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2 ND EUROPEAN FREQUENCY AND TIME FORUM NEUCHÂTEL - 16 - 17 - 18 MARCH 1988

Session TF-05 Number 3

Time dissemination via the LF transmitter DCF77 using a pseudo-random phase-shift keying of the carrier

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Summary

In addition to the amplitude modulation by second markers the carrier of DCF77 is phase modulated using a pseudo-random phase-shift keying (PRPSK). The PRPSK is accomplished according to a pseudo-random binary sequence of maximal length with 2^9 states. Therefore the mean value of the carrier phase is not influenced. At the receiver side the arrival time of the PR cycles can be determined by cross correlation. Time comparisons with reference to UTC(PTB) demonstrate the capability of the cross correlation method combined with PRPSK also in the LF band. This technique makes better use of the frequency spectrum available and results in a more precise LF time distribution. The arrival times of the pseudo-random cycles received 273 km from the transmitter fluctuate during day time (except in winter) within a standard deviation of less than half a period (< 6.5 µs) of the carrier.

1. Introduction

Due to a great number of harmful interferences, atmospheric disturbances and man-made noise, amplitude-modulated time signals in the LF band can often be recorded only with narrow bandwidth. Narrow-band reception has, however, the consequence that the accuracy with which the arrival times of the amplitude-modulated time signals are ascertained decreases with the degree of filtering. When the arrival times of the amplitude-modulated time signals received are to be determined with as low an uncertainty as possible, it is therefore necessary to use a broad band. Broad-band reception results, however, in a high liability to interferences. This can in part be compensated for by suitable demodulation techniques, averaging methods and other methods of noise suppression. All conventional reception and averaging techniques for amplitude-modulated time signals have, however, the disadvantage that any phase-synchronous harmful interference which falls within the effective bandwidth results in spurious modulation which superposes the time signals and cannot be separated from them. A technique with which this disadvantage is avoided because it allows various signal components in the same frequency range to be separated and which also excels by a high immunity from disturbance is the cross correlation /1/. The advantages which are known from the theory of the correlation technique can, however, be made full use of only when pseudo-random noise (PRN) signals are used /2/. In contrast to the conventional second markers, the duration of correlation is shorter for PRN signals, which means a higher resolution in time. For this reason, in the following, a correlation method adapted to the transmitter DCF77, combined with PRN signals is described which optimally utilizes the bandwidth available.

2. Cross correlator configuration for time transmission

Fig. 1 shows the principle of a correlator configuration for time transfer. Essential elements of this configuration are the delay network and the multplier with subsequent integrator.

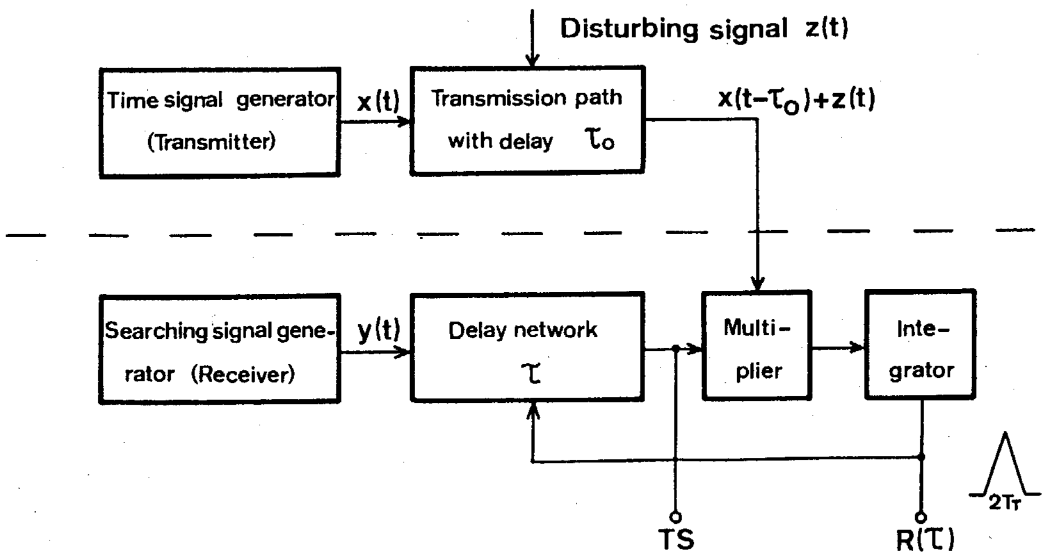

Fig. 1: Simplified cross correlator block diagram: Arrival time determination of the received time signal (TS) by means of the cross correlation function $R(\tau)$.

