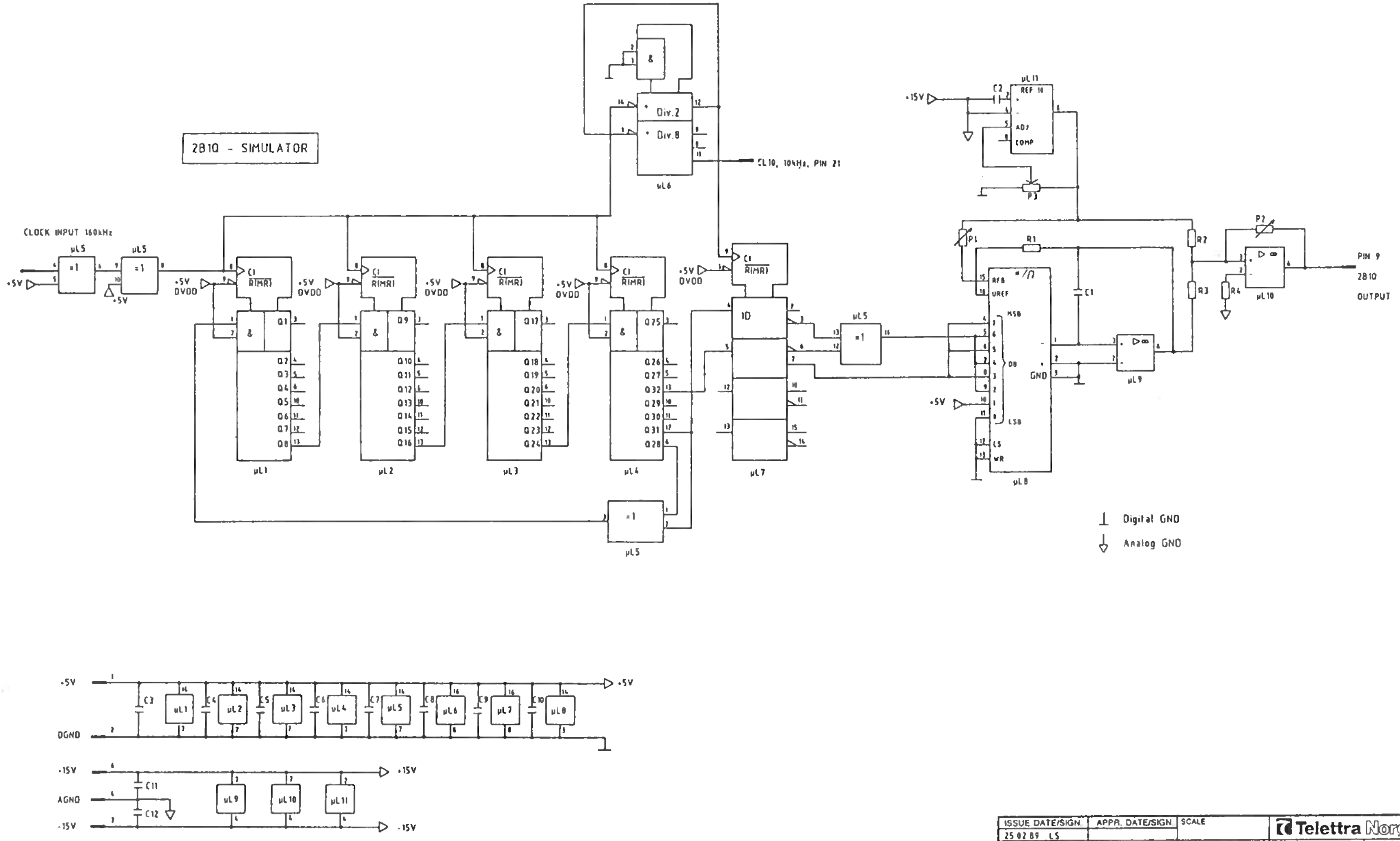
APPENDIX-C: CIRCUITS DIAGRAMS

This appendix contains the circuit diagrams of the 2B1Q simulator, the NEXT-simulator and the output-stage. The component values used are listed below.

