International Telecommunication Union

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(09/2021)

Protection criteria for systems in the standard frequency and time signal services

TF Series

Time signals and frequency standards emissions

International Telecommunication Union

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REPORT ITU-R TF.2487-0

Protection criteria for systems in the standard frequency and time signal services

(2021)

Scope

Standard Frequency and Time Signal (SFTS) services constitute an important means to provide accurate time and frequency traceable to national standards. The individual systems and their characteristics are described. The aim of this Report is to provide protection criteria for SFTS systems. This information should be used for studies regarding the sharing/compatibility with other services and systems, e.g. non-beam wireless power transmission (WPT) systems which have been described in Report ITU-R SM.2303 – Wireless power transmission using technologies other than radio frequency beam.

Keywords

Standard frequency and time signal services, protection criteria, interference, wireless power transfer

Abbreviations/Glossary

AGC	Automatic gain control
AM	Amplitude modulation
ASK	Amplitude shift keying
AWGN	Added white Gaussian noise
BER	Bit error rate
BCD	Binary coded decimal
BPSK	Binary phase shift keying
DST	Daylight saving time
OEM	Original equipment manufacturer
PM	Phase modulation
RCC	Radio-controlled clock
SFTS	Standard frequency and time signal service
SNIR	Signal to noise-plus-interference ratio
SNR	Signal-to-noise ratio
UT1	Universal time (astronomical time)
UTC	Coordinated universal time
WER	Word error probability
WPT	Wireless power transfer

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1 Introduction

This Report will deal with the description and protection criteria for systems in the Standard Frequency and Time Signal (SFTS) services. The individual systems and their characteristics will be described. The aim of this Report is to provide protection criteria of SFTS systems. The systems affected are those operating at 4 996 kHz in Russia; 9 996 kHz in Russia; 14 996 kHz in Russia; 20 kHz globally; 40 kHz in Japan; 50 kHz in Russia; 60 kHz in the United Kingdom, the United States, and Japan; 66.6 kHz in Russia; 68.5 kHz in China; 77.5 kHz in Germany; 100 kHz in China; and 162 kHz in France. This information is intended for performing analyses of radio-frequency interference impact on SFTS from radio sources other than those used in the SFTS. In particular this information should be used for compatibility study of non-beam wireless power transmission (WPT) systems.

2 Recommendation ITU-R P.372 environmental noise

Recommendation ITU-R P.372 provides a background on radio noise. The graphical curves in Recommendation ITU-R P.372 provide the information required for an assessment of the expected receiving environment for the standard time and frequency signals as listed in the Introduction. The graph included below as Fig. 1 reproduces Fig. 2 from Recommendation ITU-R P.372.

The quantities plotted are the external noise figure $F_a = 10 \log f_a$ *(in dB)* with $f_a = p_n/(k\, T_0\, b)$, with the designations:

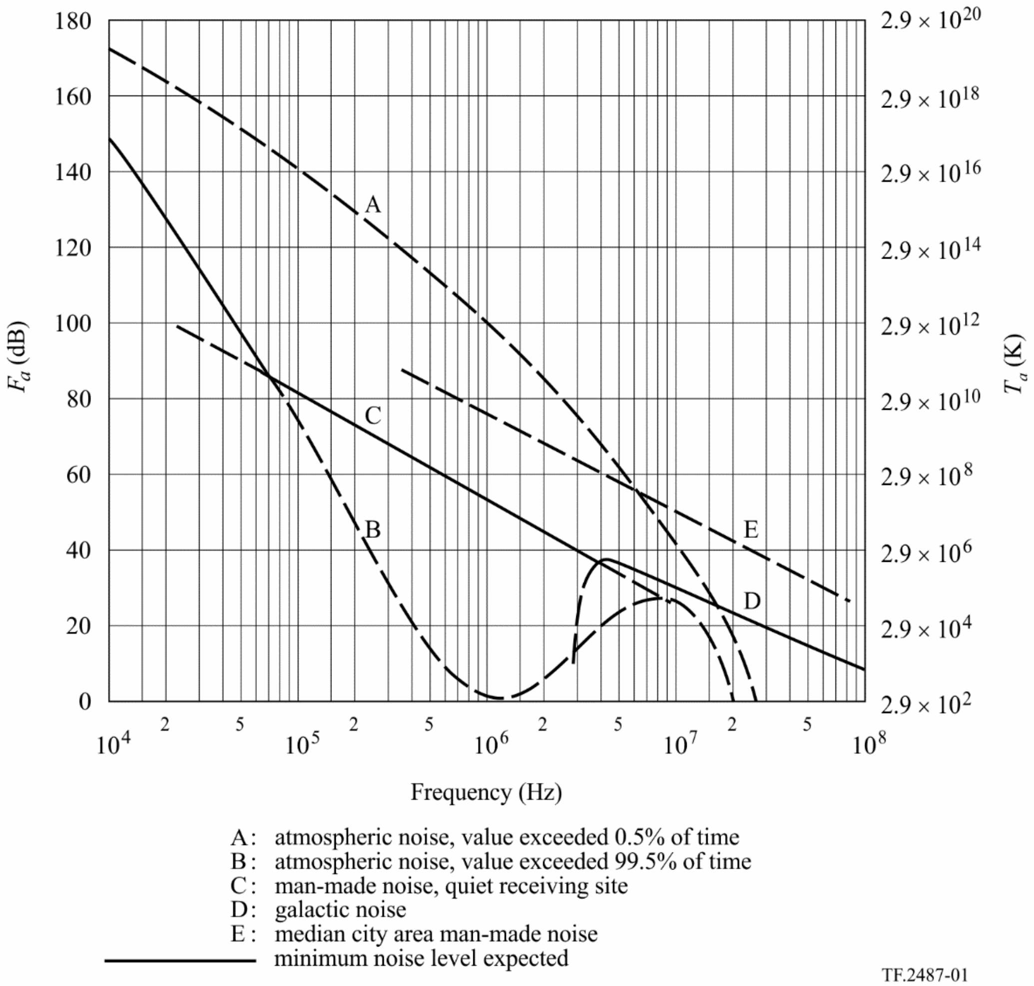
p_n:	available noise power from an equivalent lossless antenna
k:	Boltzmann’s constant = 1.38×10^{-23} J/K
T_0:	reference temperature (K) taken as 290 K
b:	noise power bandwidth of the receiving system (Hz).

The second vertical axis represents $T_a = f_a \times T_0$.

Curve A represents the upper limit of atmospheric noise that is only exceeded for 0.5% of the time. It indicates a value of F_a around 150 as the expected worst case during nearby lightning activity in the interesting frequency range. Curve B represents the lower limit of atmospheric noise that is exceeded during 99.5% of the time. At, e.g. 60 kHz are read F_a values of around 90, for both atmospheric noise (curve B) and man-made noise (curve C). It shows that atmospheric noise is the dominant noise source affecting SFTS.

FIGURE 1

F_a versus frequency (10^4 to 10^8 Hz)



Recommendation ITU-R P.372 also provides the following reference for calculation of the field strengths for a given F_a. The physical size of the receiving antenna for long-wave SFTS is only centimetres and very short compared to a wavelength between 2 km and 5 km. The actual antenna in a given receiver will have its own inherent gain and this gain/attenuation will apply equally to both the wanted signal and all other received external signal sources such as interference and environmental noise. The ratio between the aggregate background noise and the wanted signal is an important requirement for reception quality.

"For a short ($h << \lambda$) vertical monopole above a perfectly conducting ground plane, the vertical component of the RMS field strength is given by:

$$E_n = F_a + 20 \log f_{MHz} + B - 95.5 \qquad \text{dB}(\mu\text{V/m}) \qquad (1)$$

where:

- E_n: field strength in bandwidth b Hz
- f_{MHz}: centre frequency (MHz);
- B: 10 log (bandwidth Hz).

To give an example, this translates for a typical receiver used for MSF at 60 kHz in the following way.

Such a receiver has a narrow bandwidth crystal filter which will result in a noise bandwidth around 50 Hz in the receiver detector. The RMS electrical field strength of the background noise is thus:

For F_a of 150 $E_n \geq 47$ dBμV/m for less than 0.5% time as the worst-case level.

For F_a of 90 $E_n \geq -13$ dBμV/m for 99.5% time as the maximum usual level.

Atmospheric noise usually dominates for the worst case and this is closely associated with thunderstorms/lightning activity. So in different geographic areas around the world the levels may be significantly lower on average in some areas than others. For example, the UK is in a temperate zone with relatively low levels of thunderstorm activity throughout the year. This consideration should apply to most of the services discussed in this Report.

3 Information on the UK SFTS service

The operational characteristics of the UK MSF SFTS service and the typical protection ratios for Gaussian interference are provided in this Report. There is also a description of the operational parts of a typical domestic radio-controlled clock receiver derived from manufacturers' data sheets.

The section includes an analysis of the theoretical expected behaviour of domestic clock receiver units given the selectivity characteristics of the antenna. It also describes the behaviour that will result with standard ON/OFF keying modulation and the data format utilised by MSF where the time information is sent as a sequential block of data over a one-minute interval.

The expected average time to achieve a good time signal lock decode versus the received Gaussian signal-to-noise ratio (SNR) is also provided. It should be noted that many domestic radio-controlled clocks only allow a checking time of six minutes to achieve a data lock before they declare a state of no usable input signal. This means that a protection ratio towards 25 dB signal to aggregate noise and interference is preferable, with some allowance for imperfections in implementation, or there will be many occasions when initialisation start-up will fail after a change of battery or other routine maintenance.

The signal from MSF provides typically up to 40 dBμV/m (100 μV/m) at a range of 1 000 km from the transmitter, and the UK is in a temperate zone with infrequent thunderstorm atmospheric noise activity. Recommendation ITU-R P.372 gives a value of −13 dBμV/m for the background noise signal over 99.5% of the time, which allows a good margin for users of the signal to place their receivers with allowance for orientation coupling loss and building penetration loss, together with scope for manufactures to use cost-effective techniques in their designs.

3.1 The National Physical Laboratory MSF signal

FIGURE 2

Transmitter location

The MSF transmission from Anthorn (latitude 54° 55' N, longitude 3° 15' W), close to the border between Scotland and England, is one of the principal means of disseminating the UK national standards of time and frequency, which are maintained by the National Physical Laboratory (NPL). The effective monopole radiated power is 15 kW and the antenna is substantially omnidirectional. The signal strength is greater than 10 mV/m at 100 km and greater than 100 μV/m at 1 000 km from the transmitter. This provides coverage for the whole of the United Kingdom and Northern Ireland including the islands to the north of Scotland and Wales, plus the Channel Islands and the Isle of Man. The signal is also widely used in northern and Western Europe. The carrier frequency is maintained at 60 kHz to within two parts in 10^{12}.

The MSF time and date code format is summarised below. ON/OFF carrier modulation is used; the exact rise and fall times of the carrier are determined by the combination of antenna and transmitter. The timing of these edges is governed by the seconds and minutes of Coordinated Universal Time (UTC), which is always within one second of UT1. Every UTC second is marked by a 100 ms minimum carrier OFF period, preceded by at least 500 ms of carrier ON, and this second marker is transmitted with an uncertainty of 1 ms. See Fig. 3 for an illustration.

The first second of the minute begins with a longer period of 500 ms with the carrier off, to serve as a minute marker. The other 59 (or, exceptionally, 60 or 58) seconds of the minute always begin with at least 100 ms carrier off and end with at least 700 ms with carrier on. Seconds 01-16 carry information for the current minute about the difference (DUT1) between astronomical time (UT1) and UTC, and the remaining seconds convey the time and date code.

The time and date code information is always given in terms of UK civil time and date, which is UTC in winter and UTC+1 hour when British Summer Time is in effect, and it relates to the minute following that in which it is transmitted.

A full description of the bit allocations and their timing during each minute of the transmission is detailed in Annex 1 to this section.

The successful reception of a full set of data over the minute requires decoding of the following data contents:

– at the zero start of the initial second in the minute:
  5 consecutive 100 ms logic “1” carrier off markers;
– at the beginning of each of the next 59 seconds:
  1 initial “first” 100 ms logic “1” carrier off marker;
– in each of the next 59 seconds in a standard typical minute:
  2 data bits in the A/B timeslots as either logic “0” or “1”.

This gives a total of 182 bits (359 + 5) to be received correctly in their sequence in a complete frame across the minute interval.

The data includes sets of odd parity bits to help with error detection.

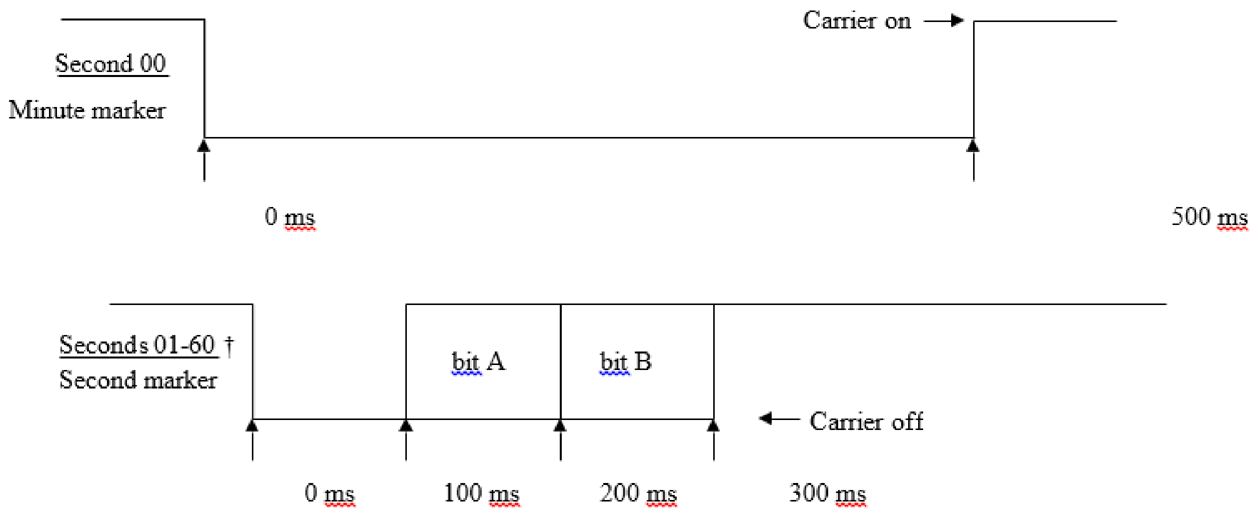
There are 18 bits currently reserved for future use set at static values of carrier ON, i.e. logic “0”.

The ON/OFF keying 100% amplitude modulation format results in wanted signals with a maximum modulation frequency of 5 Hz over short intervals (100 ms ON and 100 ms OFF as a sequence). This gives a narrow bandwidth signal format for reception purposes.

The duration of continuous carrier ON sequences are maximised to facilitate the use of MSF as a reference frequency standard signal as well as a time signal.

FIGURE 3

MSF 60 kHz time and date code



Two 100 ms time slots are used for data bits, labelled xxA and xxB for second xx.

NB signalling uses carrier “off” as logic 1 and carrier “on” as logic 0.

† Most minutes have seconds numbered 00-59, second 60 is a leap second in UTC.

3.2 ON/OFF keying Gaussian noise-to-signal requirements

3.2.1 ON/OFF keying standard detection by threshold

The basic principle of detection is to first find the average peak transmitter carrier ON signal level, denoted by *A*. Then check if the actual received signal at the sample times exceeds *A*/2 for a carrier ON or if the received signal is less than *A*/2 for a carrier OFF.

The probability of correctly receiving a data bit is influenced by the amount of additive unwanted noise on the detected signal. The statistical characteristics of the actual interferers are needed to determine correct protection ratios against those interferers.

In the absence of any defined format for WPT the only option is to assume Gaussian random noise to provide the following guidance.

Genuine transmitted signals at these frequencies are essentially constant and only partially subject to moderate ionospheric fading between day and night within the main service area. In the following a wanted signal will be considered to be a constant level and the interferer a Gaussian noise source.

An interfering Gaussian noise signal can be characterised by its RMS level denoted by the standard deviation, and the cumulative percentage chance in time that the envelope will exceed a given level is available from the statistics of a binomial distribution.

The distribution characteristics of the voltage Gaussian noise are symmetric about zero: the chance of a negative excursion and the chance of a positive excursion being greater than a given value are equal. The aggregate received signal is the sum of the wanted signal and the interferer, assessed here as Gaussian noise.

As the noise level rises relative to the wanted signal the chance of a decoding error in the logic 1 level and the logic 0 level are equal if the threshold is set half way between the two values of level for carrier OFF and ON. The critical value for the additive noise excursion needed to cause a detection error is $\pm A/2$ to push the aggregate signal back across to the opposite (wrong) side of the detection threshold.

3.2.2 Derivation of bit error probability and overall error probability for ON/OFF keying

The next steps only consider the actual data bits for detection, not the steady state during other intervals.

The probability total is:

$$P = P_{zero} + P_{one} = 1$$

There is a roughly equal chance overall of a wanted signal data bit being a logic 0 or logic 1:

$$P_{zero} = P_{one} = 0.5$$

For a Gaussian distribution the chance of an error overall in the data bits with ON/OFF keying is:

$$Probability\ error\ overall = P_{zero} * Perror_{zero} + P_{one} * Perror_{one}$$

$$P\ error\ overall = 0.5 * Perror_{zero} + 0.5 * Perror_{one}$$

$$P\ error\ overall = 0.5 * (Perror_{zero} + Perror_{one}).$$

For ON/OFF keying threshold detection there are equal error probability rates per bit:

$$Perror_{zero} = Perror_{one} = Perror_{per_{bit}}$$

$$Perror\ overall = 0.5 * 2 * Perror_{per_{bit}}$$

$$Probability\ of\ error\ overall = probability\ of\ error\ per\ bit$$

The wanted signal amplitude provides the reference levels "Vsig" = 0 or peak *A* volts.

The threshold for detection is the central value of *A*/2.

The noise signal is assumed to be Gaussian with RMS power given by its standard deviation, σ.

The fractional area under the Gaussian curve exceeding $A/2$ will then provide the fractional chance of an error in detection, and the chance of good reception as 1 minus the fraction above the threshold.

The probability that a bit is received correctly is equal to [1 – (integral minus infinity to $A/2$ of noise distribution)].

3.2.3 Gaussian theory statistics

The following theory ONLY applies to a Gaussian noise-limited signal interferer. For WPT, the characteristics of commercially available systems are currently unknown.

For a Gaussian noise signal $v(t)$ with standard deviation σ and mean of zero.

The amplitude density probability function of v is:

$$f_{0(v)} = \frac{1}{\sqrt{2\pi\sigma^2}} e^{-v^2/2\sigma^2}$$

For the density $f_1(v)$ the interference is shifted to an average DC level A by substituting (v-A) for v.

This can be used directly to consider the case of interference to the carrier OFF state. The same statistics apply for the carrier ON state, the only difference being that a voltage shift is applied equally to both the wanted signal and the interferer mean level.

Considering the carrier OFF state, the interest is in the condition that the interferer is greater in amplitude than the detection threshold of $+A/2$ where A is the wanted carrier-on signal voltage level.

The density function is symmetrical about zero and in the case of the carrier ON state the interest is in the condition that the interferer is more negative than $A/2$ to again cross the detection threshold, but this time heading towards zero volts.

By symmetry both logic state error integrals become the same, so only the carrier OFF state needs to be considered.

The probability of a detection error in the zero carrier OFF state is also the overall bit error rate:

$$P_e = Probability\left(v > \frac{A}{2}\right) = \int_{A/2}^{\infty} f_{0(v)}\, dv$$

This is an example of the standard error function and complementary error function and can be found in standard reference tables. Excel spreadsheet functions are available to provide numerical evaluations.

3.2.4 Summary of ON/OFF keying signal to Gaussian interferer statistics

The summary Table below gives the individual bit error rate of reception for a given SNR. For the MSF signal there are 182 bits in each one-minute sequence that all need to be detected correctly.

For guaranteed reception the signal-to-aggregate-noise ratio should be above 18 dB.

An allowance will also have to be applied for imperfections in implementation in practical detectors, And 3 dB is a figure often used for this allowance for deviations from theory.

There is a relatively rapid change into a low quality of reception around a SNR value of 15 dB.

TABLE 1

ON/OFF keying bit error rates versus signal-to-Gaussian-noise ratio

Voltage SNR (dB)	Interferer Relative linear level	Cumulative good bit reception probability P	Bit Error rate 1 - P
0	1	0.691462461	0.308538
1	0.8912509	0.712604377	0.287396
2	0.7943282	0.735476912	0.264523
3	0.7079458	0.759989495	0.240011
4	0.6309573	0.785949841	0.21405
5	0.5623413	0.813035999	0.186964
6	0.5011872	0.840770877	0.159229
7	0.4466836	0.868506827	0.131493
8	0.3981072	0.895431721	0.104568
9	0.3548134	0.92061072	0.079389
10	0.3162278	0.943076851	0.056923
11	0.2818383	0.961973961	0.038026
12	0.2511886	0.976734031	0.023266
13	0.2238721	0.987239303	0.012761
14	0.1995262	0.993893617	0.006106
15	0.1778279	0.997536029	0.002464
16	0.1584893	0.999196923	0.000803
17	0.1412538	0.999799731	0.0002
18	0.1258925	0.99996431	3.57E-05
19	0.1122018	0.99999583	4.17E-06
20	0.1	0.999999713	2.87E-07
21	0.0891251	0.99999999	1.01E-08
22	0.0794328	1	1.54E-10
23	0.0707946	1	8.17E-13
24	0.0630957	1	0
25	0.0562341	1	0

3.3 Lock up times with "N bits correct sequence over 1 minute"

3.3.1 Signal structure for decoding – the need for a multiple bits correct sequence

Odd parity checking signals are included across four separate sections of the transmitted sequence and there is also a specific worst-case synchronisation check sequence, which must be received correctly before the data can be utilised.