Mathematically, the measuring operation which is carried out by the cross correlator can be described by the following equation:

$$R(\tau) = \lim_{T\to\infty} \frac{1}{2T} \int_{-T}^{+T} \left[x(t-\tau_0) + z(t)\right] \cdot y(t-\tau)\, dt \qquad (1)$$

When the averaging is expressed by a bar instead of the time integral, eq. (1) may be written as:

$$R(\tau) = \overline{x(t-\tau_0) \cdot y(t-\tau)} + \overline{z(t) \cdot y(t-\tau)} \qquad (2)$$

Equations (1) and (2) describe the principle of the cross correlator: On the transmission path, the time signal $x(t)$ emitted by the time signal transmitter is superposed by a harmful interference $z(t)$. Furtheron, it undergoes a delay τ_0 before it arrives at the receiver end at an input of the multiplier. To the second input of the multiplier, via a delay network, the search signal $y(t)$ is supplied which is generated in the receiver and has the same properties as $x(t)$.

In the subsequent averaging of the product signal over a sufficiently long integration time, the second term in eq. (2) disappears when $y(t)$ and $z(t)$ are incoherent for all values of τ and are statistically independent. At its output the correlator thus provides a correlation function $R(\tau)$ which is dependent only on $x(t)$ and $y(t)$ and free from superposed interferences. When the time displacement τ due to the delay network is equal to the delay τ_0 on the transmission path, $R(\tau)$ becomes a maximum and the search signal $y(t)$ tracked by the delay line represents the time information which is emitted by the transmitter and largely free from interferences.

Pseudo-random noise signals have proved to be particularly suitable for the correlation technique. Such signals have a triangular correlation function with a correlation time of the duration of 2 chips. As will be shown in the following, this correlation function allows to easily obtain a control signal with correct sign in order to adjust the delay network with the aid of a control circuit to the maximum of $R(\tau)$.

3. AM-compatible, pseudo-random phase-shift keying of the DCF77 carrier

For DCF77, the modulation with a phase noise is made as pseudo-random phase-shift keying of the carrier according to a binary random sequence. In order to affect the use of DCF77 as a standard frequency transmitter to as small an extent as possible, a small phase swing $\Delta\varphi = 10°$ is used. At a period duration of 12.9 µs, this corresponds to changes in the phase time of $\pm$ 0.4 µs.

For the production of the pseudo-random sequence, a nine-stage shift register is used which is feedbacked according to Fig. 2 and clocked with a clock frequency $f_T = 645.8\overline{3}$ Hz. This clock frequency is as close as possible to the upper frequency limit at which the transmitting antenna is still in a position to follow the phase-shift keyings. It is a subharmonic of the carrier frequency and can easily be derived from it by frequency division (77500:120).