μL 01 : HCT 164	C 1 = 15 pF	C21 = 33 nF
μL 02 : HCT 164	C 2 = 0.1 μF	C22 = 100 pF
μL 03 : HCT 164	C 3 = 10 nF	C23 = 47 pF
μL 04 : HCT 164	C 4 = 10 nF	C24 = 1 μF
μL 05 : HCT 86	C 5 = 10 nF	C25 = 1 μF
μL 06 : HCT 93	C 6 = 10 nF	C26 = 470 pF
μL 07 : HCT 175	C 7 = 10 nF	C27 = 402 pF
μL 08 : AD 7524	C 8 = 10 nF	C28 = 1 μF
μL 09 : AD 544	C 9 = 10 nF	C29 = 1 nF
μL 10 : AD 544	C10 = 10 nF	C30 = 181 nF
μL 11 : AD 587	C11 = 10 nF	C31 = 1 μF
μL 12 : HCT 283	C12 = 10 nF	C32 = 10 nF
μL 13 : HCT 283	C13 = 10 nF	C33 = 10 nF
μL 14 : LS 4081	C14 = 10 nF	C34 = 1 μF
μL 15 : AD 7111	C15 = 10 nF	C35 = 10 pF
μL 16 : AD 544	C16 = 10 nF	C36 = 22 μF
μL 17 : TL 084	C17 = 10 nF	C37 = 1 μF
μL 18 : TL 084	C18 = 10 nF	C38 = 1 μF
μL 19 : TL 084	C19 = 33 pF	C49 = 22 μF
μL 20 : LM 318	C20 = 680 pF	C40 = 10 nF
		C41 = 10 nF
		C42 = 1 μF
		C43 = 1 μF