A 5 bits carrier OFF sequence is to be detected on the first second in each minute.

A beginning of second and two data bits are to be detected in each of the next 59 seconds.

This implies correctly decoding 5 + 59 × 3 = 182 logic signals in a sequence to get a data lock.

This means that the bit error rate per individual bit will need to be raised to the power of 182 to assess the actual MSF system reception performance. This will also have an impact on the waiting period that may be required before a reception lock is achieved. With reasonably stable clocks running on their own separate internal crystal oscillators, which is now the usual situation with standard domestic receivers, the clock can run on unaided for many hours before another check synchronisation is needed. However, at first switch-on or restart after maintenance a reasonable on-demand lock up acquisition time is necessary.

It has been found that clocks from at least one major manufacturer have a start-up window of 4 to 6 minutes depending on the model before they declare that there is no signal available. This same 4 to 6-minute time window is also applied for resynchronisation checks typically at hourly intervals. Depending on the initial power up time, if there is no manual intervention the actual indicated clock time is arbitrary relative to true local time until lockup is achieved.

The RF tuner sub-unit in the clocks examined is switched on and off by the separate clock unit. The RF unit is left switched off at all other times by the microprocessor included within the separate CMOS digital internal clock to conserve the batteries.

For equipment being switched on for a fresh start-up the waiting time is critical, and if too long the consumer will probably declare the item as faulty. The signal must be available at any unconstrained time of day and especially during normal daytime working hours.

This means that a signal-to-noise-plus-interference-ratio (SNIR) biased towards perfect reception at 23 dB to 24 dB must be given to ensure lock up within the 4 to 6-minute window allowed by the clocks to avoid consumer market issues with the products both on sale and already installed in their millions.

3.3.2 Actual MSF lock up performance and signal to noise ratios

If the individual fractional bit error rate = x

Then chance of a good received bit = $1 - x$

To receive the complete time encoded data sequence requires 182 tests and all must be good

Chance of getting 182 right in a sequence = $(1 - x)$ raised to the power 182

For small x this approximates to = $1 - 182 \times x$

This condition is usually valid since the reception will be poor quality if x is not small.

The complete sequence takes one full minute for each attempt once the AGC levels are set.

A single bit in error in the sequence means another minute wait for the next attempt.

The bit error rates for a Gaussian interferer with the MSF 100% on/off keying modulation are given in Table 1 above.

An Excel spreadsheet was used to evaluate statistically the bit error rates and the respective SNRs needed to obtain results varying from one good minute of reception in 1 minute (perfect reception) down to 1 good minute of reception in 8 minutes. The latter means an average potential 8 minutes waiting time before obtaining a signal lock, which is longer than the time allowed by the domestic units sampled. If no lock was obtained in 6 minutes, they would have declared a no-usable-signal state and carried on unlocked.

As with all statistical quantities an average implies there would be many occasions with much longer or shorter waits. The classic example is tossing a coin to see if it comes up heads or tails, which is a 50% chance for each outcome, but in practice long sequences of the same outcome can be obtained. This is summarised in Table 2, where the 50% chance item is the second row in the Table.

TABLE 2

Waiting times to obtain a lock based on Gaussian SNR

		all OK	sequence length		
WAITING TIME			**182**		
Tries = minutes	**No right**	**chance**	**bit success rate**	**Perror/bit**	**SNR (dB)**
1	1	1	1	0	23.70971
2	1	0.5	0.996198742	0.003801	14.54852
3	1	0.333333	0.993981851	0.006018	14.01798
4	1	0.25	0.992411934	0.007588	13.72614
5	1	0.2	0.991195919	0.008804	13.52929
6	1	0.166667	0.990203469	0.009797	13.38292
7	1	0.142857	0.98936514	0.010635	13.26714
8	1	0.125	0.98863952	0.01136	13.17206

There is clearly a sharp transition in effect where a few dB change in wanted signal-to-aggregate-noise-and-interference ratio around 15 dB makes a large difference, and this is without any allowance for practical implementation losses.

Below 15 dB SNR there will be a state where the domestic consumer will perceive failure to synchronise and will probably declare the unit as faulty. For example, a shop sales demonstration would become unreliable or impossible. A piece of equipment which had been switched off then restarted after maintenance such as a battery change would appear to be faulty on many occasions.

3.4 Typical receiver characteristics for domestic OEM receiver units

3.4.1 Introduction

The following is based on available OEM catalogue data from suppliers of some of the sub-assemblies used in the RF tuner front end of radio-controlled clocks, and information gained by dismantling of sampled units. It summarises the typical configuration found in the domestic consumer radio-controlled clocks available in the UK.

The following may also apply worldwide to radio-controlled domestic clock receivers for the reception of on/off keyed time data signals. The individual suppliers' items were usually generic, only requiring different antenna tuning and output filter to suit the market for the local LF SFTS frequencies such as 40 kHz, 60 kHz and 77.5 kHz. The actual final signal processing following the generic RF carrier level detector was in each case via a separate dedicated controller digital clock unit which could be programmed according to market, detecting the type of signal being fed into it and processing it as required.

The RF module was in every case a small roughly thumbnail-sized printed circuit board with a custom integrated circuit covering all the functions between the separate tuned input antenna and the logic level carrier on/off output signal to the separate digital clock microprocessor.

The power consumption of the tuner when active is only approximately a milliamp and to save batteries there is a control input to allow the module to be powered off to a sub micro amp level. This results in most clocks radio control units only being active for short periods of about four to six minutes each hour. The crystal-controlled separate digital clocks then continue to run on their own oscillators until each determined checkpoint, and if no signal is detected they again continue to run without correction.

The actual receiving antennas are basically a ferrite rod with a coil overwound acting as a magnetic field detector. The physical size and magnetic properties of the ferrite rod and the directly associated number of coil turns and tuning capacitor vary as design alternatives between units. Large wall clocks could take advantage of the size of the unit to use larger components while smaller clock units had to use much smaller sampling antennas. The parallel tuning capacitor to resonate the antenna was in each case located on the small RF unit printed circuit board adjacent to the ferrite rod antenna.

This magnetic field detector inductance/capacitance parallel tuned circuit is the only tuner front end selectivity to protect the unit against any other RF signals. There is a separate narrow bandwidth crystal filter in the signal channel further down the processing chain, but this only influences the noise bandwidth of the tuner for co-channel selectivity.

The crystal filter cannot assist with front end amplifier adjacent channel signal overloading. The amplifier AGC is derived from after the narrow band filter so there is no adaption to a high level adjacent channel interferer outside of the crystal filter bandwidth and everything is dependent on the headroom allowed in the amplifier AGC level setting.

3.4.2 Structure of the tuner

The tuner shown as a schematic block diagram in Fig. 4 consists of the following, as a classic tuned radio-frequency radio combined with the wanted carrier signal level detector:

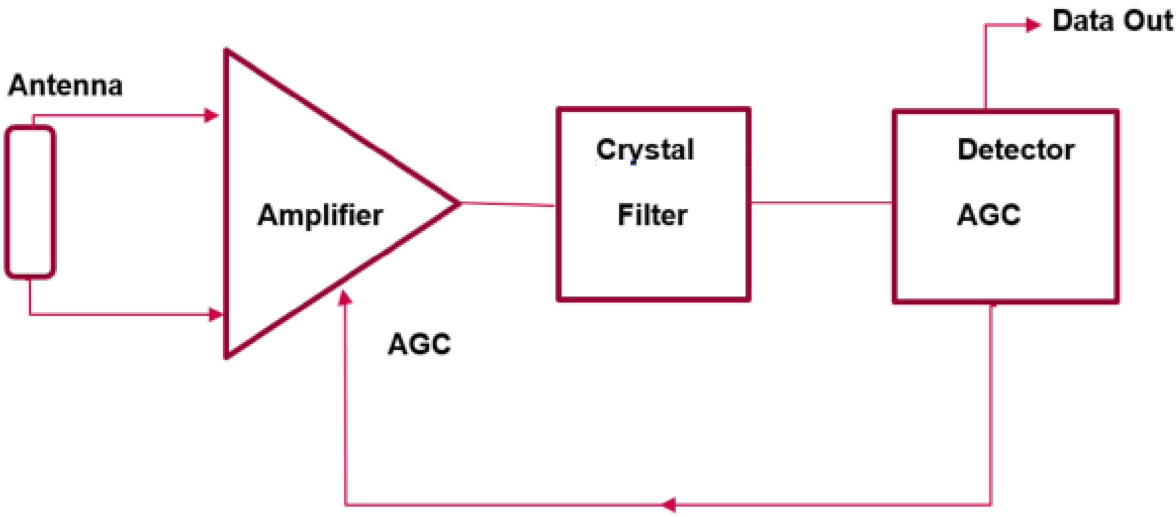
– antenna as a coil on a small ferrite rod magnetic field collector and inductor;
– the coil is parallel tuned with an external capacitor to the operating frequency (for example 60 kHz);
– a high impedance input and high gain RF amplifier with AGC feedback;
– a narrow band crystal filter centred on the wanted signal;
– a detector unit to convert the wanted RF carrier level into a logic level output signal;
– the detector works on the post crystal filter signal, also deriving the RF amplifier AGC.

The operation of the tuner/detector system can be assessed, and the behaviour is dependent on three main sub-stages:

– antenna selectivity of the parallel LC tuned circuit and its operating “*Q*” factor;
– linear dynamic range of the RF amplifier for a given wanted signal AGC setting;
– the detector decision of signal high/signal low carrier power level detection;
– the AGC system has a role to play in the setting signal levels appropriate to the detector thresholds but only within the narrow crystal filter bandwidth so it is essentially wanted signal level control only.

FIGURE 4

Tuner block diagram



3.4.3 Antenna selectivity curve

The antenna consists of a small ferrite rod, on which an inductive pick up coil is placed and parallel tuned with a capacitor. The parallel tuned circuit output voltage is fed to the following RF amplifier stage which has a high input impedance and low capacitance input. The typical data sheet figures quoted in application notes are an antenna loaded working "*Q*" of around 100 and an amplifier input impedance around 1 MΩ.

These values represent a classic tuned circuit LC input pair and an indication of its selectivity characteristic as a starting point. Clearly there are many possible combinations so typical values have been taken from the sub-unit OEM manufacturers' data sheets.

The following analysis is based on MSF at 60 kHz, but can be done for other frequencies.

Classical parallel tuned circuit equations:

$$\omega = 2\pi F$$

$$\omega^2 = 1/LC$$

$$Q = \omega L/Rloss$$

Tuned coil capacitor impedance: $$Z = Q\omega L$$

This provides a standard selectivity curve:

L with all circuit losses assigned as a series resistance

L with losses in parallel with a perfect capacitor (capacitor losses are usually much lower)

Based on complex numbers for component $impedance = R + j\omega X$

From the above:

For inductor $R_{loss} = \omega L/Q$

$Z_{ind} = R_{loss} + j\omega L$

For capacitor $Z_{cap} = 1/(j\omega C)$

$C = \frac{1}{\omega^2 L}$

Parallel pair resonance $$\frac{1}{Z_{total}} = \frac{1}{Z_{ind}} + \frac{1}{Z_{cap}}$$

At resonance the parallel impedance will be high with normal practical tuning undertaken to maximise the impedance of the parallel pair.

3.4.4 Data parameters from OEM manufacturers

The target Q of around 100 is relatively high and any value greater than 40 or 50 will result in only minor tuning adjustments. The impedance of the inductor is initially treated as a pure component to get an approximate starting value with a very small error. The error will be less than 1% for a Q of 100. An exact solution if required can be found rapidly by iteration, for example with the Excel spreadsheet optional "solver" add-in.

One IC module supplier's datasheet gave for 60 kHz a Q of around 100 and a nominal inductance of 2.4 mH as typical values.

An OEM antenna supplier's catalogue gives the following:

L = 3.2 mH approx.

Z > 65 kΩ as the working tuned circuit impedance leading to:

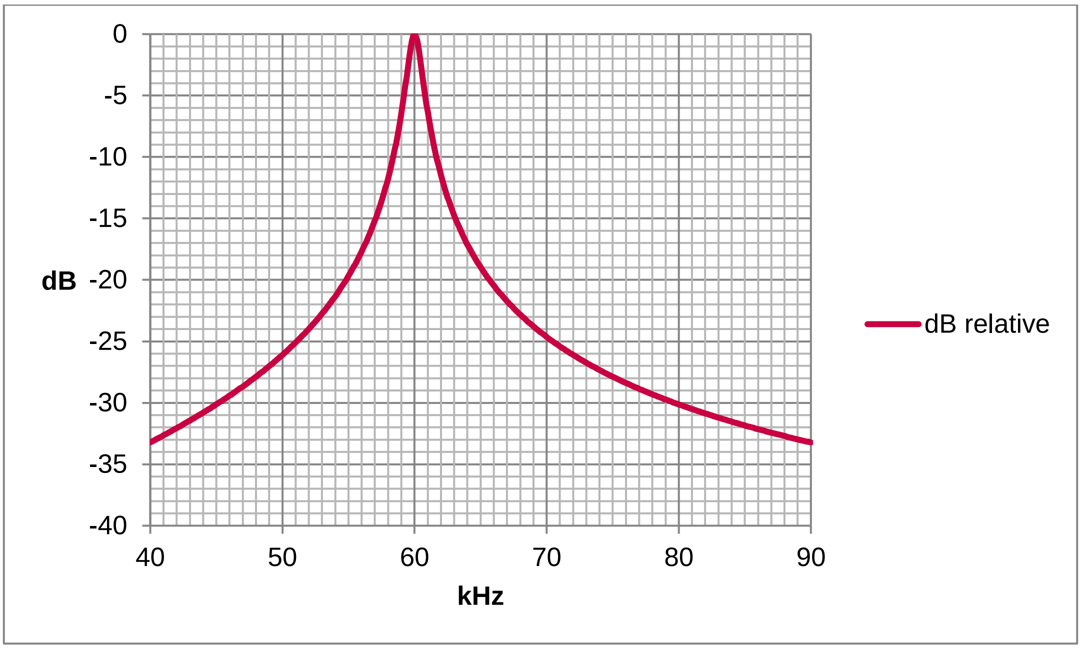
Q > 55 minimum as derived worst case

C = 2 200 pf as the parallel tuning capacitor.

This example leads to a selectivity curve as shown in Fig. 5.

FIGURE 5

Typical MSF Tuner antenna selectivity



3.4.5 Background information on ferrite rod antennas

Note ferrite rods do not perfectly couple all the magnetic field applied to them from end to end. There is maximum flux at the centre with progressive leakage towards the ends. This means an allowance for magnetic coupling loss.

The sense antenna cannot tell the difference between a wanted and an unwanted magnetic field. It will pick up both equally well with the only discrimination being the angle of the length of the rod to the direction of the individual field. Peak when aligned and a minimum not necessarily a perfect null, when at 90 degrees to the magnetic field. All locally generated magnetic fields are also by definition near fields as they will be coming from equipment much closer than the 5-kilometre wavelength at 60 kHz.

Most genuine RF transmissions at these frequencies are from large capacity hat loaded vertical masts above very significant earth mat arrangements which radiate an E field as vertical and an *H* field as horizontally polarised rings travelling outwards from the antenna as the guide to sense antenna orientation. The rod length should be at right angles to the direction towards the source transmission and in the horizontal plane to align with the *H* field.

These magnetic field pickups experience notable screening within and behind steel framed reinforced concrete buildings. This a similar action to the standard technique of mu-metal screening classic mains electrical power supply transformers to minimise leakage hum fields into adjacent electronic circuits.

A suitably pre-wound coil inductance on a former can be tuned to resonance by sliding it along the length of the ferrite rod during manufacture. As an alternative a few more turns than needed for worst case manufacturing tolerances can be used and then turns removed to tune. Once tuned the coil is fixed in final position by hot wax/glue. This takes out component parameter spreads.

The scale drawings in the OEM antenna module show a slot available from the centre to one end of the ferrite rod which could be used for this purpose. This then presumes the coil maximum external physical length is around a quarter of the rod length. The rod being around 22 mm long by 4 mm diameter giving a bobbin size around 5 mm to 6 mm long. The maximum diameter quoted was 6.8 mm and the bobbin is around 5 mm internal diameter.

A bobbin of this size wound with 0.04 mm enamelled copper wire or thinner can hold sufficient turns (estimated 230) to realise the required inductance and pickup sensitivity parameters.

The magnetic reluctance issue makes it difficult to realise very high values of actual effective permeability in a working ferrite rod configuration. The total magnetic path being along the ferrite rod length but with a return external air path between the magnetic poles being generated and the air path will limit the maximum rise in impedance which can be realised with the ferrite coil core. From experience values of an effective permeability are around 40 is possible without resorting to specialised components.

The local magnetic field is sensed via the area cross section of the ferrite rod. This field is essentially coupled entirely through the coil by the ferrite rod with very little added by any air path linkage coupling.

The result is an induced current (i.e. energy) in the coil. The basic voltage across the coil being $L \times d_i/d_t$ with magnification by the Q of the parallel tuned circuit. The resulting peak voltage output will be a peak at the parallel resonance of the LC pair. This is a voltage being generated for a given input magnetic field and the dB power conversion selectivity curve is in the form of 20 log10 (parallel circuit impedance) as shown in Fig. 5.

The flux density inside the ferrite rod is a maximum of $B = \mu\mu_0 H$ and the coupling area is a maximum of the cross-sectional area of the ferrite rod “A”. The μ value being the effective permeability for the overall assembly as noted above not the headline value for the ferrite as usually given for a toroid closed loop of the material. There will be some losses of magnetic field as not all the flux will pass down the whole length of the ferrite rod.

3.4.6 Antenna pickup sensitivity

The ferrite rod acts as a magnetic field pickup directing the flux through the coil winding with a known inductance. The pickup and transfer is not perfect but is adequate and far more practical than many other possibilities at these kilometre plus long wavelength frequencies.

The following is all basic electromagnetic theory and stems from coils of "*N*" turns being linked by a magnetic flux *B* at a frequency *F*. An estimate of the voltage generated for a given inductance and parallel capacitor used as the pickup can be derived by the rate of change of the magnetic field flux linking the detector coil via the ferrite rod core of cross section area *A*. *NB* the inductive coil has a voltage across it as $L \cdot d_i/d_t$.

$$Total_{Flux} \quad B = \mu\mu_0 H sin(\omega t) * A$$

The voltage output generated by the total rate of change of flux linkage with the coil is multiplied by the number turns on the coil each linked to the flux and an ω multiplier is obtained from the inductor differentiation and the signal becomes *cos(*ω *t).*

Finally, the inductor is part of a lightly loaded parallel tuned circuit with the external tuning capacitor so the final terminal voltage will be *Q* times higher giving a total multiplier of *Q* ω *A*.

$$Wanted\ signal\ far\ field \quad H = \frac{E\_field}{120\pi}$$

$$Tuned\ Coil\ Volts = \ Q\omega\mu\mu_0 \frac{E\_field}{120\pi} cos(\omega t) A \cdot (No. of turns\ on\ coil)$$

NB a local interferer will be a near field signal at these kilometre wavelengths and E/H fields cannot be determined by the equation above.

3.4.7 Summary of typical values from data sheets

Typical values from an OEM antenna supplier's data sheet.

Ferrite rod cross section diameter 4 mm approx. and about 22 mm long.

Coil winding placed between one end and centre of ferrite rod:

A =	π (16/4) × 10^{-6} m^2
Q	minimum of 55
μ_0 =	4π × 10^{-7} H/m standard physical constant
μ =	40 as the expected effective permeability of the ferrite rod in situ

No of coil turns = estimated at 230 with a μ of 40 to give required matching inductance value.

3.4.8 MSF signal levels versus distance from transmitter

The NPL MSF signal stated characteristics are typically

ω = 2 π ·60 000 radians/s

at 100 km	E field greater than 10 mV/m
at 1 000 km	E field greater than 100 μV/m.

Square law estimates from MSF 1 000 km value give the following field strengths at example locations within the UK as approximate distances:

Edinburgh, Scotland	120 km	6 944 μV/m
Western side of Northern Ireland	320 km	976 μV/m
South Wales	400 km	625 μV/m
North of Scotland	400 km	625 μV/m
Central London	450 km	494 μV/m
Dover, south east England	520 km	370 μV/m
Channel Islands	650 km	237 μV/m
Shetland Islands	660 km	230 μV/m

3.4.9 Output signal levels for the antenna based on OEM datasheet

With the above assumptions and ignoring losses for the antenna not being correctly orientated for maximum signal the following typical output is obtained:

1 mV/m signal gives a sense coil output around 8 μV.