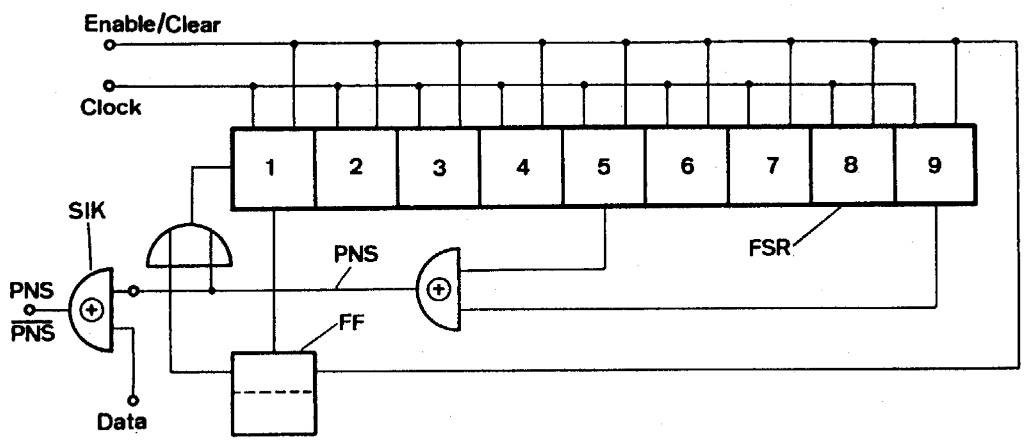

Fig. 2: Nine-stage feedback shift register FSR for generation of the pseudo noise sequence PNS. Flipflop FF forces the FSR out of the "all zero state". Sequence inversion keying (SIK) for data modulation.

With the specified number of stages $n = 9$ and the duration for 1 chip of $T_T = 1/f_T \approx 1.55$ ms the following values result for the length N and the duration T_K of one noise cycle: $N = 2^9 = 512$, $T_K = N \cdot T_T \approx 793$ ms.

Because of the signal structure of DCF77, the pseudo-random generator is not operated continuously. An enable/clear pulse serves to start it 0.2 s after the beginning of each second marker and to stop it after completion of one complete noise cycle, about 7 ms before the beginning of the next second marker. This method ensures that the falling edge defined as beginning of the second remains undisturbed and the noise cycles fall within the range of the 100 % amplitude also in the case of extended 0.2 s second markers. As in each noise cycle the states zero and one occur equally frequently (256 times each), the mean value φ_m of the carrier phase subjected to pseudo-random phase-shift keying between $\varphi_m + \Delta\varphi$ and $\varphi_m - \Delta\varphi$ also remains unchanged.

The flipflop FF ensures that the pseudo-random generator can start from the state zero. At the beginning of the cycle, FF is set to one so that at the first clock pulse after the release, a one is read into the first register stage. As soon as this one has traversed the first stage, FF is set back to the state zero.

The feedbacked shift register is followed by an exclusive OR gate with which binary data can be added by modulo 2 adding to the pseudo-random sequence generated /3/. In the case of the data state zero, the pseudo-random sequence PNS traverses this data adder SIK without a change, whereas in the data state one, it takes the inverted form $\overline{\text{PNS}}$. With each noise cycle, one bit is transmitted, the binary information transmitted with SIK (sequence inversion keying) being the same as the conventional one which is emitted by pulse duration modulation of the AM-second markers /4/. In the case of SIK, only the minute marker identification is different: Instead of omitting the 59th second marker, 10 inverted $\overline{PNS}$ sequences are transmitted in the seconds 0 to 9.