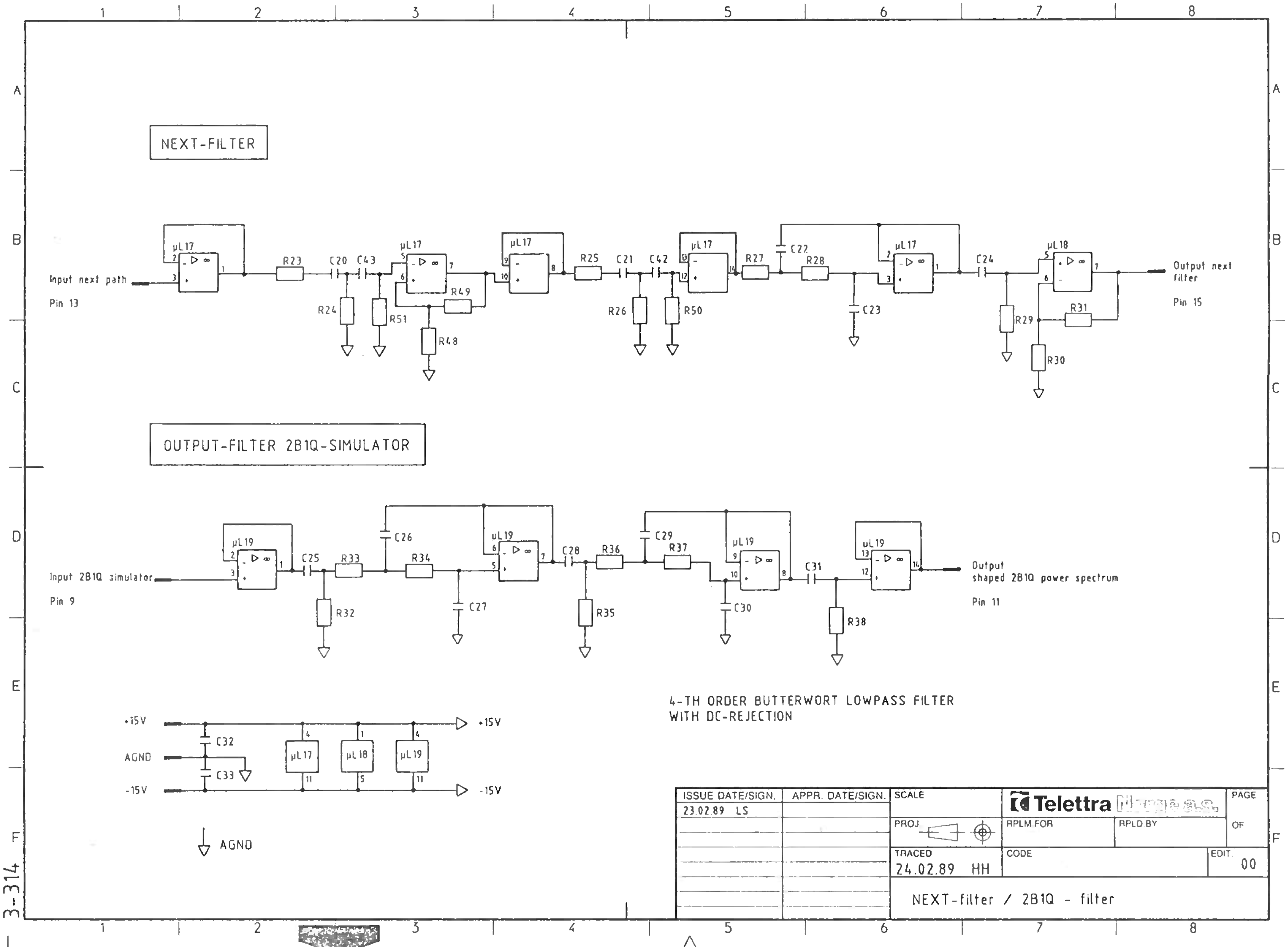
TR1 : BDY 58 IX	P1 = 2 kΩ	RD1 : LED
TR2 : BDY 58 IX	P2 = 20 kΩ	RD2 : LED
TR3 : 2N4401	P3 = 10 kΩ	RD3 : 1N4148
TR4 : 2N4403	P4 = 0.76 Ω	RD4 : 1N4148