3.4.10 Required Integrated circuit tuner/detector module input

The OEM IC manufactures data sheet gives a working signal range of:

Minimum 0.4 μV to maximum of 20 mV under AGC control

Based on the quoted NPL signals, the following antenna output sense voltages are obtained, which would be well matched to the OEM combined integrated circuit amplifier/detector unit and provided by a unit less than 30 mm long by about 20 mm across maximum.

1 000 km	0.8 μV
100 km	80 μV
10 km	8 mV

3.5 Amplifier

Gain block with AGC and high impedance input.

Quoted Zin resistive differential minimum of 600 kΩ for German DCF77 at 77.5 kHz.

NB the overall gain can be split between the front end and the detector but results in the following.

System will have sufficient gain to lift the signal level to logic voltage levels at max gain (e.g. 5 V).

System maximum voltage gain estimate around 5 V/0.4 μV, i.e. 1.25 million – no AGC.

The AGC is capable of automatically setting the desired working level for a 20 mV max input.

System minimum voltage gain under AGC of 5 V/20 mV, i.e. 250 with maximum AGC action.

Implied AGC voltage range estimate of 1 250 000/250 = 5 000 to 1 (70 dB in power).

The first amplifier is only protected by the front end selectivity action of the antenna sense coil and its parallel tuning capacitor. This implies that whatever adjacent channel protection ratio is required for the unit to operate should be set mainly by the distortion/potential overload in this front-end part of the system.

Using a pure CW test signal at varying offset frequencies to measure relative selectivity while maintaining a constant AGC level indicated that the gain is distributed in two separate stages with a low gain stage front end prior to the narrow bandwidth crystal filter which then allows a dynamic range greater than 50 dB above a typical wanted signal without overloading.

The products used for the reception of MSF, WWVB and JJY keyed modulation signals in a domestic environment are all very similar and should with only variation is the centre tuning of the ferrite rod antennas and selecting the crystal filter centre frequency show similar performance in each case.

The integrated circuits in the detector are designed to look for and react to signal level amplitude changes so a CW interferer is the most benign case and any noise like impulsive amplitude variations in the interferer will result in significantly worse protection ratios.

Co-channel measurements comparisons between a pure continuous CW interferer and a Gaussian noise interferer with an MSF clock receiver indicated that a CW interferer could be up to 8 dB above the wanted signal to reach a complete failure to lock point, whereas for a Gaussian random noise signal the interferer caused lock failure at 18 dB below the wanted signal. This indicates a 26 dB greater sensitivity to noise-like amplitude varying signals than pure CW interferers.

The following narrow band filter and detection stages cannot remove any signal distortion products generated before it in the front end which fall into the narrow band filter. Further the AGC is derived after the narrow band filter and is unable to provide any significant mitigation for signals outside of that narrow wanted signal passband.

3.6 Narrow band filter

The actual signal modulation for MSF is on/off keying of the transmitter. The mark/space interval time slots are 100 ms. This leads to a maximum modulation frequency with a short interval with a single cycle in 200 ms, i.e. 5 Hz. The signal will therefore be a 100% modulated AM format with the carrier at 60 kHz and two sidebands when modulated with a modulation in frequency space reaching a maximum in amplitude with peaks at +5 Hz and −5 Hz either side of the carrier. There will be a $\sin(x)/x$ generic characteristic from the 100% amplitude modulation impressed on all the sub signals.

The overall transmission bandwidth is still very narrow and well suited to a classical quartz crystal series filter which at these frequencies has a narrow pass band characteristic in the tens of Hz region, for 3 dB bandwidth with a sharp cut off. The exact −3 dB bandwidth will depend very much on the quartz crystal parameters but also on the stray capacitances of the components around it.

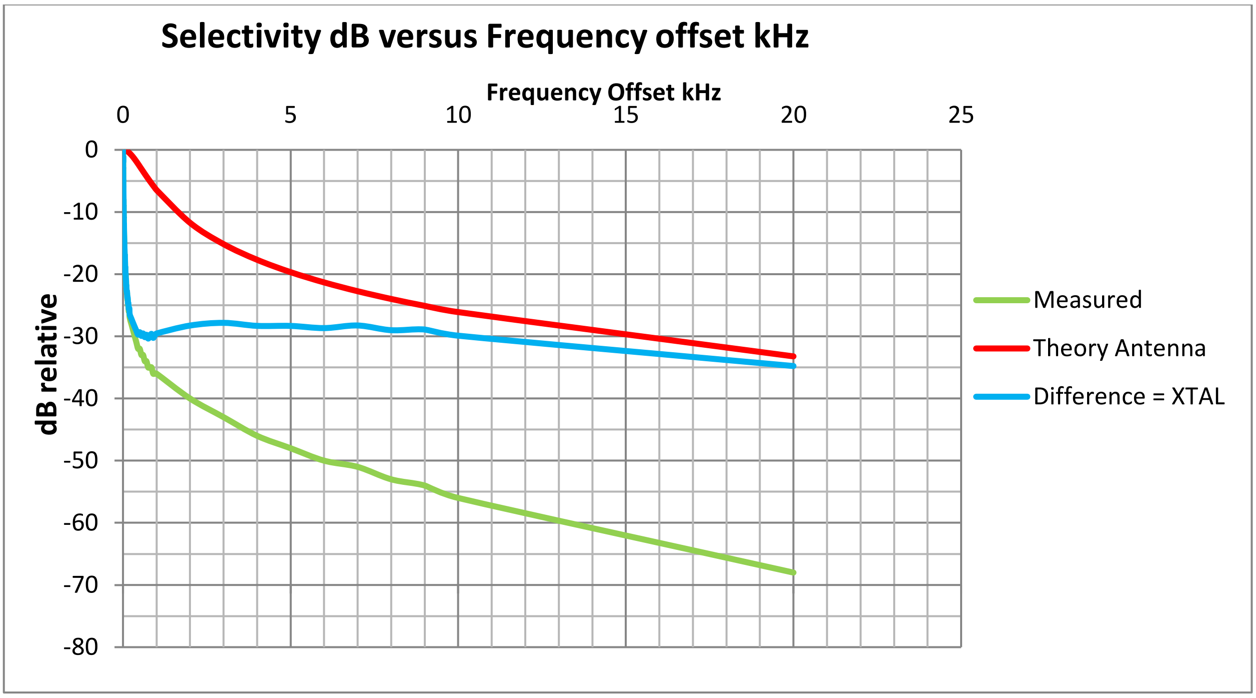
Out of passband attenuation for a single crystal in circuit is expected to be around 20 to 30 dB and achieved in less than 500 Hz from the centre frequency.

Measurements of relative selectivity using a CW input signal and maintaining a constant AGC voltage while varying the frequency offset were undertaken on one of the larger sized products where access was possible. This unit also had well separated internal component parts so leakage between circuitry was probably at a minimum compared to miniaturised products.

The results are shown in Fig. 6 for the overall relative selectivity versus frequency offset, the theory ferrite rod antenna selectivity alone and as the difference between them the potential characteristic of the crystal filter after the pre-amplifier antenna interface.

FIGURE 6

Relative selectivity versus frequency offset



For sharing study purposes a working figure of 25 dB suppression just for the crystal filter alone at greater than ±1 kHz from the wanted centre frequency could be used. This can be added to the attenuation from the antenna input tuning to estimate the overall characteristic of the receiver.

This narrow band filter will have a significant action in improving signal to thermal/atmospheric environmental noise under normal expected design working conditions and with the gain split as described above gives significant extra protection from out-of-band signals.

This will however not mitigate any issues with distortion and intermodulation products being generated in this passband by unwanted signals exceeding the dynamic range of the input stage which is somewhere in excess of 50 dB. It is to be expected that the first stage preamplifier front end antenna/amplifier pair will be the most open to unwanted adjacent channel signals causing issues.

3.7 Detector

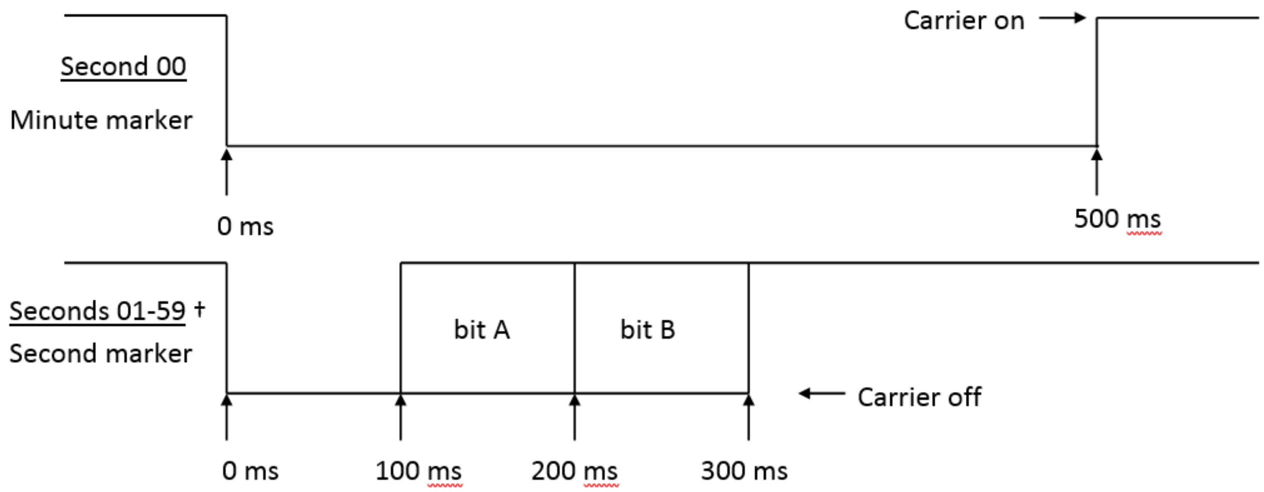
There is minimal information in any of the data sheets on the operating characteristics of the detector except the clue that there is a time lag to the logic level outputs following a carrier ON/OFF or vice versa transition. The time lag is in the tens of milliseconds so it is likely that there is some form of level detection with an integration time constant. There is a capacitor noted in the typical data sheet as being for the detector which would fit with this requirement. This would bring the detection mechanism into the classic configuration for analysis with simple smoothing which is built into the typical Gaussian signal to noise ratio analysis elsewhere in this Report.

Annex 1
to Section 3

National Physical Laboratory MSF transmission description

FIGURE 7

MSF 60 kHz time and date code



The two bits, numbered xxA and xxB for second xx bit polarity: 0 = on, 1 = off.

† Most minutes have 60 seconds numbered 00-59; exceptionally a UTC minute can have 61 seconds with the extra second being numbered 60 (a positive leap second), or 59 seconds (a negative leap second).

The MSF transmission from Anthorn (latitude 54° 55' N, longitude 3° 15' W) is the principal means of disseminating the UK national standards of time and frequency which are maintained by the National Physical Laboratory. The effective monopole radiated power is 15 kW and the antenna is substantially omnidirectional. The signal strength is greater than 10 mV/m at 100 km and greater than 100 µV/m at 1000 km from the transmitter. The signal is widely used in northern and western Europe. The carrier frequency is maintained at 60 kHz to within 2 parts in 10^{12}.

The MSF time and date code format is summarised in the diagrams above. Simple on-off carrier modulation is used, the rise and fall times of the carrier are determined by the combination of antenna and transmitter. The timing of these edges is governed by the seconds and minutes of Coordinated Universal Time (UTC), which is always within 0.9 seconds of Universal Time (UT1), an astronomical time scale based on the Earth’s rotation descended from Greenwich Mean Time (GMT). Every UTC second is marked by an OFF preceded by at least 500 ms of carrier, and this second marker is transmitted with an accuracy better than ±1 ms.

The first second of the minute begins with a period of 500 ms with the carrier off, to serve as a minute marker. The other 59 (or, exceptionally, 60 or 58) seconds of the minute always begin with at least 100 ms OFF and end with at least 700 ms of carrier. Seconds 01-16 carry information for the current minute about the difference (DUT1) between astronomical time and atomic time, and the remaining seconds convey the time and date code. The time and date code information is always given in terms of UK clock time and date, which is UTC in winter and UTC+1 h when Summer Time is in effect, and it relates to the minute following that in which it is transmitted.

The allocation of the signalling bits is detailed below and on the continuation sheet. Bits 17B-*51B inclusive, and bits 01A-*16A inclusive, are currently set to '0', but may be used in the future. Bits *52B and *59B are currently set at '0' but they may be used in the future.

Minute identifier

Bits *53A, *54A, *55A, *56A, *57A and *58A are all set permanently at '1', and are always preceded by bit *52A at '0', and followed by bit *59A at '0'. This sequence 01111110 never appears elsewhere in bit A, so it uniquely identifies the following second 00 minute marker.

*In minutes lengthened or shortened by a positive or negative leap second all these numbers are correspondingly increased or decreased by one (i.e. during these 61- or 59-second minutes the position of the time and date code is shifted by one second relative to the start of that minute).

DUT code

The difference UT1-UTC is known as DUT1 and is signalled to the nearest 100 ms in the range ±800 ms. A positive value means that UT1 is ahead of UTC. Bits 01B-16B are used to represent the DUT1 code in the following way, with bits not specified set to '0'.

TABLE 3

DUT1 (positive) code

DUT1	positive
0 ms	no bits set to '1'
+100 ms	bit 01B '1'
+200 ms	bits 01B and 02B '1'
+300 ms	bits 01B-03B inclusive '1'
+400 ms	bits 01B-04B inclusive '1'
+500 ms	bits 01B-05B inclusive '1'
+600 ms	bits 01B-06B inclusive '1'
+700 ms	bits 01B-07B inclusive '1'
+800 ms	bits 01B-08B inclusive '1'

TABLE 4

DUT1 (negative) code

DUT1	negative
0 ms	no bits set to '1'
−100 ms	bit 09B '1'
−200 ms	bits 09B and 10B '1'
−300 ms	bits 09B-11B inclusive '1'
−400 ms	bits 09B-12B inclusive '1'
−500 ms	bits 09B-13B inclusive '1'
−600 ms	bits 09B-14B inclusive '1'
−700 ms	bits 09B-15B inclusive '1'
−-800 ms	bits 09B-16B inclusive '1'

Time and date code

TABLE 5

BCD Year

Binary-coded-decimal year (00-99)							
80	40	20	10	8	4	2	1
*17A	*18A	*19A	*20A	*21A	*22A	*23A	*24A

TABLE 6

BCD date

BCD month (01-12)					BCD day-of-month (01-31)						Day-of-week †		
10	8	4	2	1	20	10	8	4	2	1	4	2	1
*25A	*26A	*27A	*28A	*29A	*30A	*31A	*32A	*33A	*34A	*35A	*36A	*37A	*38A

† 0=Sunday to 6=Saturday

TABLE 7

BCD time

BCD hour (00-23)						BCD minute (00-59)						
20	10	8	4	2	1	40	20	10	8	4	2	1
*39A	*40A	*41A	*42A	*43A	*44A	*45A	*46A	*47A	*48A	*49A	*50A	*51A

Parity bits

Bit *54B, taken with bits *17A-*24A inclusive, provides an odd number of bits set to '1'.

Bit *55B, " *25A-*35A inclusive, " " "

Bit *56B, " *36A-*38A inclusive, " " "

Bit *57B, " *39A-*51A inclusive, " " "

Summer time

When UK civil time is subject to a one-hour positive offset during part of the year, this period is indicated by setting bit *58B to '1'. Bit *53B is set to '1' during the 61 consecutive minutes immediately before a change, the last being minute 59, when bit *58B changes.

In the event of UK civil time undergoing an additional permanent offset, bit *58B will need to be changed once without any corresponding change in UK clock time.

Further information

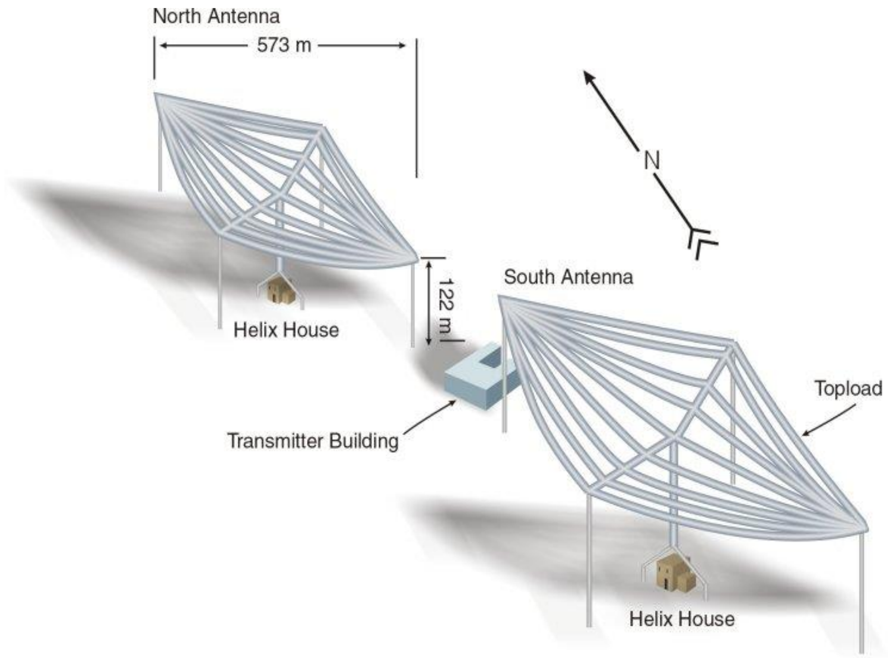
For further information please visit the website: https://www.npl.co.uk/msf-signal (valid at 2019-05-30).

4 Information on the US SFTS service

The Time and Frequency Division of the National Institute of Standards and Technology (NIST) operates a low frequency radio signal broadcast at 60 kHz from radio station WWVB located near Fort Collins, Colorado. The station went on the air on July 5, 1963, broadcasting a standard carrier frequency. On July 1, 1965 an amplitude modulated (AM) binary coded decimal (BCD) time code was added to the broadcast containing the day of year, hour, minute, second, and flags that indicate the status of leap years, leap seconds, and Daylight Saving Time. In 1999 the station power was increased to 50 kW and soon thereafter increased to 70 kW of radiated power. In 2006, the AM modulation depth was changed to improve reception in the presence of noise (discussed below), and in 2013 a phase modulation (PM) signal was added while maintaining the AM signal format. The addition of the PM code was to also improve reception in the presence of noise (discussed below). WWVB uses two nearly identical top-loaded dipole antennas consisting of four 122 m (400 ft.) towers arranged in a diamond shape (Fig. 8). The south antenna is located at 40^0 40’ 28.3” N / 105^0 02’ 39.5” W and the north antenna is located at 40^0 40’ 51.3” N / 105^0 03’ 00.0” W.

FIGURE 8

WWVB antenna

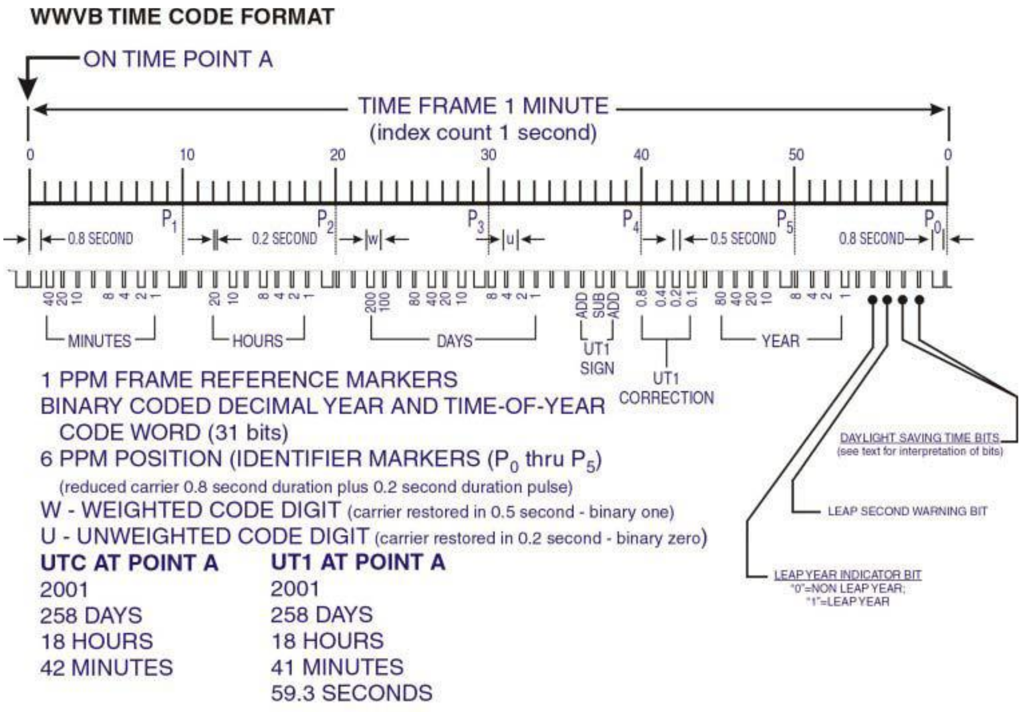


4.1 WWVB modulation and time code format

The WWVB AM time code is synchronized with the 60 kHz carrier and is broadcast continuously at a rate of 1 bit per second by use of pulse width modulation. The time code bits are produced by amplitude shift keying (ASK) of the carrier, where the carrier power is reduced at the beginning of each second and then restored to full power after a predefined interval. If full power is restored after an interval of 200 ms, it represents a binary ‘0’; if full power is restored after 500 ms, it represents a binary ‘1’. A frame marker is sent by holding the carrier low for 800 ms. A binary coded decimal (BCD) format is used, where four binary digits (bits) are combined to represent a decimal number. The complete AM time code format is shown in Fig. 9.