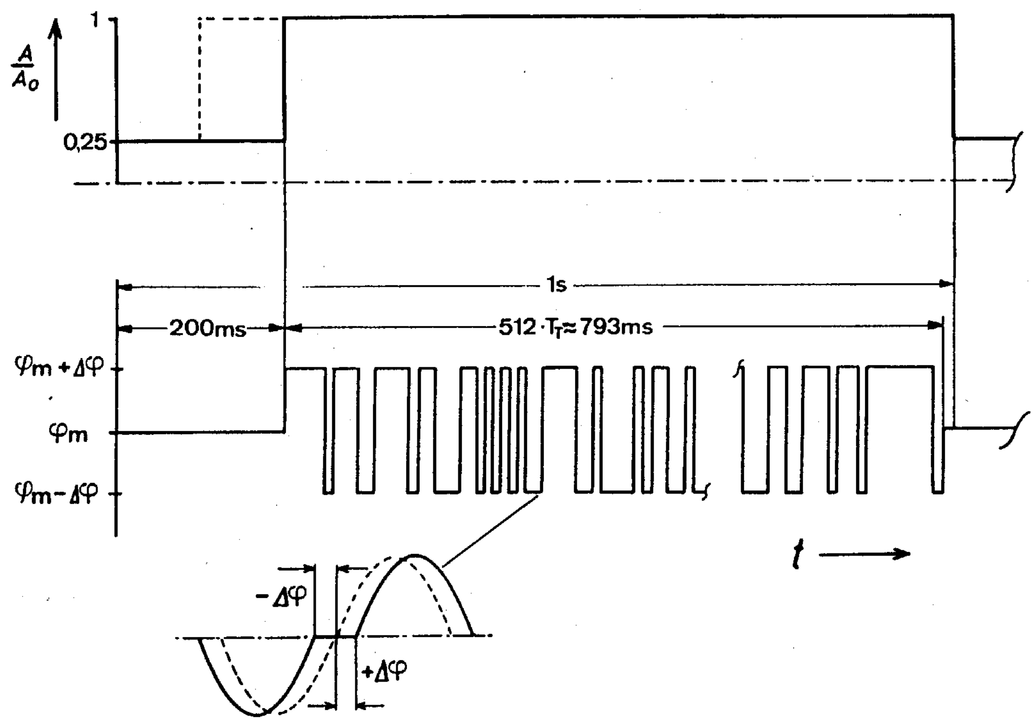

Fig. 3: Amplitude and phase of the DCF77 carrier during one second.

To illustrate the relations described, in Fig. 3 the amplitude response and the phase response are plotted over the duration of a second for the data state zero. To the state "low" at the output of the data adder the phase $\varphi_m + \Delta\varphi$, and to the state "high" the phase $\varphi_m - \Delta\varphi$, were assigned. In the steering signal the phase shift keyings and the amplitude changes at the beginning and at the end of the time markers take place in positive zero crossings of the carrier at the defined times. This ensures that the carrier, the amplitude-modulated time signals and the noise cycles are always phase-locked. At the receiving end, these sudden changes are not, however, recognizable, as phase and amplitude build up only with great time constants.

4. Frequency spectrum of the DCF77 signal

In the frequency domain the spectral density envelope of the DCF77 signal follows a (sinx)/x distribution with minima at multiples of the clock frequency f_T. The envelope is filled with spectral lines whose frequency distance is f_T/N. The amplitudes of the line components are dependent on $\sin\Delta\varphi$ while the carrier amplitude is a function of $\cos\Delta\varphi$ /5/. Figure 4 shows the measured frequency spectrum of the DCF77 PN code. Due to the resolution bandwidth of the spectrum analyser the spectral lines degenerate to a continuous spectral density envelope.