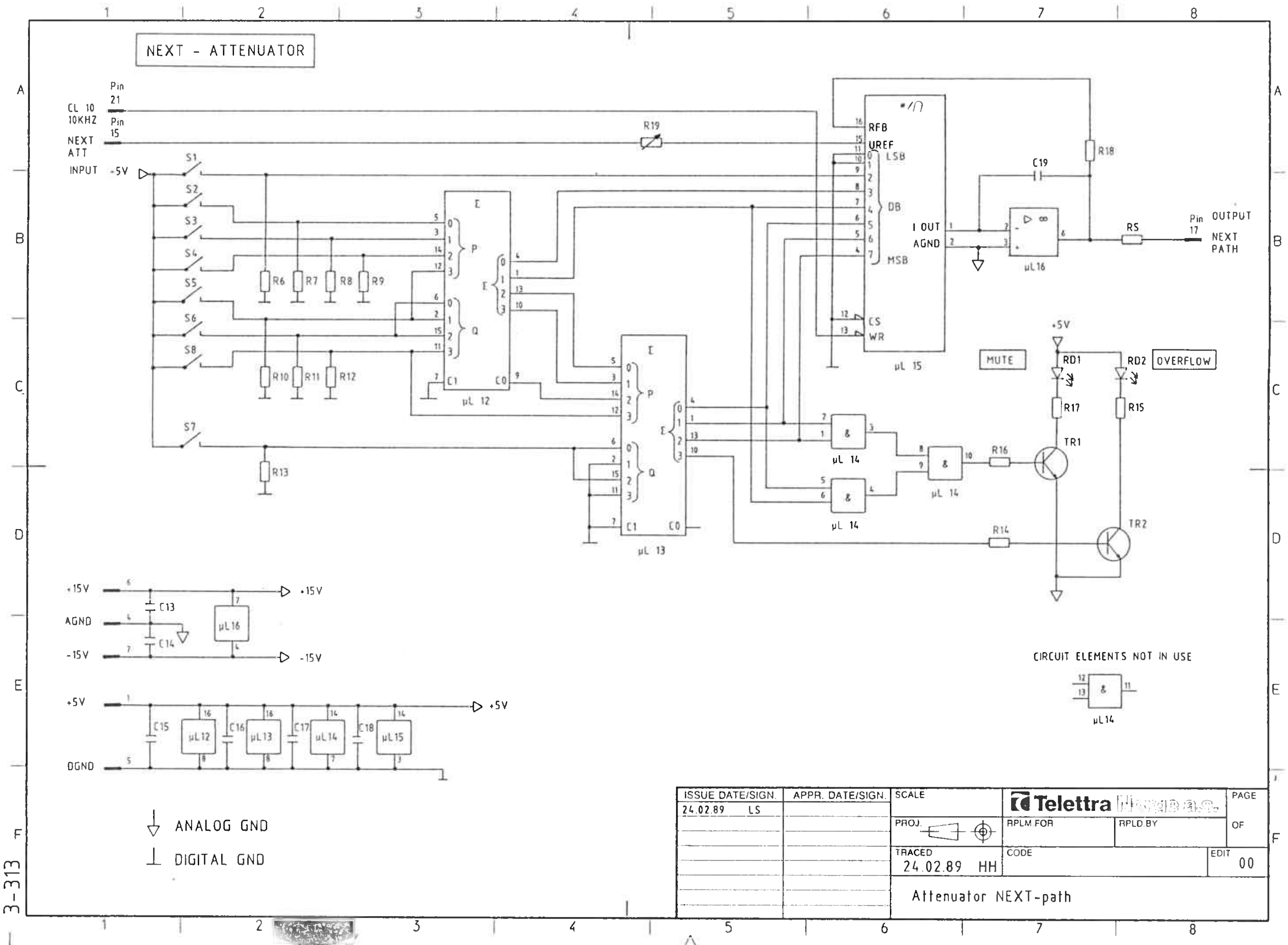
R 1 = 1 kΩ	R18 = 180 kΩ	R35 = 1 MΩ
R 2 = 20 kΩ	R19 = 20 kΩ	R36 = 4.1 kΩ
R 3 = 10 kΩ	R20 = 7.425 kΩ	R37 = 4.1 kΩ
R 4 = 5 kΩ	R21 = 75 Ω	R38 = 1 MΩ
R 5 = 75 Ω	R22 = 75 Ω	R39 = 300 kΩ
R 6 = 2 kΩ	R23 = 2.3 kΩ	R40 = 1.05 kΩ
R 7 = 2 kΩ	R24 = 733 Ω	R41 = 2.74 kΩ
R 8 = 2 kΩ	R25 = 854 Ω	R42 = 3.3 kΩ
R 9 = 2 kΩ	R26 = 927 Ω	R43 = 3.3 kΩ
R10 = 2 kΩ	R27 = 4.7 kΩ	R44 = 15 Ω
R11 = 2 kΩ	R28 = 4.7 kΩ	R45 = 15 Ω
R12 = 2 kΩ	R29 = 1 MΩ	R46 = 1.5 kΩ
R13 = 2 kΩ	R30 = 1 kΩ	R47 = 1.5 kΩ
R14 = 3.5 kΩ	R31 = 4.6 kΩ	R48 = 1 kΩ
R15 = 402 kΩ	R32 = 1 MΩ	R49 = 6.9 kΩ
R16 = 3.5 kΩ	R33 = 4.1 kΩ	R50 = 1 MΩ
R17 = 402 kΩ	R34 = 4.1 kΩ	R51 = 1 MΩ