FIGURE 9

WWVB time code format



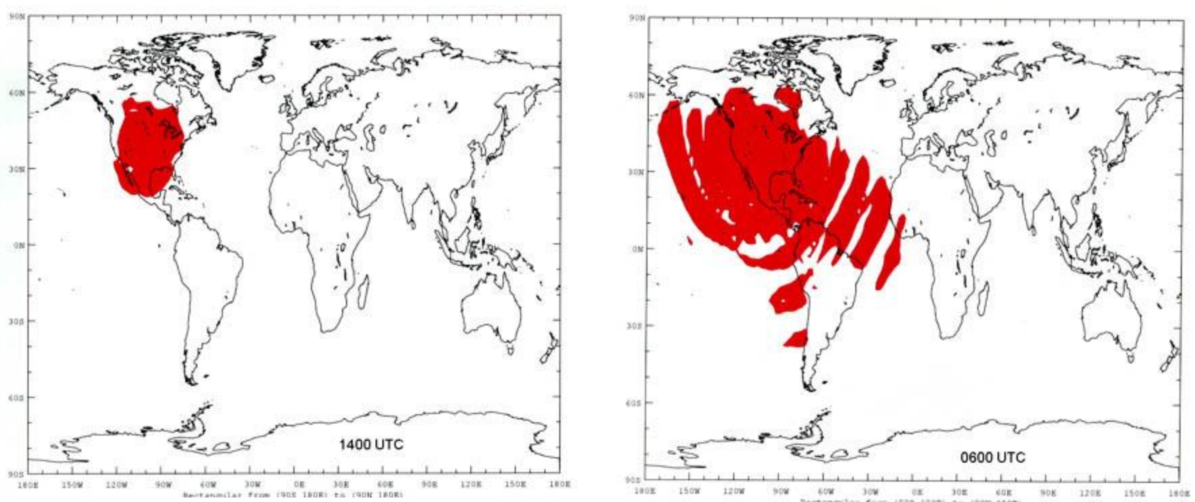
4.1.1 Coverage area for WWVB

Figure 10 shows the calculated day and night coverage areas of WWVB at a field intensity of 100 µV/m, which is barely sufficient for AM reception in most consumer-market devices. Such field intensity, when coupled with the gain of typical ferrite-rod based antennas that are used in common commercial radio-controlled clock (RCC) devices, results in an input signal level that is below 1 µV, which is on the order of the sensitivity level specified by vendors of receivers.

However, while this signal level may be sufficient when assuming only the natural thermal noise, whose bandwidth is typically limited within the receivers by a crystal with a bandwidth on the order of 10 Hz, additional interference may require a much higher signal level for proper reception.

FIGURE 10

Day-time and night-time coverage at 100 µV/m



4.2 AM signal analysis

The WWVB signal consists of three possible symbols transmitted at a rate of one symbol per second. The symbol for the binary ‘0’ is sent by reducing the carrier amplitude at the beginning of the second, holding it low for 200 ms, followed by 800 ms of full power. The symbol for the binary ‘1’ is sent by reducing the carrier amplitude at the beginning of the second, holding it low for 500 ms, followed by 500 ms of full power. Frame markers are sent by reducing the carrier amplitude at the beginning of the second, holding it low for 800 ms, followed by 200 ms of full power. A frame marker is transmitted at seconds 9, 19, 29, 39, 49 and 59 of each minute. The double frame marker at the seconds 59 and 0 marks the start of a minute. The symbol sent thus depends only on the signal level transmitted in two consecutive 300 ms time intervals within each second. This code is presented in Table 8. The code is equivalent to a binary, amplitude shift keyed (ASK) modulation signal. The relative performance of a binary ASK, matched filter receiver operating in a noise environment is a function of the size of the signal amplitude shift. A timing symbol is composed of two ASK binary symbols or bits. To successfully detect a timing symbol, the state for the two time-intervals (two bits) must be correctly detected. Therefore, the probability of a symbol being received in error is equal to the probability of making an error in the first time-interval or the second time-interval or both time-intervals. If the probability of an error in the first time-interval is $\boldsymbol{p}$, then the probability of an error in the second time-interval is also $\boldsymbol{p}$, since the modulation is identical in both time intervals. The probability of an error in both time intervals is $\boldsymbol{p}^2$. Thus, the total probability of making a symbol error is $\boldsymbol{p} + \boldsymbol{p} + \boldsymbol{p}^2$. It is reasonable to assume that the probability of a symbol error to be small is desired. Then, for $\boldsymbol{p} << 1$, $\boldsymbol{p}^2 << \boldsymbol{p}$ is given, so that the probability of a symbol error becomes $\boldsymbol{p} + \boldsymbol{p} = 2\,\boldsymbol{p}$. Now the probability of a time symbol error, $2\,\boldsymbol{p}$, in terms of the probability of a bit error, $\boldsymbol{p}$, for an Amplitude Shift Keyed binary modulation is given.

TABLE 8

Signal codes (original modulation depth)

		1st interval	**2nd interval**	
Symbol	0-200 ms	200-500 ms	500-800 ms	800-1 000 ms
‘0’	−10 dB	0 dB	0 dB	0 dB
‘1’	−10 dB	−10 dB	0 dB	0 dB
Frame marker	−10 dB	−10 dB	−10 dB	0 dB

4.2.1 Detection

One of two data symbols will be received. The ASK symbol ‘0’ is sent with signal amplitude $\boldsymbol{\chi_0}$. The ASK symbol ‘1’ is sent with signal amplitude of $\boldsymbol{\chi_1} = \mathbf{C}\,\boldsymbol{\chi_0}$. The value of $\mathbf{C}$ is given by:

$$\boldsymbol{C} = 10^{-X/20} \tag{2}$$

where $X = 20\log_{10}(\chi_0 - \chi_1)$ is equal to the amplitude shift in dB.

In the standard digital receiver[1] the symbol energy ($\boldsymbol{S}$) is integrated over a symbol duration. The received noise ($\boldsymbol{N}$) is also integrated over the same symbol duration. The resulting value, $\boldsymbol{S} + \boldsymbol{N}$, has a mean equal to the integrated signal plus a random noise term. Since there are two possible

[1] The standard digital receiver design is optimal for reception in additive, white, Gaussian noise (AWGN), which may not be the optimal receiver design for atmospheric noise at 60 kHz.

symbols that may be sent, the integrated symbol energy may have two possible values, S_0 and S_1. These values must differ by the transmitted amplitude shift of X dB, i.e.

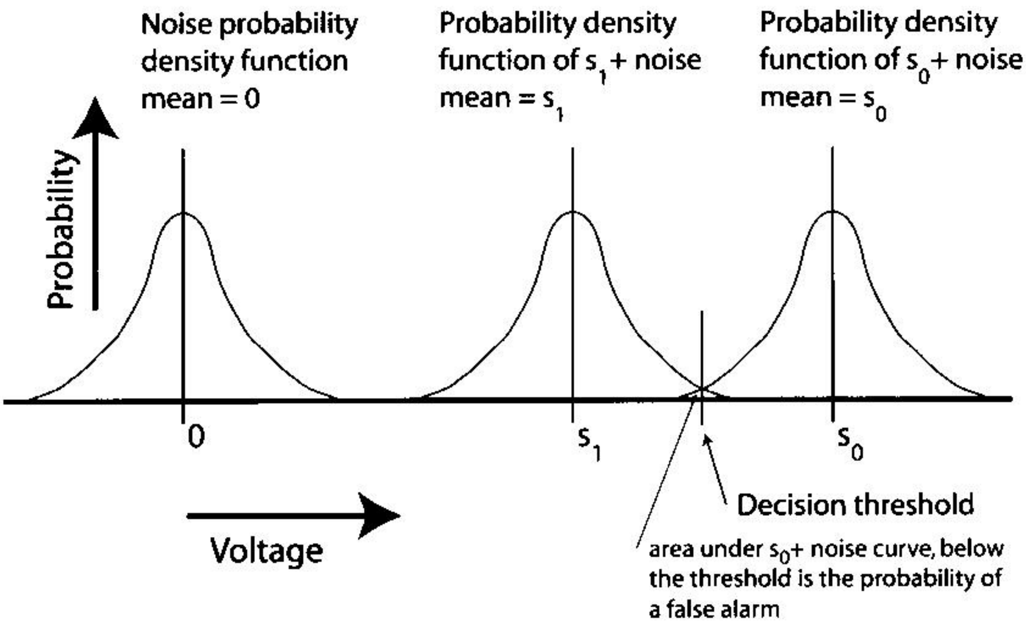
$$S1 = C\,S_0. \tag{3}$$

A value of $S_1 + N$ is developed when the low power level is received and a value of $S_0 + N$ is developed if the high-power level is received (Fig. 11).

At the receiver, a threshold is used to decide which symbol is received. When $S_0 + N$ is less than the threshold, a false detect occurs and a '1' is mistakenly received. When $S_1 + N$ is greater than the threshold, a false detect also occurs and a '0' is mistakenly received.

FIGURE 11

Diagram to determine which ASK bit was sent



Knowing that the minimum error rate occurs when the probability of false detects are equal, the threshold must be the same distance from S_0 and S_1. The separation between the values of S_0 and S_1 at the receiver determines the bit error rate (BER). It may be assumed that a separation distance of $\mathbf{y} = S_0 - S_1$ is required for a desired BER. It may now be solved for the signal level, S_0, necessary for S_0 and $\mathbf{S_1}$ to be separated by $\mathbf{y}$. Using equation (3) it can be seen that:

$$S_0 = \mathbf{y} / (1 - \mathbf{C}) \tag{4}$$

Now the dependence of this required received signal level S_0 on the amplitude shift X can be found. Having two amplitude shifts equal to X_a and X_b, then from equation (4):

$$S_0\,(X_a)/\,S_0\,(\mathbf{b}) = (1 - \mathbf{C\,b}) / (1 - \mathbf{C\,a}) \tag{5}$$

If 20 $\log_{10}$ of the ratios in equation (5) is used, the difference in dB between the received signal levels required to achieve the same bit error rate is obtained. Furthermore, let X_b be a reference shift amount, say infinity, which corresponds to keying the transmitter on and off. Then the signal level increase in dB required to match on-off keying is given by:

$$\mathbf{Sr}\,(X) = -20\,\log_{10}(1 - 10^{-X/20}). \tag{6}$$

where equation (2) has been used to substitute for **C**.

Interestingly, the result given in equation (6) does not depend on the noise level or the bit error rate. This means that the improvement in the performance of the receiver, when the amplitude shift, ***X***, is increased corresponds to a fixed increase in signal power over the entire area of reception. Comparing the original design modulation depth of 10 dB to on-off keying, or 100% modulation, gives an effective signal increase of 3.3 dB, i.e.:

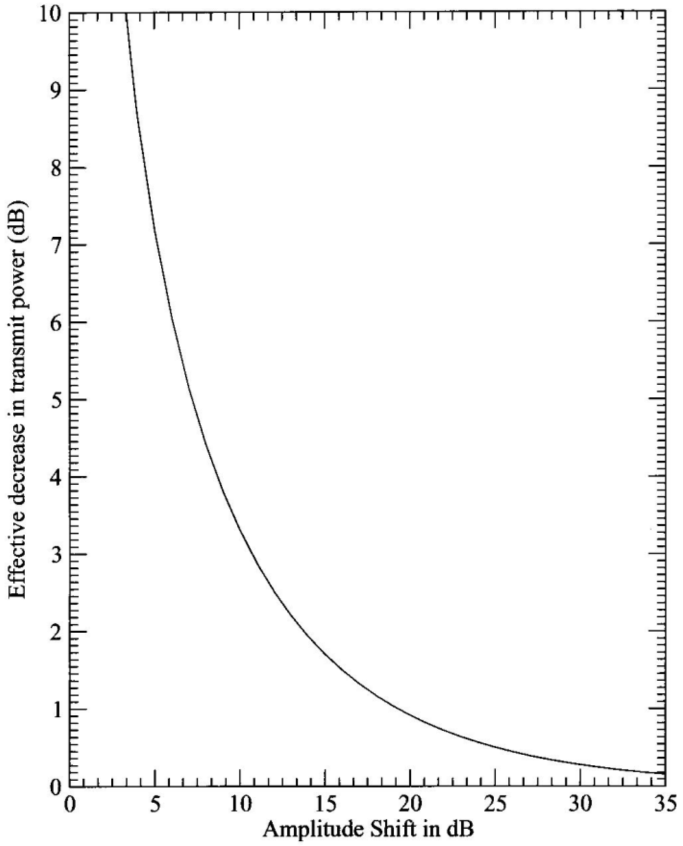
$$\boldsymbol{Sr}\,(10\text{ dB}) - \boldsymbol{Sr}\,(\infty\text{ dB}) = 3.3\text{ dB} - 0\text{ dB} = 3.3\text{ dB}$$

Of course, if 100% modulation depth were implemented there would be no low power carrier signal. The traceable frequency information would be lost during low power periods, and carrier phase tracking receivers would come unlocked.

Figure 12 shows the equivalent loss in signal level when an amplitude shift of ***X*** dB is used instead of on-off keying (100% modulation) for binary ASK modulation. A 10 dB modulation depth corresponds to 3.3 dB equivalent signal level loss, as shown previously.

FIGURE 12

The effective decrease in signal strength measured by a receiver from the optimum level of 100% modulation (on-off keying) to a modulation depth of *X* dB



4.2.2 Measurements

There is a limit to how much increase in modulation depth is acceptable. Too much reduction in power by amplitude modulation would limit the detection of the carrier frequency. A compromise is required between an increased modulation depth to improve the time-of-day code detection and the maximum low power level amplitude that still allows the continuous detection of the carrier frequency. By experimental tests of receivers located in the eastern United States (some 2000 km from the transmitter), a level of 17 dB modulation depth was determined to be acceptable. This level of modulation depth still allows continuous detection of the carrier frequency under normal propagation conditions. The effective increase of signal strength as measured by a receiver when the modulation depth is changed from 10 dB to 17 dB can be calculated by equation (6):

$$\boldsymbol{Sr}\,(10\text{ dB}) - \boldsymbol{Sr}\,(17\text{ dB}) = 3.3\text{ dB} - 1.3\text{ dB} = 2.0\text{ dB}.$$

Thus, standard digital receivers see the increase from 10 dB to 17 dB of modulation depth as if the transmitter has increased the effective radiated power by 2.0 dB. The radiated power of the WWVB broadcast is measured to be approximately 70 kW. Therefore, with this modification in the modulation format, the readability of the time code has improved to the equivalent of raising the output power to 122 kW. In 2006, the amplitude modulation depth was tested and officially changed from 10 dB to 17 dB.

4.3 WWVB phase modulation

In 2016, the WWVB format was augmented with the addition of a phase modulation (PM) code. This code was added to increase the reception capability for radio-controlled clocks (RCC) with PM capable receivers without affecting existing AM RCCs. The increase in reception potential is at least an order of magnitude better in the presence of additive, white, Gaussian noise (AWGN) with greater gains also made with efficient coding methods that reduces further errors in code detection.

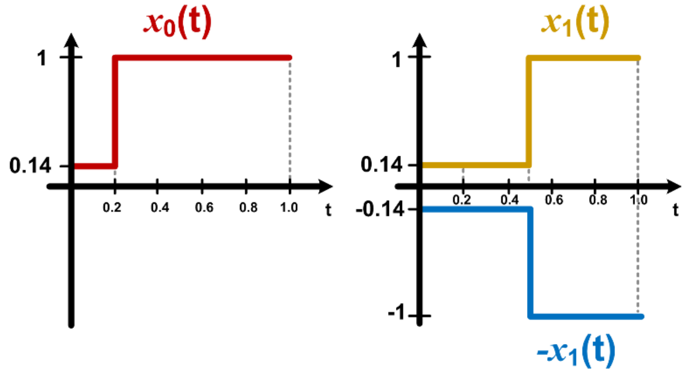
4.3.1 Binary phase shift keying modulation

The AM WWVB coding system transmits a pulse-width modulated amplitude-shift keyed waveform on a 60 kHz carrier. The 1-second reference markers are represented by a 17 dB reduction in power at the start of every second, for periods of 0.2 s and 0.5 s during the transmission of a ‘0’ and a ‘1’ respectively, as shown in Fig. 13, where the baseband waveforms for the transmitted bits ‘0’ and ‘1’ are denoted $\boldsymbol{X}_0(t)$ and $\boldsymbol{X}_1(t)$ respectively. The addition of a phase modulation to the signal as a binary-phase-shift-keying (BPSK) code, having a 180° difference in the carrier’s phase between the ‘0’ and ‘1’ symbols as represented by $-\boldsymbol{X}_1(t)$.

Hence, the modulated waveforms representing these symbols may be expressed as the products of the sinusoidal 60 kHz carrier and the baseband waveforms $\boldsymbol{S}_0(t) = \boldsymbol{X}_0(t)$ and $\boldsymbol{S}_1(t) = -\boldsymbol{X}_1(t)$, respectively, as shown in Fig. 13, where $\boldsymbol{X}_0(t)$ and $\boldsymbol{X}_1(t)$ denote the waveforms that are in use in the existing AM system. As can be seen in the figure, the phase modulation scheme is accomplished through simple sign inversion for the waveform representing the ‘1’ symbol. It should be noted that since the existing AM envelope-detector based receivers do not consider the carrier’s phase, they are not impacted by such modification.

FIGURE 13

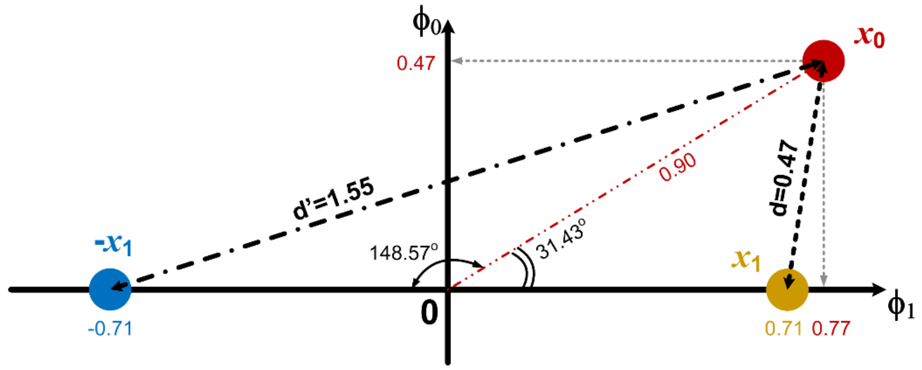
Baseband representation of the ‘0’ and ‘1’ bits in historical (AM) as well as PM modulations



A signal-space diagram for the two waveforms of the PM scheme versus those of the historical AM one is illustrated in Fig. 14. As can be seen in the diagram, the new pair of waveforms, $\boldsymbol{X}_0$ and $-\boldsymbol{X}_1$, having the same amount of energy (corresponding to their distances from origin), exhibit a much greater distance between the '0' and '1' symbols, allowing for more robust reception in the presence of additive noise. It is to be noted that the existing symbols are strongly correlated, i.e. have a very short distance between them in the signal space with respect to their energies.

FIGURE 14

The signal-space representation of the AM and PM WWVB '0' and '1' signals



Thus, by using phase modulation detection, an order of magnitude of improvement is achieved when assuming additive white Gaussian noise (AWGN). This analysis implicitly assumes that the receivers for both schemes would be optimal, i.e. based on correlation or matched filtering. In practice, the BPSK receiver may be implemented digitally in a near-optimal fashion, whereas the receivers for the AM scheme, not designed as a classical digital-communications system, are based on envelope detection. This adds an additional gap of 2-4 dB between the two, when only AWGN is considered. However, in the presence of on-frequency interference, the gain offered by realizing a near-optimal BPSK receiver may be arbitrarily higher, as is shown below. Furthermore, additional gains can be offered, as is shown in the following subs-sections, through encoding of the information, use of a known synchronization sequence, and extended-duration reception in the receiver.