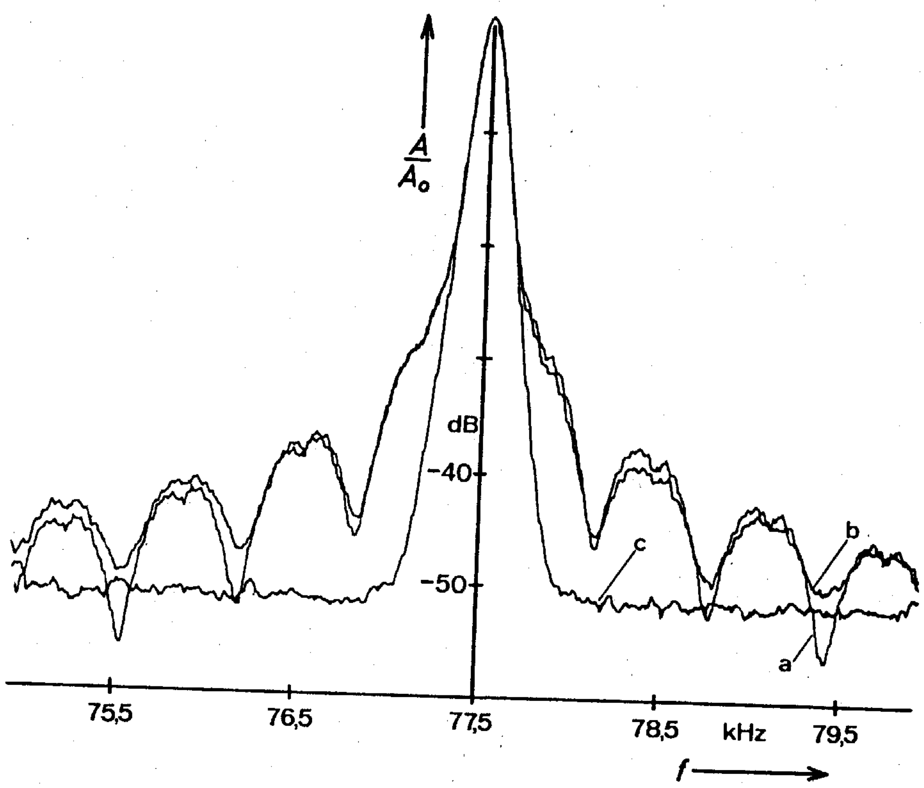

Fig. 4: Spectrum of the DCF77 signal measured with 100 Hz resolution bandwidth at the transmitter station:
a Spectrum of the generated steering signal
b Spectrum of the radiated signal
c Carrier without modulation

5. DCF77-phase noise reception, tracking and data readout

Fig. 5 shows the basic circuit of the receiving arrangement for the evaluation of the carrier noise developed at the PTB. The voltage-controlled quartz oscillator 1 provides the output frequency 6.2 MHz which allows the carrier frequency 77.5 kHz and the decadic frequencies 10 Hz and 1 Hz to be obtained by integer frequency division. After frequency division by 80 in the divider 2, the oscillator signal converted to 77.5 kHz is supplied to the phase detector 3 where it is compared with the DCF77 signal received. The received signal is supplied to the detector 3 via the band pass 5 ($\Delta f = 4\ f_T$) and the amplifier 4. If there is a phase difference between the converted oscillator signal and the received carrier signal, the phase detector 3 supplies a control voltage of corresponding size and sign which readjusts the quartz oscillator 1 through a PI control system element 6 until there is phase coincidence at the phase detector 3. This method allows a voltage to be obtained at the phase detector which on average becomes approximately zero. In the adjusted state it is superposed only by the rapid fluctuations produced by the pseudo-random phase-shift keying. The phase detector output signal thus represents the noise sequence received which is required as reference signal for the correlation control loop in the lower part of the circuit.

The phase detector 8 mixes the oscillator signal shifted by 90° with the signal received. The mixed product is a voltage which is proportional to the envelope of the carrier. The signal shaper 9 following the phase detector, which consists of a comparator, converts the envelope signal into the pulse sequence Z which according to the coding of DCF77 consists of second pulses 0.1 s and 0.2 s long. These circuit elements practically represent a synchronous demodulator for the amplitude-modulated time signals.