ISSUE DATE/SIGN	APPR. DATE/SIGN	SCALE	Telettra Norge a.s.		PAGE
25.02.89 LS		PROJ.	RPLM.FOR	RPLD BY	OF
		TRACED 25.02.89 HH	CODE		EDIT. 00
		2B1Q - Simulator			

2-082


NEXT-FILTER
Input next path
Pin 13
μL17
R23
C20
C43
R24
R51
R48
R49
R25
C21
C42
R26
R50
R27
C22
R28
C23
C24
μL18
R29
R30
R31
Output next filter
Pin 15
OUTPUT-FILTER 2B1Q-SIMULATOR
Input 2B1Q simulator
Pin 9
μL19
C25
R32
R33
C26
R34
C27
C28
R35
R36
C29
R37
C30
C31
R38
Output
shaped 2B1Q power spectrum
Pin 11
4-TH ORDER BUTTERWORT LOWPASS FILTER
WITH DC-REJECTION
+15V
AGND
-15V
C32
C33
AGND
ISSUE DATE/SIGN.
23.02.89 LS
APPR. DATE/SIGN.
SCALE
PROJ
TRACED
24.02.89 HH
Telettra
RPLM FOR
RPLD BY
CODE
PAGE
OF
EDIT
00
NEXT-filter / 2B1Q - filter
3-314



NEXT - ATTENUATOR
CL 10 10KHZ Pin 21
NEXT ATT INPUT Pin 15
-5V
S1 S2 S3 S4 S5 S6 S7 S8
R6 R7 R8 R9 R10 R11 R12 R13
R19
μL 12
μL 13
μL 14
μL 15
μL 16
RFB
UREF
LSB
DB
MSB
I OUT
AGND
CS
WR
C19
R18
RS
Pin 17
OUTPUT NEXT PATH
+5V
RD1
RD2
MUTE
OVERFLOW
R17
R15
R16
R14
TR1
TR2
+15V
AGND
-15V
C13
C14
+5V
DGND
C15 C16 C17 C18
CIRCUIT ELEMENTS NOT IN USE
ANALOG GND
DIGITAL GND
ISSUE DATE/SIGN.
24.02.89 LS
APPR. DATE/SIGN.
SCALE
PROJ.
TRACED
24.02.89 HH
Telettra
RPLM FOR
RPLD BY
CODE
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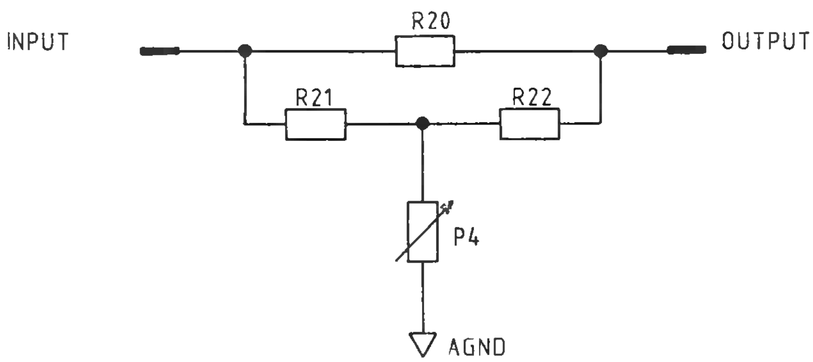
OF
EDIT
00
Attenuator NEXT-path
3-313
A.S TORRKOPI