4.3.2 Bit error rate analysis

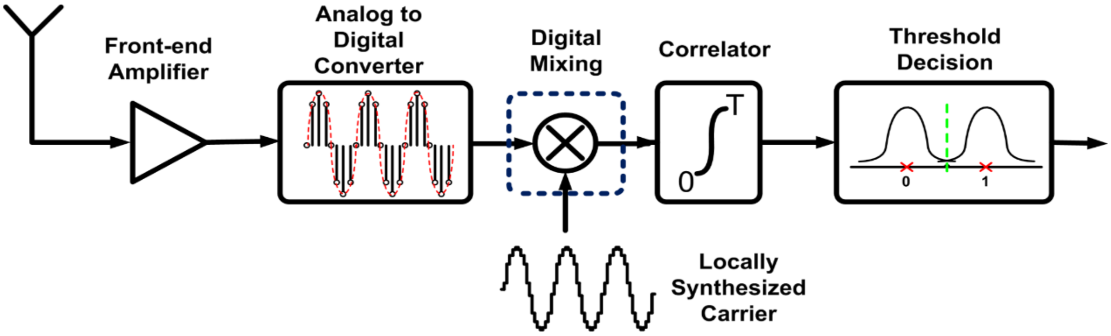
Figure 15 illustrates, in the form of a simplified block diagram, how a coherent BPSK optimal receiver may be implemented digitally. The filtering of the signal is based on the correlation operation, which is followed by a decision that is made in the presence of AWGN. The bit-error-rate (BER) performance of the receiver, for a signal to noise ratio E_b/N_o, is given by:

$$BER = Q\left(\sqrt{\frac{2 \cdot E_b}{N_o}}\right)$$

where E_b is the energy per bit and N_o is the noise density. The E_b/N_o is equivalent to the ratio between the power of the signal and the power of the noise in a bandwidth that is equal to the bit rate, i.e. E_b/N_o = SNR @ BW=R_b, where R_b represents the bit rate. The threshold decision block shown in the block diagram is where the decisions are made, and the errors occur, in direct relation to the variance of noise, which is shown to have Gaussian nature and equal variances around the '0' and '1' symbols.

FIGURE 15

A digital receiver architecture for BPSK



The BER may also be expressed as a function of the distance between the symbols in the signal space, as follows:

$$BER = Q\left(\sqrt{\frac{d^2}{2 \cdot N_o}}\right)$$

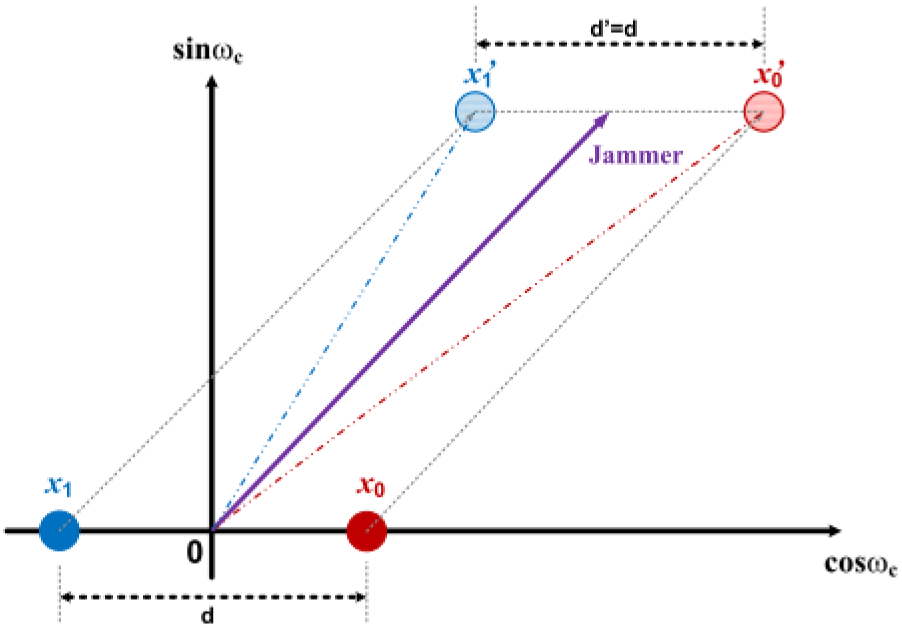
where $Q(x)$ is the tail probability of the normal distribution, i.e. $Q(x) = \frac{1}{\sqrt{2\pi}}\int_x^{\infty} exp\left(-\frac{u^2}{2}\right) du$.

The improvement obtained by the use of this phase modulation scheme assumes only the presence of AWGN in the receiver. In the presence of radio-frequency interference (RFI), such as the on-frequency signal experienced by WWVB AM receivers, mainly on the Eastern coast of the United States of America due to the British MSF broadcast, the performance improvement could be much more significant and stems from the structure of the BPSK receiver, where the demodulation is based on correlation.

Figure 16 shows a signal space wherein the constellation for antipodal BPSK is represented by the two signals $\boldsymbol{X}_0$ and $\boldsymbol{X}_1$ representing the symbols ‘0’ and ‘1’ respectively, having a distance $\boldsymbol{d}$ between them. Although an on-frequency jammer is an aggressor to the system and has no relevant information in it, it may be directly represented on this diagram, as it can be expressed as a linear combination of the functions spanning this space (orthogonal sinusoids centred at 60 kHz). The jammer, as shown in the diagram, can be added in vector form to each of the system’s two signals $\boldsymbol{X}_0$ and $\boldsymbol{X}_1$, resulting in the sums ‘***x***0’ and ‘***x***1’ respectively. The distance between these two sums in the signal space, denoted ***d*0**, is shown to be equal to ***d***, thereby allowing the receiver to operate without any performance degradation. In other words, the interfering signal only shifts signal constellation by a vector corresponding to the interferer’s level, without reducing the Euclidean distance between the ‘0’ and ‘1’, so long as the decision threshold is adjusted accordingly, and no non-linear effects are experienced in the receiver’s front-end, its performance would not be degraded at all by the additive on-frequency signal.

FIGURE 16

The effect of on-frequency RFI shown in the signal space as a vector diagram



4.4 Bit-designation in the transmitted frame

While the addition of phase modulation alone would allow for significant performance improvements in the presence of AWGN and RFI, as was shown in the previous sections, the system will benefit further from representing different information in the phase modulation than is given in its historical amplitude/pulse-width modulation. In order to maximize such benefits, the most common usages of the received signal were considered, with the following assumptions being made:

1. The received signal serves to convey the time of day (and date) to those devices who have not yet acquired it, such as a new wall-clock which has just had batteries put into it. In this scenario, the greatest amount of completely unknown data is assumed to be conveyed, for which the greatest receiver effort may be expected. Once time (and all other information) is acquired, the RCC device would have its own time-keeping capability and should not be required to repeat such acquisition. This information will be referred to as time information hereafter.

2. Periodically, devices that have already acquired the time-of-day would rely on the received signal to simply compensate for whatever time-drifts that may have experienced due to the inherent frequency error in their crystal-based oscillator (typically on the order of ±10 ppm). Naturally, this would be the most common use of the WWVB signal, since a device may require only a few acquisition operations in its lifetime but would regularly depend on the periodic time-adjustments based on the WWVB signal. This information will also be referred to as timing information hereafter.

3. Advance notification of the next daylight-saving time (DST) transition, i.e. either when entering or exiting DST, is advantageous, as it allows the receiving device to perform the one-hour time shift at the correct instance without having to receive the WWVB signal around the time of the transition. In devices that display the time and/or control systems in accordance with it (e.g. pool controllers, irrigation systems, heating/air-conditioning

controllers), it is obviously important for the correct time to be considered. However, devices that simply need to maintain synchronicity with one another may not need this. Other information that falls under this category is the advance notification of an imminent leap second. The DST schedule and leap second notification information are referred to as additional information hereafter.

For each of these three purposes (or types of information) listed above, based on its need and criticality, an efficient and robust way for representing the information was identified. This was done under the constraint of the same 60-second frame defined in the AM protocol, while also considering the marker symbols in it, which are of lesser energy, having a duration of only 0.2 s of high power. For this reason, the use of the markers has been avoided in conveying bits of information and were limited in their use to known components of a synchronization word, as shown in Table 9. While three of the seven markers are within the fixed synchronisation word, the four remaining marker symbols are reserved (denoted 'R' in Table 9) and their use is not defined here.

TABLE 9

Bit-designation in AM vs. PM WWVB one-minute frame

Second	**0**	**1**	**2**	**3**	**4**	**5**	**6**	**7**	**8**	**9**
Amplitude	Marker	Min[6]	Min[5]	Min[4]	0	Min[3]	Min[2]	Min[1]	Min[0]	Marker
Phase	sync[12]	sync[11]	sync[10]	sync[9]	sync[8]	sync[7]	sync[6]	sync[5]	sync[4]	sync[3]
Second	**10**	**11**	**12**	**13**	**14**	**15**	**16**	**17**	**18**	**19**
Amplitude	0	0	Hour[5]	Hour[4]	0	Hour[3]	Hour[2]	Hour[1]	Hour[0]	Marker
Phase	sync[2]	sync[1]	sync[0]	timepar[4]	timepar[3]	timepar[2]	timepar[1]	timepar[0]	hour[19]	R
Second	**20**	**21**	**22**	**23**	**24**	**25**	**26**	**27**	**28**	**29**
Amplitude	0	0	Day[9]	Day[8]	0	Day[7]	Day[6]	Day[5]	Day[4]	Marker
Phase	hour[18]	hour[17]	hour[16]	hour[15]	hour[14]	hour[13]	hour[12]	hour[11]	hour[10]	R
Second	**30**	**31**	**32**	**33**	**34**	**35**	**36**	**37**	**38**	**39**
Amplitude	Day[3]	Day[2]	Day[1]	Day[0]	0	0	DUTS[2]	DUTS[1]	DUTS[0]	Marker
Phase	hour[9]	hour[8]	hour[7]	hour[6]	hour[5]	hour[4]	hour[3]	hour[2]	hour[1]	R
Second	**40**	**41**	**42**	**43**	**44**	**45**	**46**	**47**	**48**	**49**
Amplitude	DUT[3]	DUT[2]	DUT[1]	DUT[0]	0	Year[7]	Year[6]	Year[5]	Year[4]	Marker
Phase	hour[0]	min[5]	min[4]	min[3]	min[2]	min[1]	min[0]	dston	leap	R
Second	**50**	**51**	**52**	**53**	**54**	**55**	**56**	**57**	**58**	**59**
Amplitude	Year[3]	Year[2]	Year[1]	Year[0]	0	LYI	LSW	DST[1]	DST[0]	Marker
Phase	dlpar[2]	dlpar[1]	dlpar[0]	dst[5]	dst[4]	dst[3]	dst[2]	dst[1]	dst[0]	sync[13]

4.4.1 Representation of timing information using a known sequence

Extracting timing from a digitally modulated received signal is best accomplished when a known sequence, having good autocorrelation properties, is embedded within it. This allows for a correlation operation in the receiver to reveal the timing of the received signal even in low SNR conditions, for which the recovery of individual bits within the sequence might have involved high error probabilities. The successful identification of the known sequence does not require the recovery of the individual bits comprising it, and directly corresponds to the total energy in the known sequence, which is proportional to its duration. Therefore, the duration of the known sequence in the frame was maximized, while weighing this against the need to send the time information in a robust fashion, i.e. with redundancy. As can be seen in Table 9, a total duration of 14 seconds is allocated to the known sequence, starting from the last second of a frame and ending 13 seconds into the next frame. Hence, the amount of energy invested in the timing information is on the order of a quarter of the total energy in a minute frame.

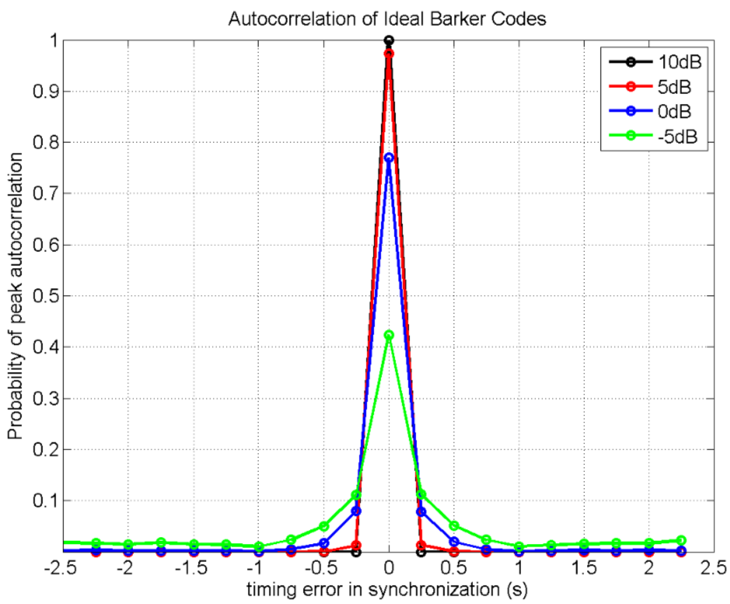
Since the timing adjustment operation is more frequent (assuming that the initial setting of the time may also be accomplished manually or by other means), it is useful to allow it to take place at SNR values that are below the minimum required for reliable acquisition. This justifies a long synchronization word.

If the correlation operation in the receiver were to be limited to when the signal is at high amplitude, then the total duration of useful correlation would be at least 7 seconds (the presence of a few 0.2 s AM markers in the synchronization word is compensated by the presence of 3 AM symbols that are fixed at zero, for which the high-amplitude duration is 0.8 s).

When the timing of the received signal is known to within a defined maximal timing error, the correlation operation may be limited in time accordingly, allowing the design of the known waveform to exhibit good autocorrelation throughout that entire correlation interval. However, when the timing is not known at all (e.g. when acquiring the time or when excessive drift has been experienced since the last successful timing adjustment), the correlation operation is more challenging, as it may be impractical to guarantee that the known sequence would exhibit minimal correlation with all possible values of the unknown data, to minimize the chances of incorrect identification. Figure 17 shows the probability of correct identification of a barker sequence of 11-bits, having good autocorrelation properties, for different SNR values, as measured in a bandwidth of a single bit. As can be seen in the figure, even for 0 dB, i.e. when the signal's power is equal to that of the noise in a 1 Hz bandwidth, the barker sequence would be identified to within 0.25 seconds of accuracy with over 90% probability (the sum of the three central results in the histogram). The probability of the synchronization resulting in an error of 1 second or more is negligible at such SNR (i.e. below 0.1%). The consequences of such a rare event can be avoided by considering a measure of confidence that may be generated in the receiver based on reception conditions.

FIGURE 17

Timing identification probability histogram for an example 11-bit barker code



4.4.2 Efficient representation of time information

Rather than allocating a separate field to each element of time and encoding this information in BCD, as is done in the historical AM protocol, where a total of 31 bits are consumed, the time may be represented more efficiently and robustly in one merged word as described below.

The entire time information, which includes the minute, hour, and date, is allocated a total of 26 bits, to which 5 redundant parity bits are added, which is, coincidently, the same number of bits dedicated to the representation of time in the AM protocol. Within the 26-bit time word, 6 bits are designated to the minute counter (0-59) and the remaining 20 bits represent the number of hours that have elapsed since the beginning of the century.

The number of hours in a century is limited to 100×365.25×24=876,600. Therefore, the 20-bit field is used efficiently, since log2(876600) = 19.74. The 6-bit portion representing the minute is also used efficiently, since log2(60) = 5.9, which is very close to 6 (60 of the possible 64 combination are used).

4.4.3 Linear coding of time word

The 26-bit time word is encoded into a 31-bit code-word using a Hamming systematic block code. This linear block code has the capability to correct a single error that may occur in any of the 31 bits, and the capability to detect up to two errors. If three errors were to occur, the received 31-bit word may appear like a legitimate one, resulting in erroneous decoding. Such scenario is considered intolerable and, therefore, the system is designed to minimize its probability. If, for example, the BER is on the order of 10^{-2} (1%), the probability for three errors is on the order of 10^{-6}.

The block code that was chosen is a systematic code, which means that the input data is embedded in the encoded output and may be read directly from there without necessitating decoding. This property does not come at the cost of performance, while it allows for simplified testing and reception (i.e. elimination of the decoding procedure, particularly for high SNR).

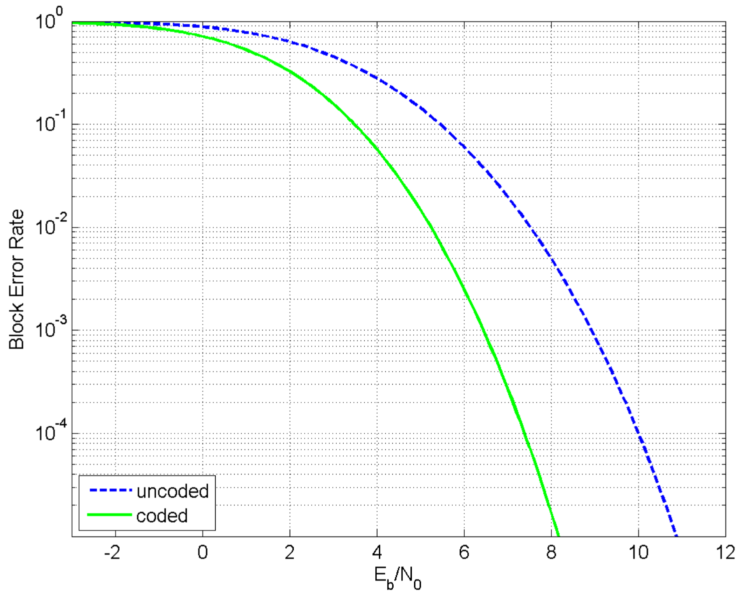
A block code may be denoted as a (n, k) code, where n and k denote the code-word size and the number of information bits respectively. For the time word $n = 31$ and $k = 26$. The equations below specify how each of the 5 parity bits, denoted "timepar[i]" (i=0, 1, 2, 3, 4), is to be calculated using the 26 bits comprising the minute counter (6 bits), and the hour counter (20 bits), which are jointly denoted "time[k]" (k=0, 1… 25).

- timepar[0]=sum(modulo 2){time[23, 21, 20, 17, 16, 15, 14, 13, 9, 8, 6, 5, 4, 2, 0]}
- timepar[1]=sum(modulo 2){time [24, 22, 21, 18, 17, 16, 15, 14, 10, 9, 7, 6, 5, 3, 1]}
- timepar[2]=sum(modulo 2){time [25, 23, 22, 19, 18, 17, 16, 15, 11, 10, 8, 7, 6, 4, 2]}
- timepar[3]=sum(modulo 2){time [24, 21, 19, 18, 15, 14, 13, 12, 11, 7, 6, 4, 3, 2, 0]}
- timepar[4]=sum(modulo 2){time [25, 22, 20, 19, 16, 15, 14, 13, 12, 8, 7, 5, 4, 3, 1]}

Syndrome based decoding, wherein a syndrome vector is calculated based on the received word in a linear fashion, may be used in the receiver. A non-zero syndrome indicates that at least one error occurred in the received word. For this Hamming code, the syndrome can correctly indicate the error location if only one error occurs in the received word. In order to guarantee the reliability of the recovered information, the receiver does not necessarily have to correct the received word when the syndrome is non-zero. Instead, a second reception, confirming the contents of the first, may be used, resulting in increased reliability at the cost of delayed acquisition. Hence, the error detection capability may be considered more important than the correction capability.

FIGURE 18

Word error rate comparison for the coded and uncoded time word



Since it is probable to have more than one error in the received word in low SNR scenarios, an erasure may occur, and the receiver will make a second acquisition attempt. In contrast, a correction can be made in high SNR scenario, where the likelihood of having two or more errors is very low.

Whenever at least one bit in a word is recovered in error despite the coding, this represents a word-error event, for which the word-error-probability (WER) is defined. The WER comparison for the coded and uncoded time-word is shown in Fig. 18. It should be noted that a WER of 10^{-3} corresponds to one error in 100 years even if time acquisition is performed 10 times in a year. The demodulation SNR for this WER is shown to be about 8.9 dB for the uncoded word and 6.4 dB for the coded word representing a coding gain of 2.5 dB.

4.4.4 Encoding of the additional information fields

The 7-bit additional information field comprises one bit indicating whether DST is in effect or not, one bit indicating whether a leap-second is to be added at the end of the current half-year, and a 5 bit DST schedule word, which serves to notify of the time and day for the next DST transition. If DST is in effect (e.g. in July), then the interpretation of the 5-bit word should refer to when it is to end. If DST is not in effect (e.g. in December), the interpretation should refer to when it is to start again. The start date and end date options are listed in Table 11 below. A total of eight specific options are supported for each, and an "out of range" possibility is defined, in case the DST schedule is changed in the future to a time that is not within those covered by the Table. Additional options are defined to allow for DST to be implemented permanently or to be cancelled altogether. With three possible values for the time at which the DST transition is to occur (1AM, 2AM, or 3AM) and a fourth option to be used for special messages, as shown in Table 10, a total of 32 combinations may exist for the DST 5-bit schedule word.