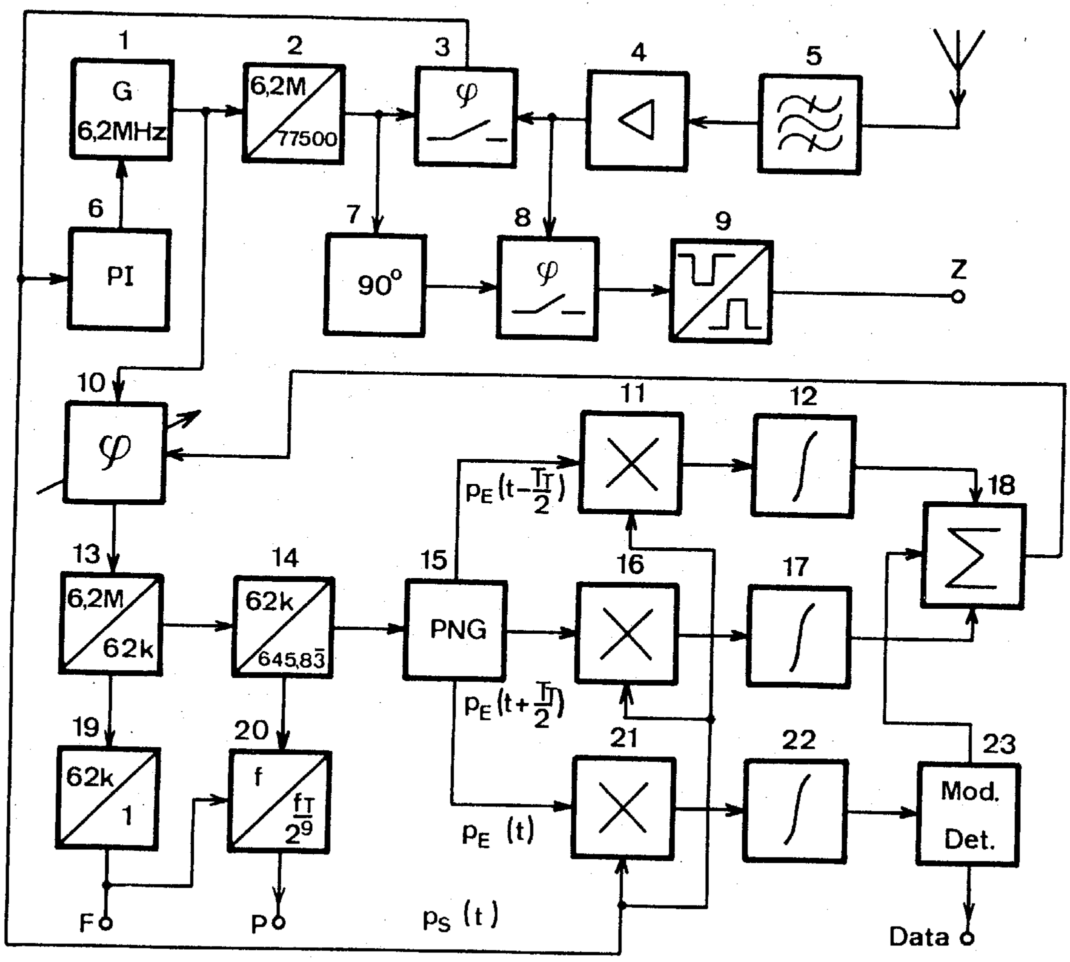

Fig. 5: Block diagram of a delay-lock tracking receiver with data readout.

The lower part of the block diagram represents a delay-lock loop with data readout. The pseudo-noise generator 15 produces three random sequences displaced in time. The two sequences $p_E(t - T_T/2)$ and p_E $(t + T_T/2)$ displaced by T_T are multiplied in the multipliers 11, 16 by the sequence $p_S(t)$ coming from the transmitter. For averaging, the product signals are subsequently supplied to the integrators 12, 17. The subtraction 18 of the integrator output voltages finally provides a correction signal to drive the phase shifter 10. If the loop is closed, the sequence $p_E(t)$ is brought to coincidence with the received sequence $p_S(t)$ and the multiplication in 21 and integrating in 22 of both sequences yield a maximum (correlation) or a minimum (anticorrelation) depending on whether a data zero or a data one is being transmitted. At the end of each noise cycle, the modulation detector detects whether correlation or anticorrelation is concerned and outputs the corresponding binary data. Furthermore, in the summation stage 18, it changes the incoming integrator output polarity so that regardless of whether a data zero or a data one is being transmitted, the control voltage will always be of the proper polarity for tracking. The clock frequency for the PN generator 15 is derived from the divider 13 (divided by 100) and the divider 14 (divided by 8·12 = 96).

Divider 19 and binary counter 20 provide the 1 Hz pulses F and P (cf. Fig. 6). These pulses enable and clear the PN generator, and before the beginning of each cycle, they erase the integrators. Furthermore, they cause the data readout at the end of each cycle.