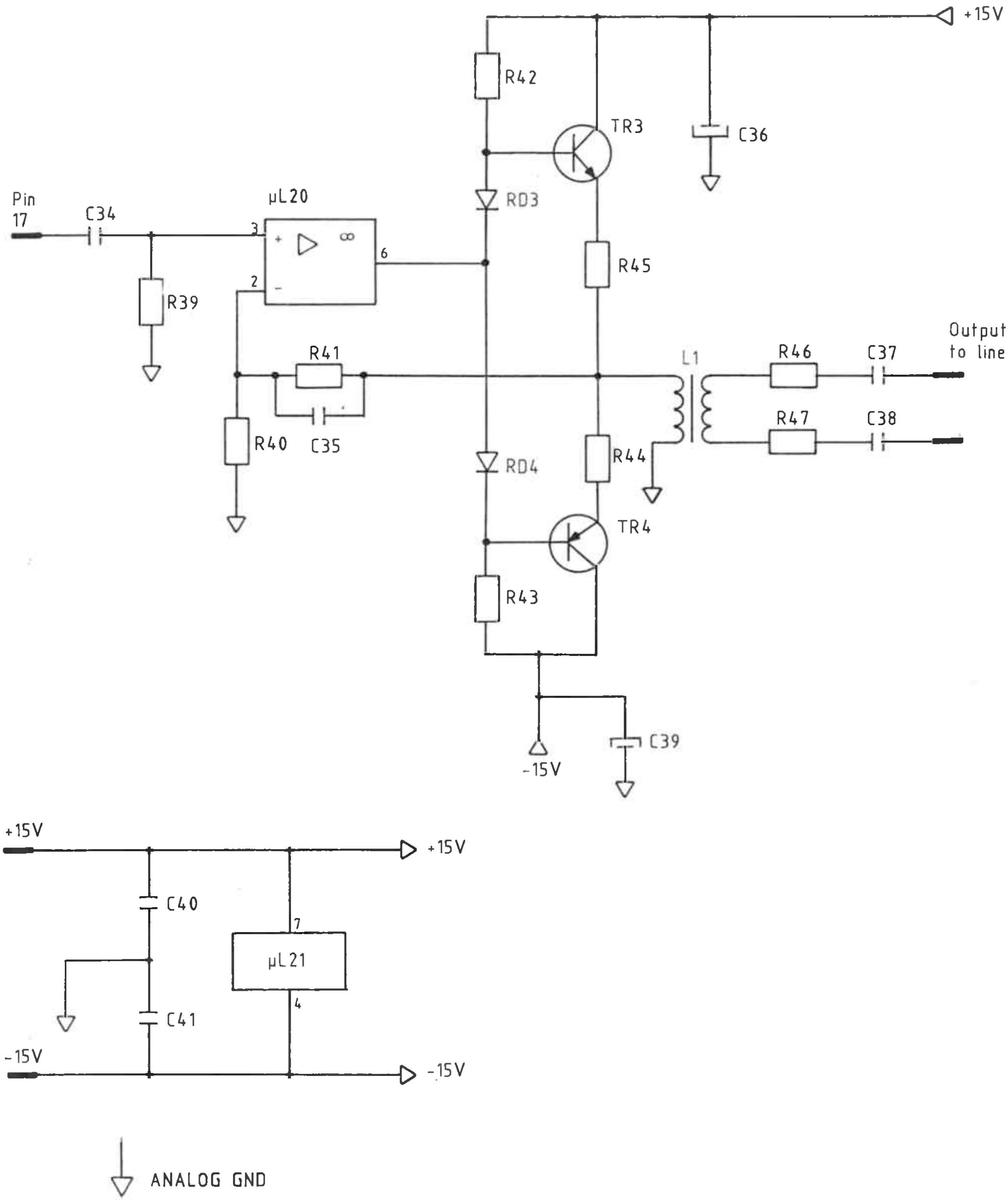
40dB- ATTENUATOR -75Ω



ISSUE DATE/SIGN.	APPR. DATE/SIGN.	SCALE	Telettra Norge a.s.		PAGE
23.02.89 LS		PROJ.	RPLM.FOR	RPLD.BY	OF
		TRACED 23.02.89 HH	CODE		EDIT. 00
		40dB - DEMPELEDD - 75Ω			

4-466

MIXER / OUTPUTSTAGE



ISSUE DATE/SIGN.	APPR. DATE/SIGN.	SCALE	Telettra Norge a.s.		PAGE
		PROJ.	RPLM.FOR	RPLD.BY	OF
		TRACED	CODE		EDIT.
		MIXER / OUTPUTSTAGE			

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APPENDIX-D:CALCULATION OF ISDN POWER INSIDE THE 1+1 SYSTEM BANDS.