TABLE 10

Designation of bits 3-4 in 5-bit word for DST schedule

bit 4	bit 3	significance
0	0	use column A to decode bits 2:0 (special messages)
0	1	next DST transition hour is 1AM, day is in bits 2:0
1	0	next DST transition hour is 2AM, day is in bits 2:0
1	1	next DST transition hour is 3AM, day is in bits 2:0

TABLE 11

Designation of bits 0-2 in 5-bit word for DST schedule

bit 2	bit 1	bit 0	column A	column B (end of DST)	column C (start of DST)
0	0	0	DST permanently off	3rd Sunday before “O”	6th Sunday since “M”[2]
0	0	1	DST permanently on	2nd Sunday before “O”	7th Sunday since “M”
0	1	0	DST out of range	1st Sunday before “O”	8th Sunday since “M”
0	1	1	reserved	last Sunday of Oct. = “O”	1st Sunday of March = “M”
1	0	0	reserved	1st Sunday of Nov.	2nd Sunday since “M”
1	0	1	reserved	2nd Sunday of Nov.	3rd Sunday since “M”
1	1	0	reserved	3rd Sunday of Nov.	4th Sunday since “M”
1	1	1	reserved	4th Sunday of Nov.	5th Sunday since “M”[2]

The Sundays in column C, indicating the start date of DST, are not in chronological order, since it was advantageous to designate the same word to the first Sunday of November and to the second Sunday of March, being the current end and start dates respectively. This allows for more efficient representation of the information under the assumption that this DST schedule, which is currently in use, will likely remain the schedule for many years to come. The other optional schedules were defined to allow some margin around what appears to be a possible schedule that would be instated in the future. It is to be noted that the last Sunday in October or in March may be either the fourth or the fifth Sunday of that month.

The 2-bit word comprising the DST status bit, and the leap-second notification bit may be used immediately upon reception and is of high importance. These bits are also somewhat unpredictable, and, therefore, have high information content. Hence, this 2-bit word is encoded into 5 bits using a shortened Hamming systematic code that provides relatively high robustness, as detailed in a following section.

In contrast, due to the highly disparate a priori probability for the DST transition schedule options, a non-linear code is used to encode the 5-bit DST schedule word into a 6-bit code-word, offering non-uniform distancing for the various code-words, with the most probable one having the highest protection, i.e. the greatest Hamming distance from all other code-words.

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2 The first Sunday in April could be either the fifth or sixth Sunday since the beginning of March.

4.4.5 Linear coding for DST and leap second indicators

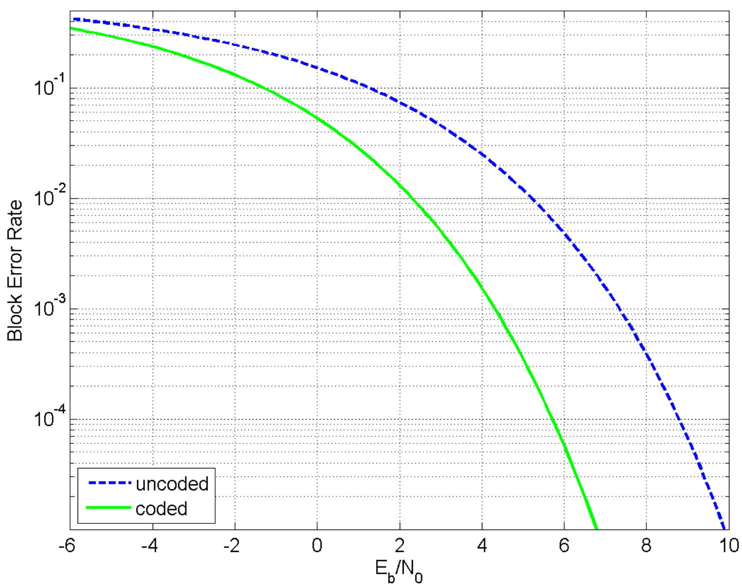
The (5,2) shortened Hamming systematic code that is used to encode these two information bits into a 5-bit code-word is derived from a (7,4) Hamming systematic code. The equations below specify how each of the three parity bits, denoted "dlpar[i]" (i=0, 1, 2), is to be calculated using the two input bits, which are denoted dston and leap.

- dlpar[0] = dston
- dlpar[1] = sum(modulo 2){leap, dston}
- dlpar[2] = leap

Figure 19 shows the performance gain obtained through the use of this linear code. As can be seen there, at WER=10^{-3}, corresponding to one error in 100 years if time acquisition is performed 10 times in a year, the demodulation SNR is shown to be about 7.3 dB for the uncoded word and 4.3 dB for the coded word, representing a coding gain of 3 dB.

FIGURE 19

Word error rate comparison for the coded and uncoded DST and leap second indicators



4.4.6 Non-linear coding of DST schedule word

The non-linear code for the advance notification of the next DST transition is designed such that one code-word will have maximum protection, corresponding to a maximal minimum Hamming distance, denoted d_{min}. The remaining code-words were selected to have a maximal number of codewords of maximum d_{min}. Table 12 shows the code-words and their d_{min}. The first codeword, having a maximum d_{min} of 3, is mapped to the most probable DST schedule, which is the one established most recently (i.e. the DST period on the 2nd Sunday of March and ending on the 1st Sunday of November, as shown in green in Table 12). Since the transition will most likely remain at 2AM and will be implemented on the same Sundays that it has been recently moved to, this combination was selected as the most probable one, for which the maximal coding protection was assigned in the non-linear code.

The schedule shown in red in Table 12 represents the one previously in use, which is assumed to have the second highest probability. Hence, these words have been designated code-words with the second highest minimum distance of 2, when coded into the 6-bit code-word.

TABLE 12

The 6-bit DST non-linear code-word representing the advance notification for the DST transition

codeword index	input word of 5 bits					output (coded) word of 6 bits						dmin
	bit 4	bit 3	bit 2	bit 1	bit 0	bit 5	bit 4	bit 3	bit 2	bit 1	bit 0	
0	0	0	0	0	0	0	0	0	0	0	0	**3**
1	1	0	0	0	0	1	1	0	0	0	1	2
2	0	1	0	0	0	1	0	1	0	0	1	2
3	1	1	0	0	0	0	1	1	0	0	1	2
4	0	0	1	0	0	1	0	0	1	0	1	2
5	1	0	1	0	0	0	1	0	1	0	1	2
6	0	1	1	0	0	0	0	1	1	0	1	2
7	1	1	1	0	0	1	0	0	0	1	1	2
8	0	0	0	1	0	0	1	0	0	1	1	2
9	1	0	0	1	0	0	0	1	0	1	1	2
10	0	1	0	1	0	0	0	0	1	1	1	2
11	1	1	0	1	0	1	1	1	0	0	0	1
12	0	0	1	1	0	1	1	0	1	0	0	1
13	1	0	1	1	0	1	0	1	1	0	0	1
14	0	1	1	1	0	0	1	1	1	0	0	1
15	1	1	1	1	0	1	1	0	0	1	0	1
16	0	0	0	0	1	1	0	1	0	1	0	1
17	1	0	0	0	1	0	1	1	0	1	0	1
18	0	1	0	0	1	1	0	0	1	1	0	1
19	1	1	0	0	1	0	1	0	1	1	0	1
20	0	0	1	0	1	0	0	1	1	1	0	1
21	1	0	1	0	1	1	1	1	1	1	0	1
22	0	1	1	0	1	1	1	1	1	0	1	1
23	1	1	1	0	1	1	1	1	0	1	1	1
24	0	0	0	1	1	1	1	0	1	1	1	1
25	1	0	0	1	1	1	0	1	1	1	1	1
26	0	1	0	1	1	0	1	1	1	1	1	1
27	1	1	0	1	1	1	1	1	1	0	0	1
28	0	0	1	1	1	1	1	1	0	1	0	1
29	1	0	1	1	1	1	1	0	1	1	0	1
30	0	1	1	1	1	1	0	1	1	1	0	1
31	1	1	1	1	1	0	1	1	1	1	0	1

Since the DST schedule word is followed by the synchronization sequence, it is desirable to let the most probable DST schedule word, in conjunction with the synchronization sequence, have good autocorrelation properties. Therefore, an offset word C will be added to all code-words in Table 12 in order to improve the synchronization performance while maintaining the minimum Hamming distance of the codewords.

The use of efficient coding and sequencing in the PM format increases reception capability more than 5 dB in the presence of AWGN as well as on-frequency interference, thus improving operation in low SNR conditions. The WWVB PM time-code has been successfully received in the southern parts of South America and as far away as Capetown, South Africa.

4.5 Reception of WWVB in the presence of WPT operating on-frequency

The vast majority of commercial and civilian radio receive clocks (RCC) in the U.S. decode the AM portion of the WWVB broadcast. Therefore, an analysis of an on-frequency WPT generator and how it would affect reception of these devises is required to establish a worst-case scenario base line. As shown below, a WPT that operates on-frequency (or at a harmonic/sub-harmonic) would not only add energy at the baseband reducing the *S*/*N* but the added coherence energy (in phase/out of phase) would degrade the detection capability of a typical threshold detector system.

4.5.1 AM envelope-detector-based receiver

Figure 20 provides a simplified block diagram for an envelope-detector-based receiver of this type. As can be seen in this block diagram, the AM signal is converted into an analogue equivalent baseband signal by use of a conventional nonlinear envelope detector (similar to the diode-based circuit in traditional AM receivers). A threshold operation that follows serves to determine the middle level, around which the voltages below it would be converted to a logical low level and the voltages above it to a logical high level. The digital processing stage that follows this operation measures the pulse durations and decides on the recovered symbols (“1”, “0”, or “marker”). Naturally, with such receiver topology, an on-frequency interferer that is present while the WWVB victim signal in the receiver is at its “low” state can cause the receiver to decode that symbol incorrectly. In particular, the “marker” symbols in WWVB are represented by 0.8 s of a low-level in the carrier (−17 dB) followed by 0.2 s at the high level, and are used to indicate the timing of the minute frame and the fields within it.

FIGURE 20

Existing receiver architecture

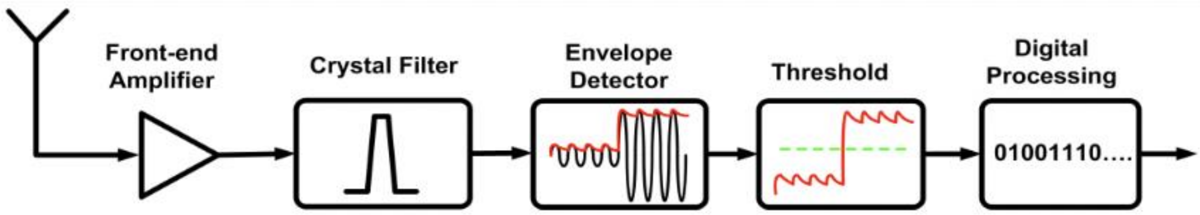
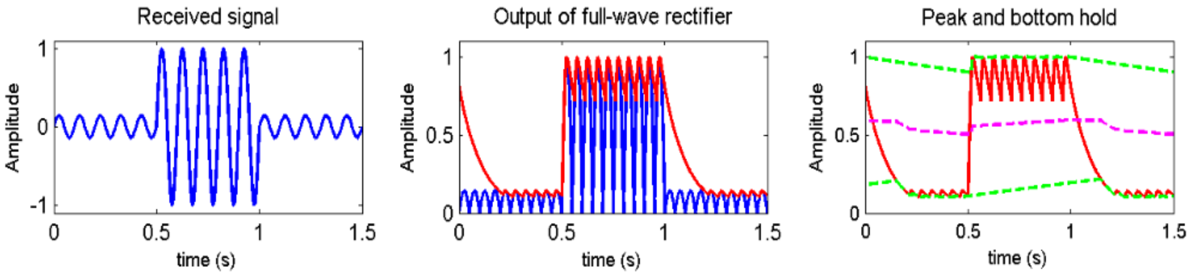


Figure 21 illustrates the principle of operation of a typical envelope-detection-based receiver. The modulated signal on the left, which is the input to the receiver, is shown in the absence of additive noise, and has two different amplitude levels, with the information represented in the durations of each of these levels. The modulation rate and carrier are not to scale, for the sake of visual clarity, which is why the decay time of the envelope detector in the middle figure appears too long with the respect to the pulse duration. The figure on the right illustrates how the high/low decision is made, by following the “low” and “high” levels with dedicated peak holders (with appropriate time-constants) and deriving the middle (average) of these two. A threshold operation (i.e. simple comparator) is then used to create the logic level signals for the digital stage that follows, where the pulse durations are measured and the “1”/ “0”/ “marker” symbol decision is made.

FIGURE 21

The analogue processing of the AM signal in a typical WWVB receiver

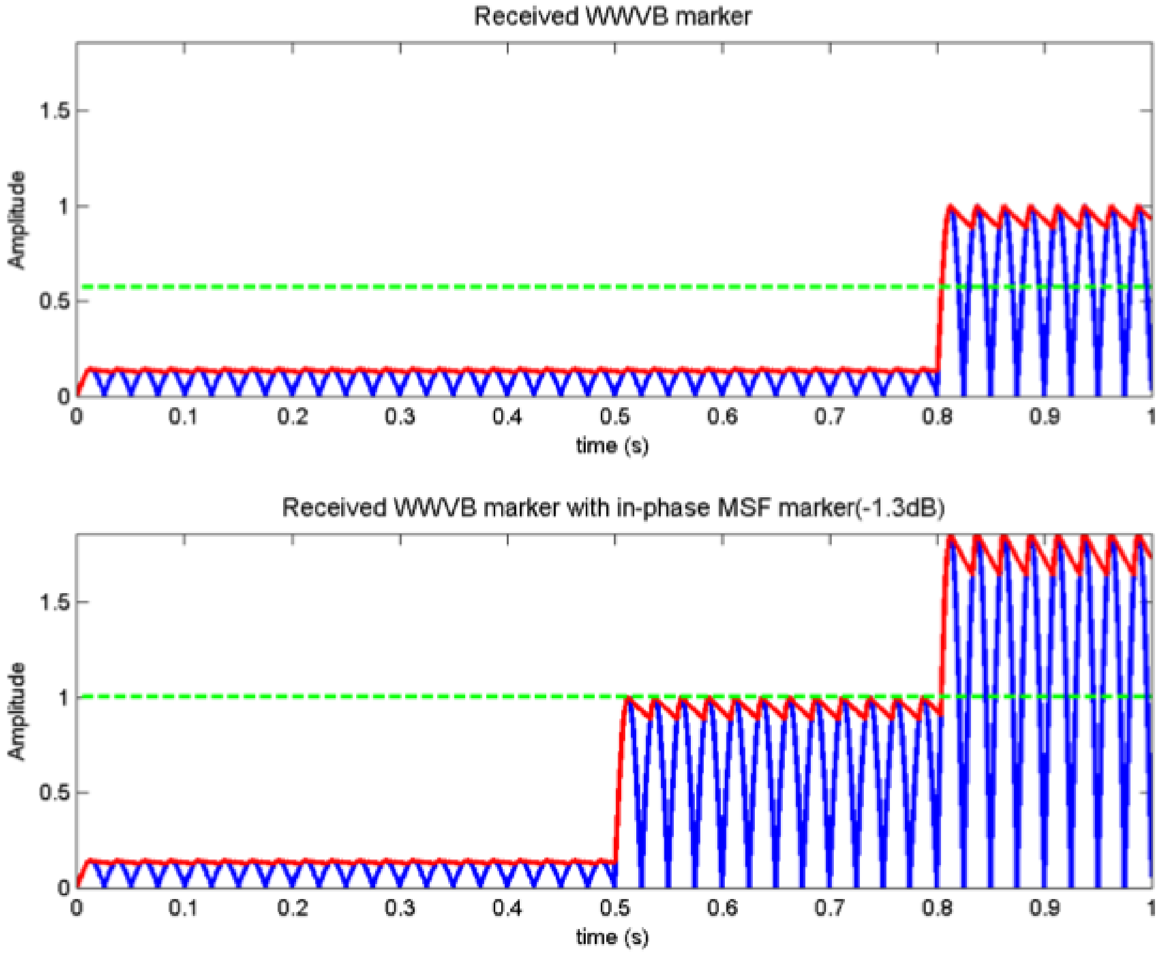


4.5.2 Detection limits for an on-frequency jammer

Figure 22 illustrates how the threshold operation can be misled by the presence of a jammer having the nature of an on-frequency WPT signal. With the high and low amplitude levels in the WWVB signal having a ratio of 1/7 (−17 dB), the lower level is shown to be at 0.14 (with the high level normalized to 1), and the threshold is set to the middle value of (1 + 0.14) / 2 = 0.57. The WWVB signal shown in the top figure is a marker, having a low level of 0.8 s followed by a high level of 0.2 s. The additive in-phase WPT signal shown in the bottom figure, for the first marker in a minute, has a low level of 0 for the first 0.5 s of the second and a high level assumed to be 86% (−1.3 dB) relative to the WWVB signal. The three different sum results shown for 0<t<0.2 s, 0.2 s<t<0.5 s, and 0.5 s<t<0.8 s are shown to be 0.14, 1, and 1.86 respectively. The maximum and minimum peak holders for these three levels would produce a middle level that is around (1.86 + 0.14) / 2 = 1, for which the signal may be high for t > 0.5 s. Consequently, even in the absence of noise, the threshold decision operation will fail for that interval and the digital stage that follows would be decoding this symbol as a “1” instead of “marker”.

FIGURE 22

The effect of an in-phase WPT interferer with 86% relative amplitude (−1.3 dB)



Even in the absence of additive noise (i.e. very high SNR), with an interfering WPT signal that is 1.3 dB below the WWVB received signal or stronger, an envelope-detector-based receiver would fail to correctly decode the received WWVB signal. For the other extreme case of out-of-phase interference (180^0 phase difference), the effect of the interferer is worse, resulting in a maximal tolerable interfering level that is about 6.5 dB below the received WWVB signal level. Between these two phase relationships, the tolerable level of the interferer is between these two values. It is expected that the phase and frequency stability of the WPT will be less than that of WWVB so that interference levels will fluctuate between these values as the phase angle changes.

It is to be noted that when the SNR is not very high, a receiver's performance may be noticeably degraded by the presence of this jammer, as it effectively reduces the noise margin in its decision function. The peak-detector based receivers typically determine the "low" and "high" levels of the received carrier and adaptively set a threshold between them, as shown in Fig. 21. However, when a sufficiently strong on-frequency jamming signal is present during the time when the WWVB signal is at its low amplitude state, the receiver may incorrectly establish that the received signal is at its high state, and would consequently not measure the pulse durations correctly, where the information is encoded.

A WPT signal could represents such a jammer if it operated at the same frequency (or at a harmonic or subharmonic frequency) of WWVB, and sets a worst-case scenario for on-frequency WPT operation. Besides, the degradation of the *S*/*N* of an on-frequency generator an additional 6.5 dB of protection criteria is required to accommodate the threshold/decoding technique. This degradation would apply to other National Physical Laboratories' LF broadcast services that use AM envelope detection RCCs. For instance, if a WPT operated at 80 kHz and generated power at the subharmonic of 40 kHz, the degradation of the *S*/*N* due to the additive power/noise at 40 kHz would require an addition 6.5dB of protection due to threshold decoding issues.

4.6 Reception of WWVB in the presence of WPT operating off-frequency

The proposed frequency band for WPT is 79 to 90 kHz and is not coherent with the WWVB operational frequency of 60 kHz. Therefore, the interference to WWVB RCCs will be of an additive noise nature and is dependent on the out-of-band power generation at 60 kHz of the WPT device. The further the frequency of operation of the WPT from 60 kHz the lower the out-of-band power is expected.

4.6.1 Additive noise tolerance for WWVB AM reception

A *S*/*N* level of 14-18 dB is required for adequate reception and decoding of the WWVB signal using AM envelope-detection-based-receivers. As shown in Fig. 20, an approximate field strength of 100 µV/m is expected at the extremities of the continental U.S. over a diurnal. Therefore, a field strength of 1.6 µV/m of additive noise/power could interfere with the operation of an RRC at far-field. The protection criteria of distance from a WPT to an RCC should be limited to a received power level of 1.6 µV/m or less.

5 Information on the German SFTS service

5.1 Description of the German SFTS service DCF77

The Physikalisch-Technische Bundesanstalt, Braunschweig, Germany, is entrusted by the "Law on Units and Time" from 2008 to realize and disseminate legal time for Germany. In this context, DCF77 disseminates standard frequency and time-of-day information as an infrastructure service of the federal government of Germany. The standard frequency of 77.5 kHz and coded time information are both generated at the transmitter site by PTB equipment are currently broadcast via facilities operated by Media Broadcast GmbH under contract. The transmitter DCF77 is located at the site of the radio station Mainflingen (coordinates: 50°01' north, 09°00' east), approximately 25 km south-east of Frankfurt am Main. A detailed description of the DCF77 including its history was published in A. Bauch, P. Hetzel, D. Piester, 2010, "Time and Frequency Dissemination with DCF77: From 1959 to 2009 and beyond," PTB-Mitteilungen, Special Issue Vol. 119, No. 3, 3-26. The information provided in this Report is limited to details relevant for the proper reception of the signal and leave out all aspects of time and frequency accuracy.