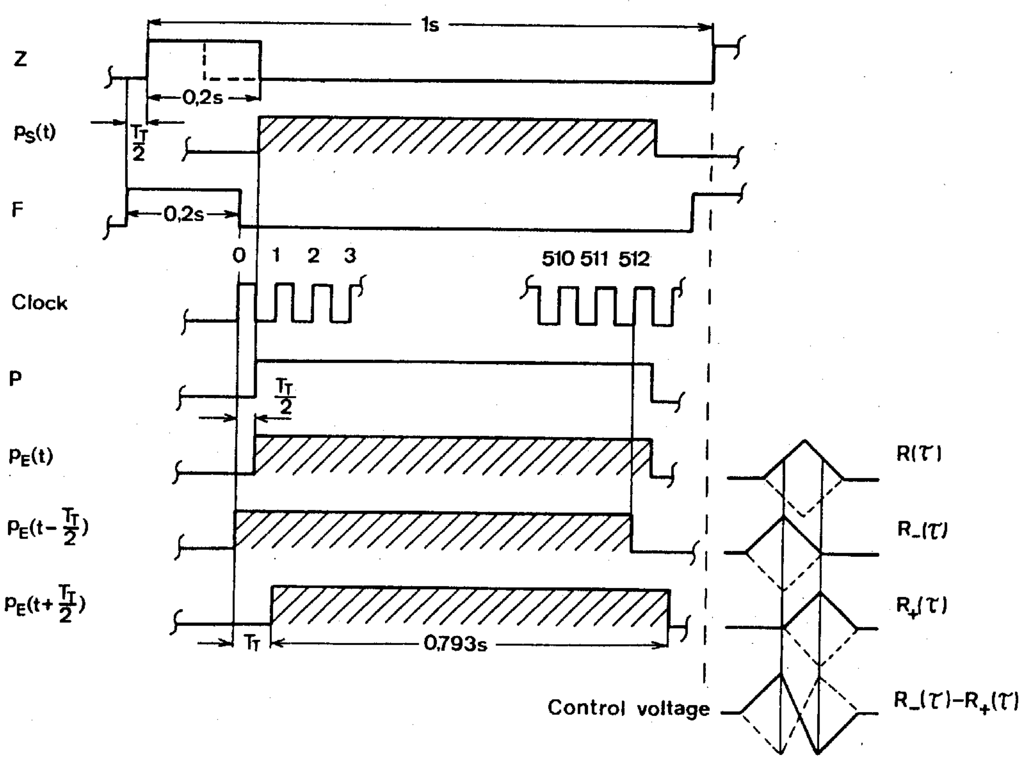

Fig. 6: Timing diagram of the individual pulses Z, F, P and the pseudo random noise sequences generated and received in the delay-lock receiver of Fig. 5.

6. Measurement results

Investigations and measurements carried out at the PTB using the PRPSK method have confirmed that both the binary data can be reliably recognized and the arrival times of the noise signals can be determined with a small uncertainty. The superiority of the pseudo-random noise signals chiefly appeared on days with strong atmospheric disturbances due to lightning discharges. Fig. 7 gives some measurement values. They show the time differences UTC(PTB) - UTC(DCF77) + K measured daily at 11.00 h and 23.00 h over about 200 days in Braunschweig (situated at a distance of 273 km from the transmitter). UTC(DCF77) designates the rising edge of the pulse F defined in Fig. 6 and K means a delay constant. It can be seen that due to the influence of the sky wave, the fluctuations occuring in the night are greater than those occurring in the day-time. It can further be seen that the scatter increases during the day when from about mid-October, the sky-wave share increases. The following table gives some typical standard deviations for various day-times and seasons at a distance of about 300 km from the transmitter:

	Typical standard deviation		
	of an individual value from the daily mean value (10000 s values)	of an individual value from the nightly mean value (1000 min values)	of a daily mean value from the monthly mean value
summer	2 µs	14 µs	4 µs
winter	9 µs	22 µs	10 µs