In section 3.2.3. theoretical average power of the three line codes inside the 1+1 system bands has been used to calculate the minimum required NEXT-attenuation. This appendix will show the calculations done to obtain the average power for these frequency bands. The frequency bands are 20 to 36 kHz and 71 to 87 kHz.

First the power spectrums must be calculated by using equation (3.1.3.)

$$G(f) = r\,|P(f)|^2 \{R_a(0) + 2 \sum_{n=1}^{\infty} R_a(n) \cos(\frac{2\pi fn}{r})\} \qquad (D.1.)$$

where r is the transmission rate, P(f) the Fourier transformed transmitted pulse, and R_a the amplitude correlation. A rectangular pulse was used.

$$P(f) = \frac{1}{r}\,\mathrm{sinc}(\frac{f}{r}) \qquad (D.2.)$$

Amplitude correlations for the three line codes are

$$\text{AMI:} \quad R_a(n) = \begin{cases} A^2/2 & n = 0 \\ -A^2/4 & n = 1 \\ 0 & n \geq 2 \end{cases}$$

$$\text{4B3T:} \quad R_a(n) = \begin{cases} 2B^2/3 & n = 0 \\ -5B^2/32 & n = 1 \\ -3B^2/32 & n = 2 \\ 0 & n \geq 3 \end{cases} \qquad (D.3.)$$

$$\text{2B1Q:} \quad R_a(n) = \begin{cases} 5D^2/4 & n = 0 \\ 0 & n \geq 1 \end{cases}$$

where A, B, and D are the distance between the amplitude levels of AMI, 4B3T and 2B1Q, respectively.

By substituting (D.2.) and (D.3.) into (D.1.) the following power spectra can be found

$$G_{AMI}(f) = \frac{1}{r} A^2 \operatorname{sinc}^2(\frac{f}{r}) \sin^2(\frac{\pi f}{r})$$

$$G_{4B3T}(f) = \frac{1}{r} B^2 \operatorname{sinc}^2(\frac{f}{r}) [\frac{2}{3} - \frac{5}{16} \cos(\frac{2\pi f}{r}) - \frac{3}{16} \cos(\frac{2\pi f}{r})] \quad (D.4.)$$

$$G_{2B1Q}(f) = \frac{5}{4r} D^2 \operatorname{sinc}^2(\frac{f}{r})$$

The transmission rate r is 160 kbaud, 120 kbaud and 80 kbaud, for AMI, 4B3T and 2B1Q, respectively. The next step is to determine the amplitude steps and find the power inside the given frequency bands. The calculation for 2B1Q will be given here, and can be done analogously for AMI and 4B3T.

The total power of 2B1Q must be 11 dBm (9 dBm for 4B3T and 10 dBm for AMI), which implies that

$$10 \log \left[\int_0^\infty G_{2B1Q}(f)\, df \right] = 11 \text{ dBm} = 0 \text{ dBr} \qquad (D.5.)$$

and by substituting (D.4.) into (D.5.) one can find

$$10 \log (D^2) = -10 \log \left[\int_0^\infty \frac{5}{4\ r} \operatorname{sinc}^2(\frac{f}{r})\, df \right] \qquad (D.6.)$$

$$10 \log(D^2) = 2.09 \text{ dBr} \quad \text{(for 2B1Q where r = 80 kbaud.)}$$

The total power in a given frequency band between f_l and f_u can now be found as

$$S = 2.09 + 10 \log \left[\int_{f_l}^{f_u} \frac{5}{4\,r} \operatorname{sinc}^2\left(\frac{f}{r}\right) df \right] \qquad (D.7.)$$

The power is given in dBr, which is converted into dBm by adding 11 (D.5), resulting in the following for the actual frequency bands.