The transmitted signal is generated from three independent atomic clocks as inputs to three time-code generators, from which one is chosen as the master source and one as the backup. In regular operation, the phases of the carriers of all outputs are kept in mutual agreement within a few tenths of a microsecond. The content of the DCF77 programme comprises traceable standard frequency as derived from the carrier phase (CP), AM SI second markers at the beginning of every second and an AM time and date code message. The additional PM is phase coherent to the beginning of the second and the same time and date information as encoded in the AM is provided. Whereas AM is widely used for applications with uncertainty requirements not below 1 ms, the PM reception allows one to refer clocks to UTC(PTB) at the level of 10 µs. The DCF77 AM and PM are transmitted at different times of each second of the DCF77 signal, so that a receiver can detect either or both modulations: AM uses only the amplitude, between 0 s and 0.2 s after the beginning of each second, whereas PM effects only the carrier phase and is transmitted during the remaining 0.8 s of each second interval.

5.1.1 Amplitude Modulation (AM)

The 77.5 kHz carrier of DCF77 is amplitude modulated with second marks. At the beginning of each second (except for the last second of each minute, which serves as identification for the beginning of the next minute) the amplitude is reduced phase-synchronously with the carrier oscillation for a duration of 0.1 s or 0.2 s to a residual level of 15 %. In Fig. 23, the falling edge of the envelopes of the carrier oscillation emitted by DCF77 (curve a) at the beginning of a second mark and the associated control signal (curve a’) are shown. The blanking interval of 250 µs in the drive signal causes a faster decay of the antenna circuit, so that the obtained decay rate is practically identical with that which would be obtained for a drive signal completely without residual amplitude. For comparison, the dashed line b illustrates the steepness that would be obtained if the drive signal was reduced directly to the residual amplitude without blanking interval (curve b'). The steeper the falling edge, the more exact is the determination of the beginning of the carrier reduction which is defined as the beginning of the second of the disseminated legal time.

The different durations of the second marks serve for the binary encoding of time and date: second marks with a duration of 0.1 s correspond to the binary zero, and marks with a duration of 0.2 s to the binary one. Once during each minute, the numbers of the minute, the hour, the day, the day of the week, the month and the year are transmitted using BCD coding (BCD: Binary Coded Decimal, every digit of a number is encoded separately). From the calendar year, only the unit place and the decimal place are transmitted, i.e. the year 2018 is transmitted only as 18. The emitted code contains the information for the minute that follows. The temporal sequence of the bits and their significance are explained by the encoding scheme shown in Fig. 24.

FIGURE 24

Falling edge of the carrier envelopes emitted by DCF77 at the beginning of a second mark

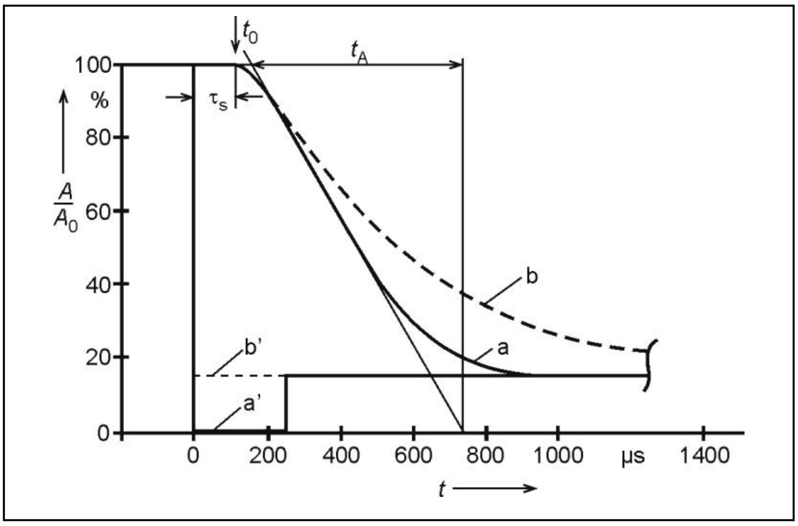
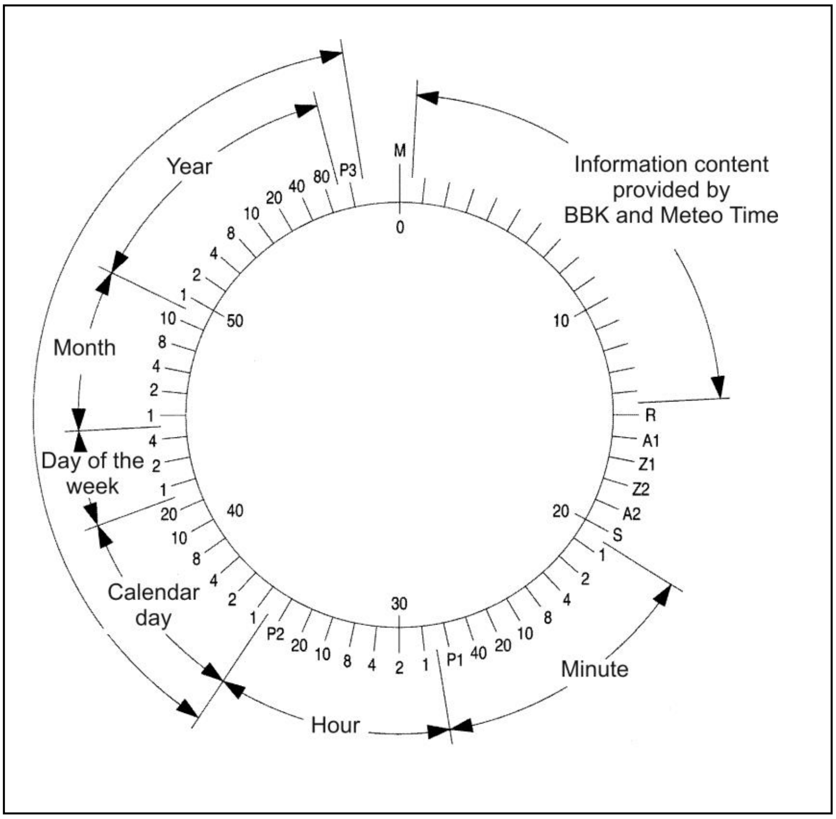


FIGURE 25

Coding scheme of the DCF77 time code frame



The designations in Fig. 25 are as follows and have inter alia been reported in Recommendation ITU-R TF.583. A more detailed description is included in the reference given above:

M: minute marker (second marker No. 0): 0.1 s

R: second marker No. 15 indicates service request to the DCF77 signal generation system

A1: announcement of a forthcoming change from CET to CEST or vice versa

Z1, Z2: zone time indication: UTC: Z1 0.1 s, Z2 0.1 s

UTC+1 h: Z1 0.1 s, Z2 0.2 s (referred to as Central European Time CET)

UTC+2 h: Z1 0.2 s, Z2 0.1 s (referred to as Central European Summer Time CEST)

UTC+3 h: Z1 0.2 s, Z2 0.2 s

A2: announcement of a leap second

S: starting bit of the coded time information, 0.2 s

P1, P2, P3: parity check bits

The information transmitted using the second markers Nos. 1 to 14 is provided by third parties. Currently weather data are transmitted, which are supplied by the Swiss Meteo Time GmbH.

A forthcoming change from CET to CEST or from CEST to CET is indicated during one hour before the change by second marker No. 16 being 0.2 s. Similarly, the second marker No. 19 is 0.2 s during the hour preceding the introduction of a leap second.

When a positive leap second (the normal case) is inserted, second marker No. 59 is emitted for 0.1 s and in the following second No. 60 of that particular minute the amplitude reduction is omitted.

The three parity check bits P1, P2, and P3 complete the preceding information words (7 bits for the minute, 6 bits for the hour and 22 bits for the date, including the day of the week) to form an even number of binary ones.

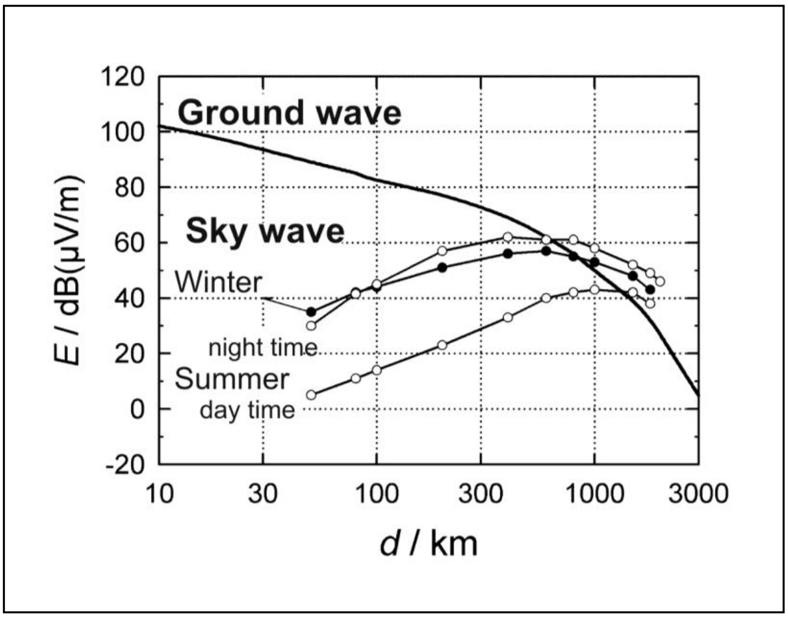
5.2 Reception of DCF77 and expected field strength

The isotropically emitted power of the DCF77 stations was estimated to 30 kW by the operators. The emitted signal reaches the place of reception in two ways: On the one hand, it propagates as ground wave along the Earth's surface and, on the other hand, it reaches the place of reception as sky wave after reflection on the ionospheric D layer. The expected field strength was calculated based on the following resources: Handbook "The ionosphere and its effects on radiowave propagation", ITU, Radiocommunication Bureau, Geneva 1998; Recommendation ITU-R P.368 – Ground wave propagation curves for frequencies between 10 kHz and 30 MHz, and Recommendation ITU-R P.684 – Prediction of field strength at frequencies below 150 kHz for ground wave and sky wave propagation. Details can be found in the reference given in § 5.1, from which Fig. 26 is copied. The largest electric field strength (100 dB(µV/m) occurring close to the transmitter corresponds to a magnetic field strength of 49 dB(µA/m), respectively 280 µA/m.

To give an impression on the geographical range of reception, the map reproduced as Fig. 27 illustrates in a simplified way lines of identical field strength as circles.

FIGURE 26

Field strengths of the ground wave and of the sky wave as a function of the distance *d* from the place of transmission, calculated based on ITU-Reports specified in the text, assuming an irradiated power of 30 kW



Note to Fig. 26: For the ground conductivity, $3 \cdot 10^{-3}$ S/m were assumed. A distinction is made between propagation in summer (open symbols) and in winter (full symbols).

FIGURE 27

Schematic view of the reach of the DCF77 transmission

5.3 Signal properties and reception experiences

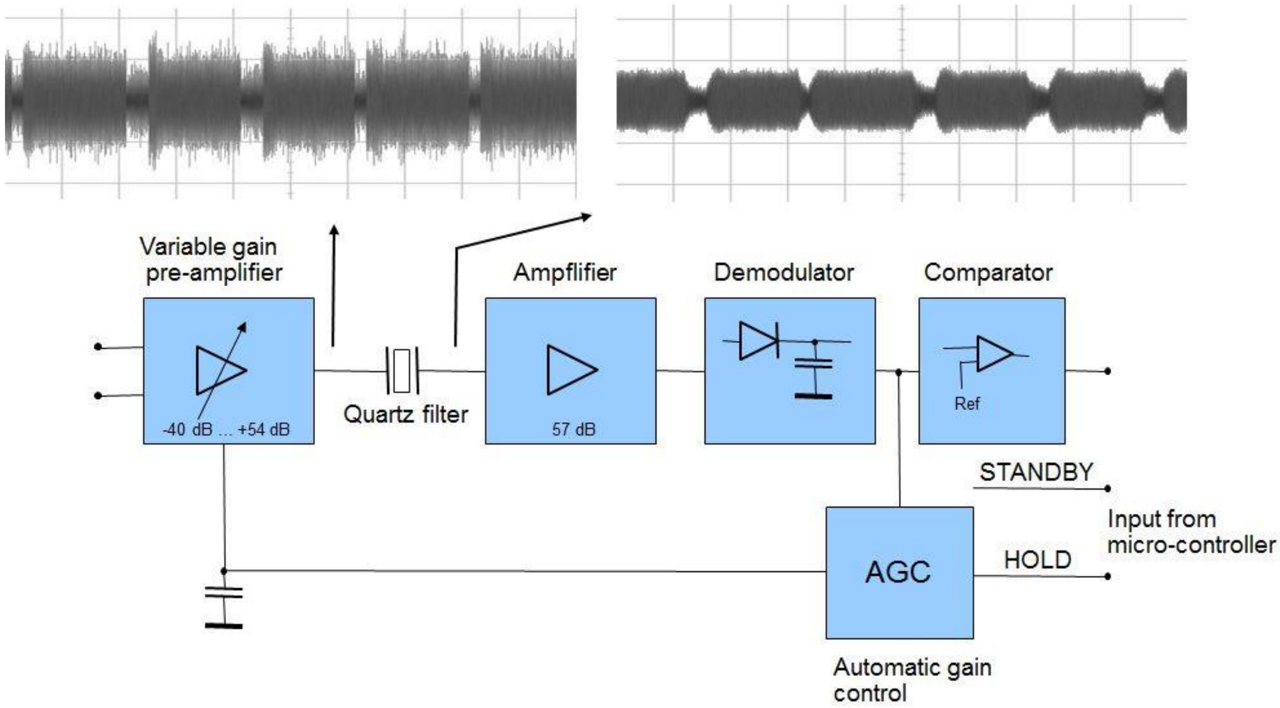
Measurements of the frequency spectrum performed at the place of transmission, as well as an estimate of the transmission bandwidth Δf from the decay time t_A of the falling edge of the carrier envelope shown in Fig. 24 according to the relation $\Delta f = 1/t_A$, have shown that Δf is in the order of magnitude of 850 Hz. The larger the fraction that is cut off from this frequency range on the receiving side, the larger the statistic uncertainty with which the arrival time of the DCF77 signal can be determined. To guarantee a reception free from interferences, many of the radio-controlled clocks on the market work with bandwidths around 10 Hz and timing uncertainties of approx. 0.1 s, which is regarded as sufficient for this purpose. In the context of this document, the use of DF77 as a standard frequency transmitter is less relevant as the number of suitable receivers is quite small compared to the number of “ordinary” DCF77 radio-controlled clocks.

In the following, experiences of receiver makers are collected, and published results about AM and PM reception in general were summarized. In § 5.4, some estimates and results that were explicitly made for the assessment of co-existence of DCF77 reception in the neighbourhood of Wireless-Power Transfer (WPT) stations are provided.

5.3.1 AM decoding Example 1

FIGURE 28

Schematic view of the components of a DCF77 AM receiver and typical signals received before and after the quartz filter (provided by HKW-Elektronik GmbH)



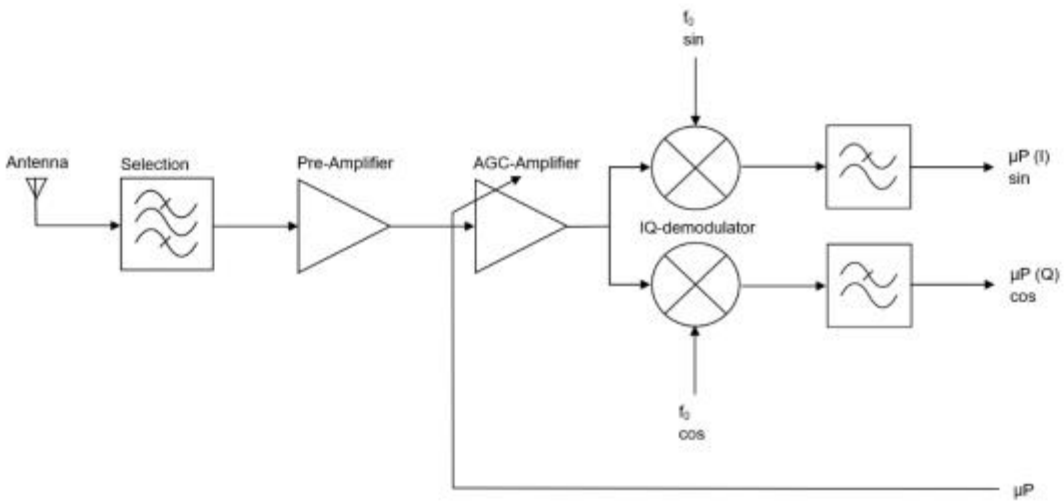
A typical receiver configuration realized inter-alia in DCF77-controlled wrist watches is depicted in Fig. 28. The variable gain pre-amplifier is essential to cope with the highly variable signal level received from the antenna (not shown). The variations are caused by the distance from the transmitter as well as by the orientation of the antenna. The gain adjustments must on the other hand be sufficiently slow in order not to react on the amplitude modulation with one-second rate.

As a rule of thumb, provided by the manufacturer HKW-Elektronik GmbH, a safe and reliable decoding of the AM code can be achieved when a disturbing signal at 77.5 kHz is less than 1/5 of the nominal use field, corresponding to a protection level of 14 dB at carrier. If the disturber is narrow band, single frequency, with stable amplitude, the conditions could likely be relaxed.

5.3.2 AM and PM reception with a quadrature receiver

FIGURE 29

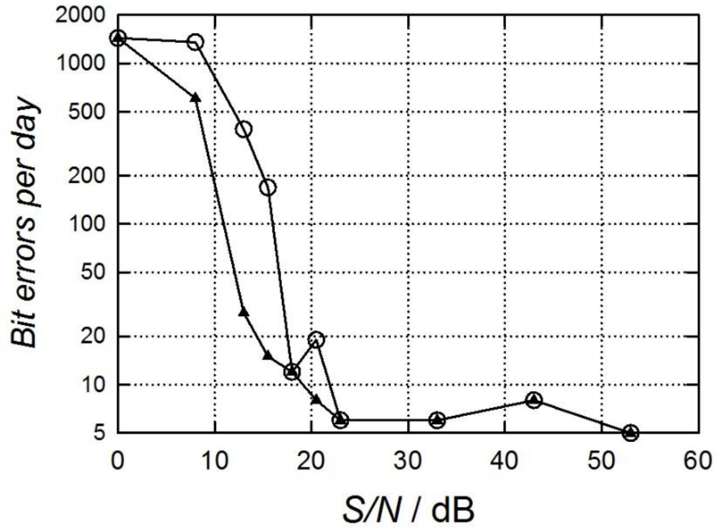
Schematic view of a quadrature demodulation receiver (provided by Meinberg GmbH)



The type of receiver as shown on Fig. 29 has been developed by Meinberg GmbH and is used in professional equipment intended inter alia as source of time-of-day information for distribution in networks via different protocols (Meinberg proprietary, IRIG-B, NTP, PTP). As an antenna, typically a 140 mm ferrite rod is used, and the antenna receive bandwidth is about 400 Hz. It is followed by an impedance matching circuit. The function of the AGC-amplifier is the same as explained before. Behind the I-Q demodulator, active low-pass filters limit the receive bandwidth to 10 Hz. A device like this was used to study the bit-error rate as a function of the SNR at the input. The results are shown in Fig. 30. The results clearly show the advantage of using the PM, but this is not an option for low-cost miniaturized consumer-electronics receivers. The DCF77 signal was generated with a signal simulator, either as AM or PM. Nevertheless, even in AM mode and at a *S*/*N* of 15.5 dB, the bit error rate is only about 0.35 per 3 minutes. An error-free reception during 3 minutes would safely allow to decode the full AM sequences and even do minimum consistency checks between successively received one-minute bit sequences.

FIGURE 30

Results of bit error rate per day as a function of *S*/*N* variation at the input to Meinberg DCF77 PZF 180 AM (open cricle) and PM (triangles) receiver in alternate operations



5.3.3 Results of further studies on DCF77 reception

DCF77 represents a valid source of time-of-day information in central Europe, in competition with the more accurate signals transmitted from Global Navigation Satellite Systems, such as GPS and Galileo. With this in mind, several studies were made regarding the performance and limitations of DCF77 receiver architecture. The results were mostly published as PhD-theses in German language, but typically they address the high-accuracy equipment, not the consumer market devices. In his paper "Performance analysis and receiver architecture of DCF77 radio-controlled clocks" (D. Engler, IEEE Transactions UFFC, vol. 59(5), pp 869-884 (2012)) the author proposes new types of receiving and data content decoding instrumentation and algorithms which are, however, mostly applicable in continuously operated, AC-powered DCF77-receivers in commercial and scientific applications. The author's findings suggest that for more advanced receiving equipment less stringent protection limits could be proposed than needed for typical devices as described in the previous sub-sections.

5.3.4 DCF77 reception under everyday conditions

The experimental studies referred to in the previous sub-sections were made under well-controlled conditions by experienced actors. The largest number of DCF77 clocks, however, is used under everyday conditions that likely deviate from such ideal situations. Here to mention are:

- Mismatch between antenna sensitivity curve and the DCF77 signal transmitter location. Typically, the minimum sensitivity is very narrow, i.e. reception with somewhat reduced sensitivity is possible at most orientations of the antenna;
- Penetration loss into buildings.