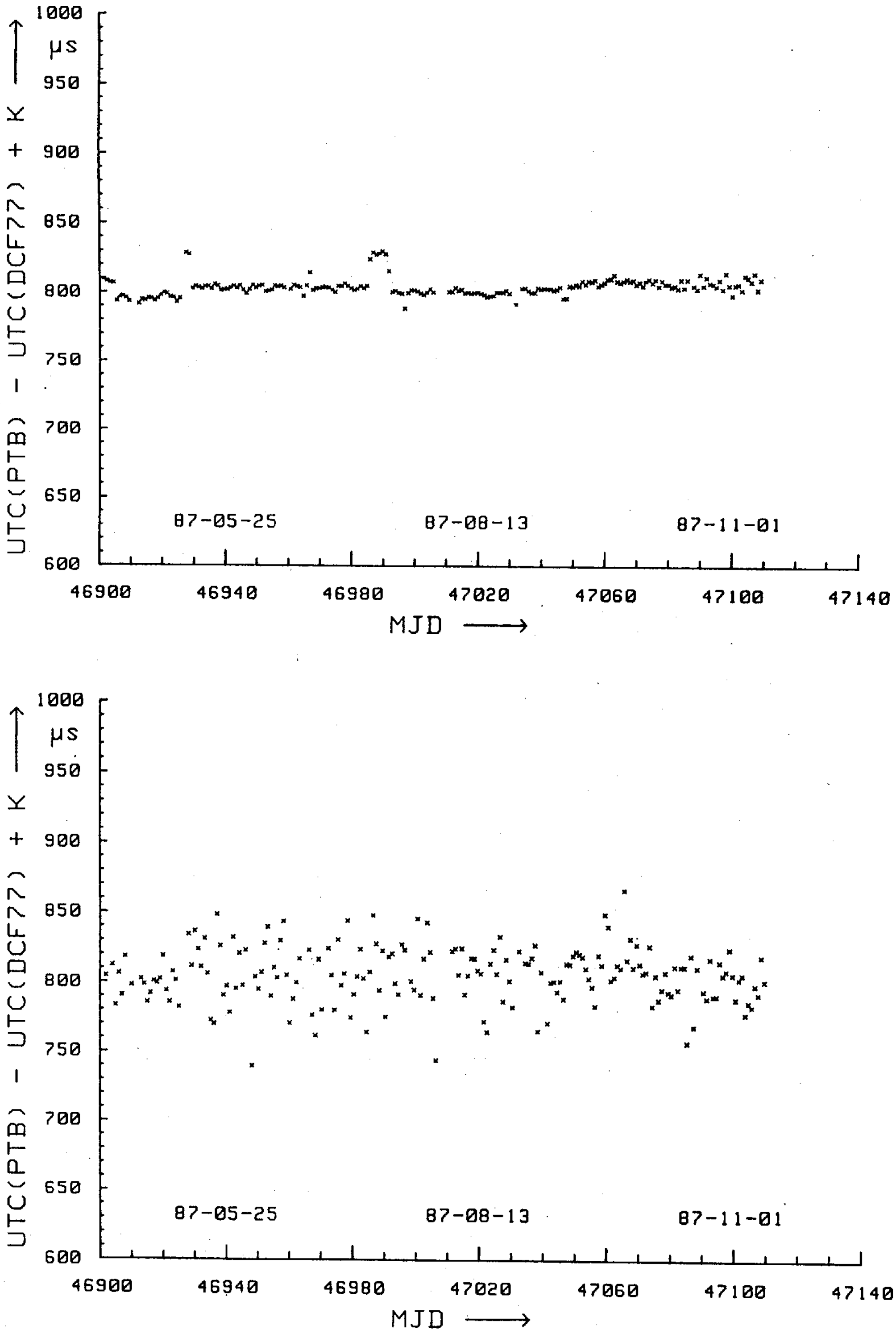

Fig. 7: Time differences UTC(PTB) - UTC(DCF77) + K measured in Braunschweig, at a distance of 273 km from the transmitter.
Upper part: single measurement values at 11 h UTC
Lower part: single measurement values at 23 h UTC

7. Conclusions

The cross correlation technique combined with pseudo-random phase-shift keying of the carrier can successfully be used also in the LF band. This technique makes better use of the frequency spectrum available and results in a more precise LF time distribution and increased reliability. Under usual receiving conditions, the arrival times of the transmitted pseudo-random noise cycles can be obtained with a smaller uncertainty than is possible with the common AM second markers.

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