Table D.1: Average power in the frequency bands used by the 1+1 system, given for three different line codes.

Line code	$P_{20-36\ kHz}$ [dBm]	$P_{71-87\ kHz}$ [dBm]
2B1Q	5.2	-17.2
4B3T	3.2	- 2.8
AMI	0.1	2.3

APPENDIX-E: INTEGRATED CIRCUITS FOR ISDN BASIC RATE ACCESS.

Here follows a list of UIC and SIC-manufacturers, the system interface used, the price and when they are expected to be available.

4B3T UIC:

PEB 2090(1/2) Siemens AG.
System interface: IOM.
Price indication: 1989
In batches of 5.000: USD 237.00
10.000: USD 162.00
20.000: USD 156.00

Available from 1989.

Am 2090(1/2) Advanced Micro Devices, Inc.
Second source Siemens AG.

MTC-2071 Mietec.
System interface: IOM compatible.

Price indication:	1989	1994
	USD 158.00	10.00

Available from 1989.

PCB2390(1/2) Philips
System interface: IOM compatible.
Price indication: 1989
USD 58.00

Available from 1989.

SGS-Thompson
System interface: IOM compatible.
Not available, only for internal use.

Alcatel

System interface: IOM compatible.

Not available, only for internal use.

2B1Q UIC:

PEB 2091(1/2) Siemens AG.

System interface: IOM.

Price indication:	1990	1991
In batches of 5.000:	USD 100.00	72.00
10.000:	USD 68.00	50.00

Samples available from 1989.

Am 2091(1/2) Advanced Micro Devices, Inc.

Second source Siemens AG.

MT 8910 Mitel Corporation

System interface: ST-bus.

Price indication:	1989	1991	1993	1995
In batches of 10.000:	USD 47.00	45.00	39.00	27.00
50.000:	USD 37.00	34.00	25.00	20.00

Available from 1989.

MC145472 Motorola Inc.

System interface: IDL

No price indication.

Available from 1990.

TP3410J National Semiconductor Corporation

System interface: IOM compatible.

Price indication:	1990	1994
In batches of 50.000:	USD 33.00	27.00

Samples available from 1990.

SIC:

PEB 2080 Siemens AG.

System interface: IOM

Price indication:		1989	1990	1991
In batches of	5.000:	USD 14.00	13.00	12.00
	10.000:	USD 10.00	9.00	8.00
	15.000:	USD 9.00	8.00	7.50

Available from 1989.

Am 2080 Advanced Micro Devices, Inc.

Second source Siemens AG.

MCT-2072 Mietec.

System interface: IOM compatible

No price indication.

Samples available 1989.

29C53 Intel Corporation.

System interface: IOM compatible.

Price indication:	1989
	USD 22.00

Available from 1989.

TP3420J National Semiconductor Corporation

System interface: IOM compatible.

Price indication:		1990	1991	1992	1993	1994
In batches of	10.000:	USD 15.00				
	30.000:	USD	11.00			
	40.000:	USD		9.00	8.00	7.50

Samples available from 1989.

MC145474 Motorola Inc.

System interface: IDL

Price indication:	1990	1995
In batches of 1.000:	USD 14.00	8.00

Available from 1990.

MT 8930 Mitel Corporation

System interface: ST-bus.

Price indication:		1989	1990	1991	1992	1993
In batches of	5.000:	USD 21.00	16.00	14.00	13.00	11.50
	10.000:	USD 16.00	13.00	12.00	10.00	9.00

Available from 1989.