In order to protect the DCF77 reception of the majority of deployed receivers, the protection limit would need to be 3 dB to 6 dB tighter than what follows from §§ 5.3.1 and 5.3.2.

5.4 Considerations and studies about WPT and DCF77 coexistence

Question ITU-R 210-3/1 (Wireless Power Transmission – WPT) reads "*What steps are required to ensure that radiocommunication services, including the radio astronomy service, are protected from WPT operations?*". Resolution **958** of WRC-15 identified the need for urgent studies in preparation for the 2019 World Radiocommunication Conference to assess the impact of WPT for electric vehicles and to investigate suitable harmonized frequency ranges which would minimize the impact of WPT systems on radiocommunication services in response to WRC-19 agenda item 9.1, issue 9.1.6.

According to ETSI EN 303 417 and ECC Report 289, the frequency band 79-90 kHz has been proposed for use in WPT. According to ERC 70-03 and cited in ECC Report 289, regarding the transmitter DCF77, the maximum magnetic field strength of a WPT signal in the frequency range 77.5 kHz ± 250 Hz at 10 m distance from a WPT installation shall not exceed 42 dB(μA/m). This corresponds to an electric field strength of 93 dB(μV/m), which is reached only in the very near of the transmitter, as can be inferred from Fig. 26.

5.4.1 Comparison of ETSI WPT operations regulations to DCF77 reception

The maximum permitted field strength in 10 m distance generated by a WPT device operated at the edge of the respective band, here at 79 kHz, is 72 dB(μA/m) according to ERC 70-03 and 68 dB(μA/m) according to ETSI EN 303 417.

The higher of the two values corresponds to an electric field strength of 123 dB(μV/m). It is contrasted to the receive field strength of 60 dB(μV/m), to be expected under favourable conditions up to 1 000 km from the transmitter (see Fig. 26). Given the 1.5 kHz offset of the disturbing field, a damping by in total 40 dB by antenna and quartz filter can be expected. Still a further damping by 23 dB + 14 dB would be needed to reach the condition of stable AM decoding as discussed in § 5.3.1 (disturbing signal < 1/5 of wanted signal at 77.5 kHz). This condition would be met at distance of > 4 × 10 m, assuming a faster signal decay than ~ 1/(distance)^2 which is typical in the near field. Of course, the situation would be more favourable if the WPT operating frequency was higher than 79 kHz.

5.4.2 Summary of studies on WPT operations with DCF77 reception by standard receivers

The German Administration initiated and conducted experimental tests of the compatibility of DCF77 reception with WPT emissions. The results were published as part of ECC Report 289. In brief, 11 DCF77 clocks were subjected to a DCF77 signal with 50 dB(μV/m) field strength (see Fig. 26) and simultaneously to a disturbing signal of variable amplitude and in the frequency range between 79 kHz and 90 kHz. A failure of reception has been defined as “no synchronization to the DCF77 AM time signal or extended duration of the synchronization by more than 1 minute than nominal (typically 3 minutes)”. It should be stressed that in all battery-powered DCF77 clocks the receiving circuit is indeed activated only for short intervals in order to extend the battery lifetime. The criterion is thus relevant for the largest part of clocks in use. Figures 31 and 32 present the so-called compatibility distance, the minimum distance from an active WPT installation at which DCF77 reception is still possible. The eleven devices were from different manufacturers, and the spread in the results, i.e. the large difference in their immunity against disturbing signals, can be explained with the detailed construction of the receiver, in particular of the quartz filter. It is noted that most of the receivers tested were somewhat less performant than what could be naively expected from the findings reported in § 5.3.2.

FIGURE 31

Compatibility distances at different wanted DCF77 field strength for a WPT frequency of 79 kHz. (Copy from ECC Report 289)

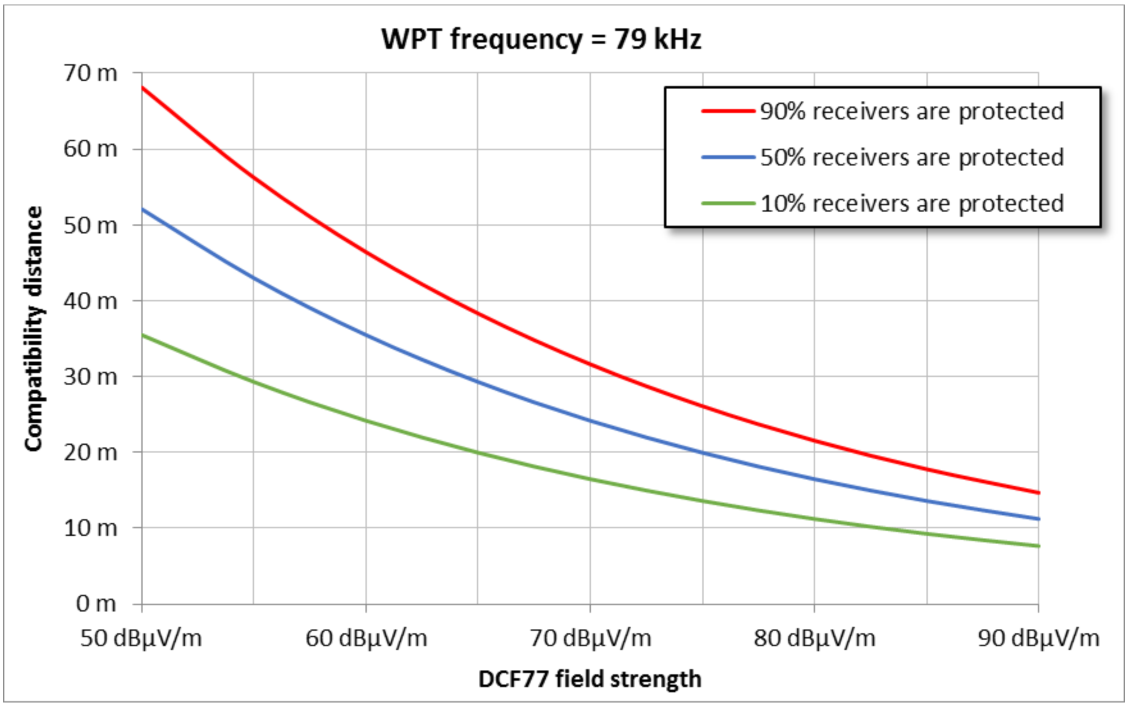
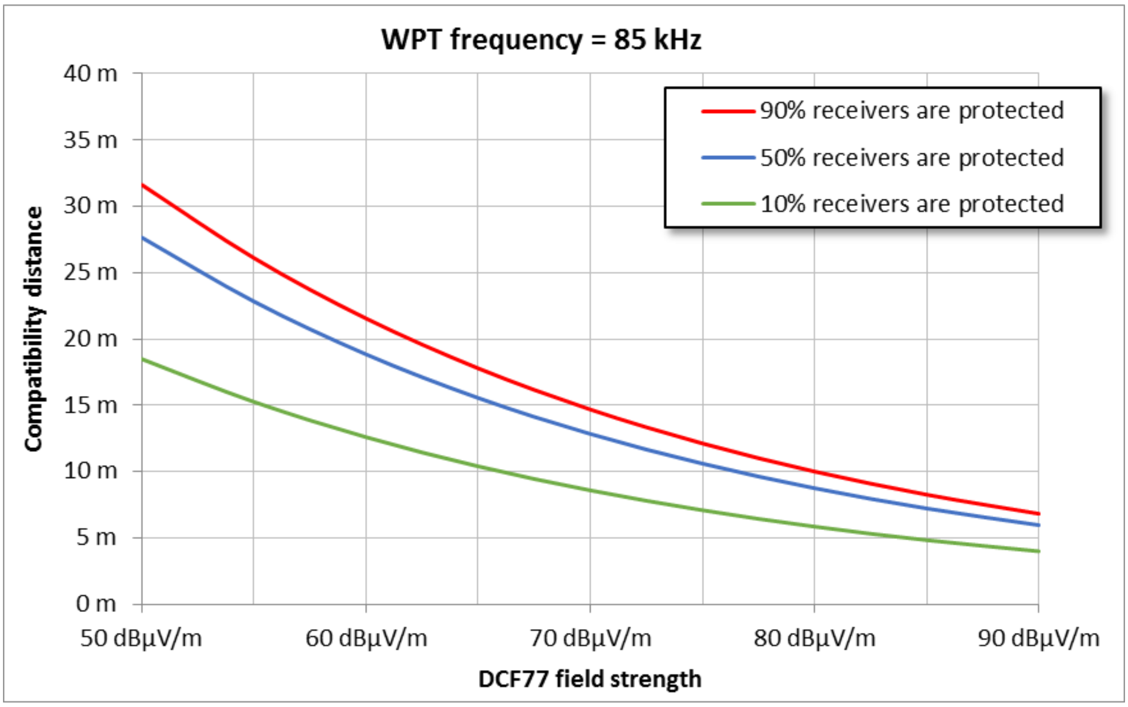


FIGURE 32

Compatibility distances at different wanted DCF77 field strength for a WPT frequency of 85 kHz. (Copy from ECC Report 289)



In summary, the prevailing regulations on WPT emissions in the band 79-90 kHz would prevent the reliable operation of consumer-grade DCF77 clocks at the distance of 10 m from an active WPT installation in a wide region of the traditional reception region (see Fig. 27).

6 Information on the Japanese SFTS service

6.1 Outline of the JJY service

The National Institute of Information and Communications Technology (NICT) determines and maintains the time and frequency standard and Japan Standard Time (JST) in Japan as the sole organization responsible for the national frequency standard. Japan Standard Time and Frequency generated by the NICT are transmitted throughout Japan via standard radio waves (JJY). JJY is the call sign of the radio station and a registered trademark (T4355749) of NICT.

Even though a transmission signal is precise, the precision of received signal may be reduced by factors such as conditions in the ionosphere. Such effects are particularly enhanced in the HF region, causing the frequency precision of the received wave to deteriorate to nearly 1×10^{-8} (i.e. the frequency differs from the standard frequency by about 1/100,000,000). Therefore, low frequencies, which are not susceptible to ionospheric conditions, are used for standard transmissions, allowing received signals to be used as more precise frequency standards. The precision obtained in LF transmissions, calculated as a 24-hour average of frequency comparison, is 1×10^{-11} (i.e. the frequency differs from the standard frequency usually by less than 1/100,000,000,000).

Standard radio waves transmitted by low frequency (LF) stations contain time-coded information on time, which is superposed on a highly precise carrier frequency signal. Signals from the two LF stations, Ohtakadoya-yama and Hagane-yama, cover almost all over Japan.

Ohtakadoya-yama LF Standard Time and Frequency Transmission Station has been transmitting the standard wave (40 kHz) since June 1999. Hagane-yama LF Standard Time and Frequency Transmission Station has been transmitting the standard wave (60 kHz) since October 2001.

Table 13 shows the characteristics of the stations, both of which operate 24 hours a day.

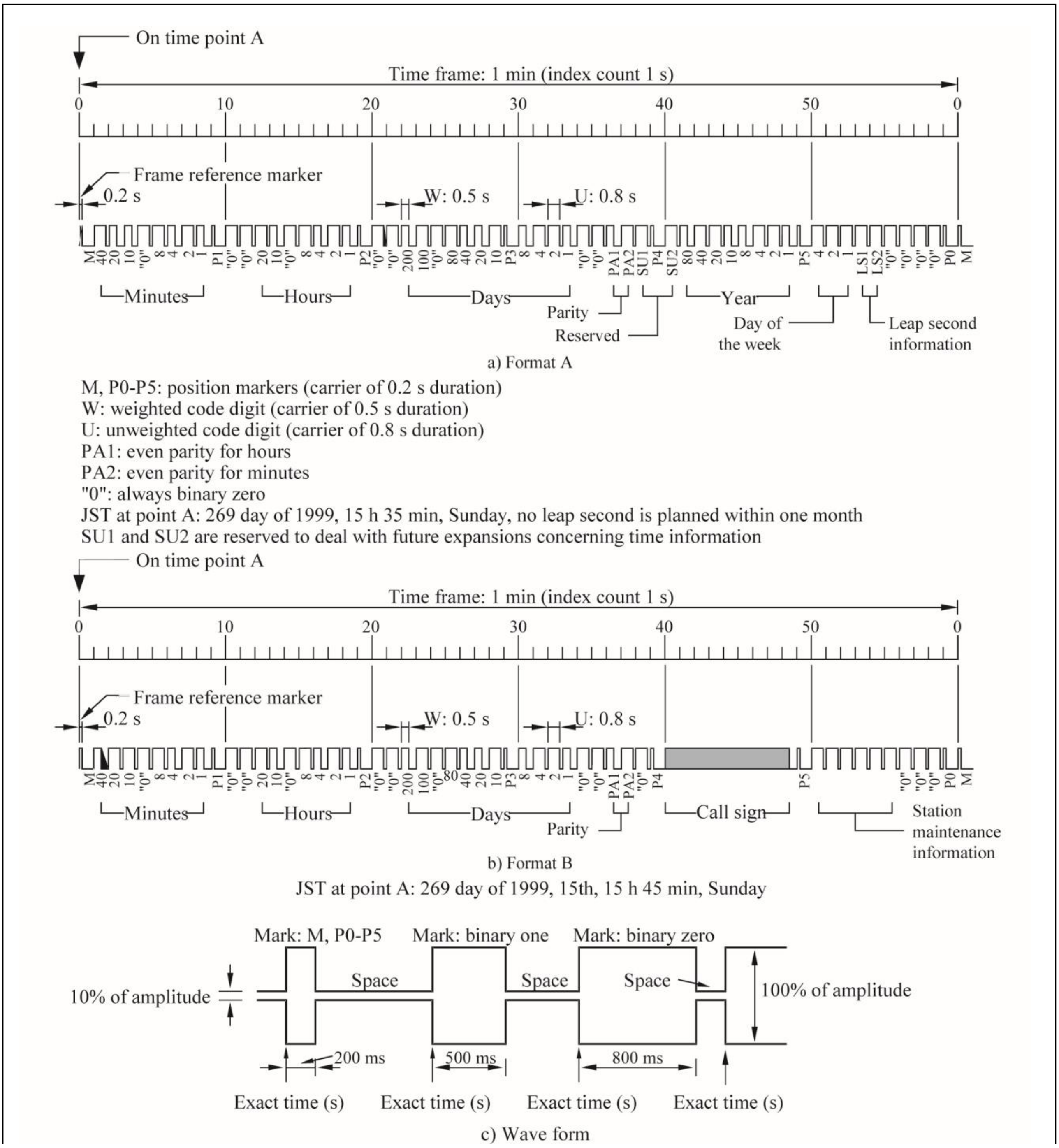
TABLE 13

Characteristics of JJY stations

Transmission station	Ohtakadoya -yama	Hagane –yama
Frequency	40 kHz	60 kHz
Antenna power	50 kW (antenna efficiency: approx. 25%)	50 kW (antenna efficiency: approx. 45%)
E.R.P	13 kW	23 kW
Frequency form	A1B	A1B
Operation time	Always	Always
Antenna height (Antenna form)	250 m (Umbrella type antenna)	200 m (Umbrella type antenna)
Latitude	37°22' N	33°28' N
Longitude	140°51' E	130°11' E
Elevation	790 m	900 m

FIGURE 33

JJY time code format



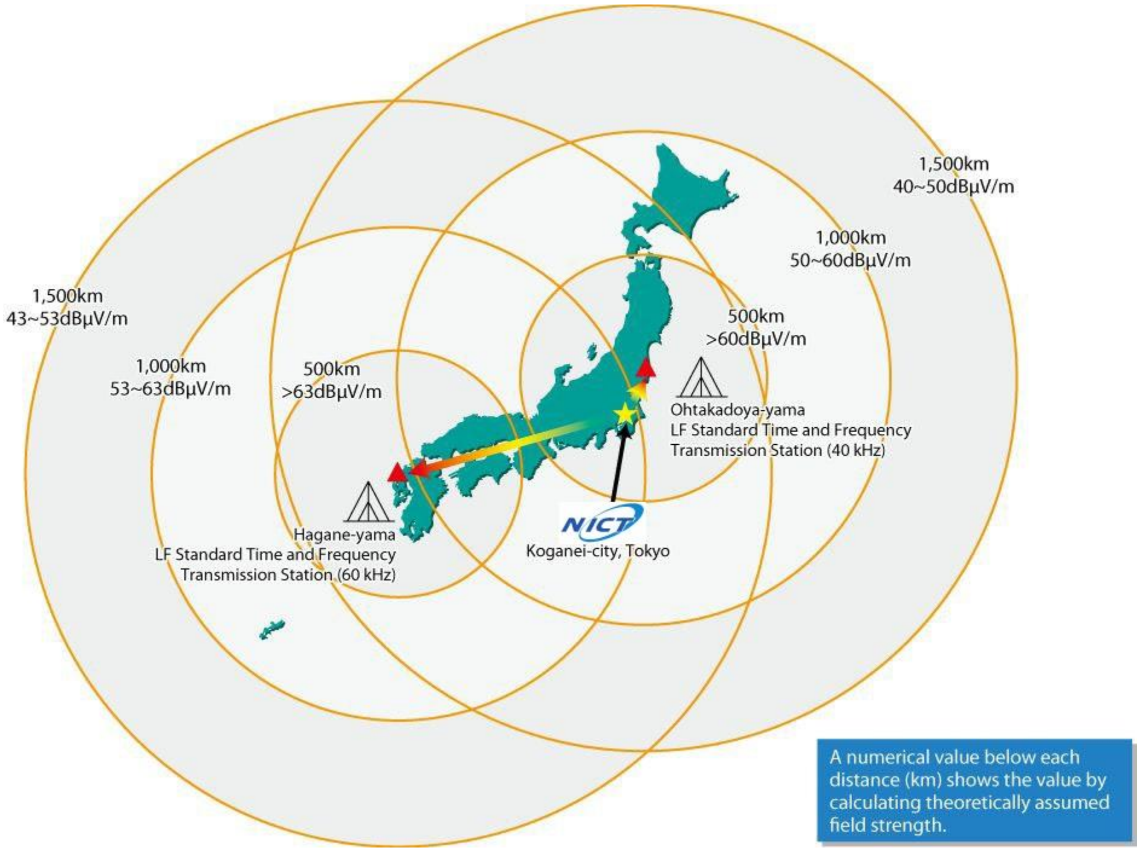
Note 1 – The time code format A is transmitted except in the 15th and the 45th minutes when the time code format B is transmitted.

The LF standard time and frequency transmission signal includes a time code which contains the following information as shown in Fig. 33: minute, hour, annual date (counted from January 1), year (last two digits of the dominical year), and day of the week.

This time code is used for reception devices such as radio-controlled clocks with automatic time correction functions.

FIGURE 34

Range of LF standard time and frequency transmission



6.2 Example characteristics of receivers

Figure 34 shows the approximate locations and the roughly estimated field strength of the two LF stations. More precise estimated values of field strength have been published on the NICT website https://jjy.nict.go.jp/fs/index-e.html (Valid at 5/30/2019).

In this website, NICT provides the expected LF field strength value in each prefecture in Japan, the graph of expected LF field strength value by the distance, the time and map of expected LF field strength value by the distance and the time, and the calculation program of expected value of LF field strength. This calculation method of LF & MF field strength is almost in conformity with Recommendation ITU-R P.684-7. NICT has provided the whole of Japan with the expected values of field strength data and has thus encouraged receiver makers to adapt their receiving circuits to the prevailing filed strength and other characteristics of the service.

The SFTS radio wave launched by Japan is used for various applications. There are two different directions of demand for users. The first is aimed at high-precision time and frequency comparison, which is mainly received by R & D institutes and operators of radiocommunication systems. In the first application, the SFTS radio waves which are constantly changing every time are continuously received to perform statistical instrumentation. Receivers of this type have sufficient stability and are carefully designed to reduce the effects of fading and man-made noise. Since the data rate of the SFTS is 1 bps (bit per second or 1 Hz) and the relative frequency inaccuracy never exceeds 10^{-11}, the required reception bandwidth of ±100 Hz may be sufficient. Such dedicated receivers comply with Article **3** of the RR, specifically Articles **3.12** and **3.13**.

On the other hand, the second application is simple time synchronization of clock/watch devices used by general consumers. A lot of radio-controlled clock/watch devices receiving the SFTS signal are available in the Japanese market. The characteristics of these receivers are described in § 6.3, and it is obvious that the immunity performance for eliminating jamming is not in all cases sufficient. It is considered that it is the result of effort for reducing the manufacturing cost of the commercial products and reducing their size and weight. For a considerable time, no effort has been made to improve the receiver characteristics, as there have been no significant interference sources in the environment and in the surrounding frequency bands to be expected. Therefore Japan Administration recognizes that the receiving characteristics of such devices do not adhere to the provisions of the Article **3** of the RR. They should be used under self-responsibility even if there is performance degradation due to interference.

6.3 Results of the study of impact to the clock/watch devices receiving SFTS from WPT for EV

6.3.1 Potential interference from WPT

The study of impact to the clock/watch devices receiving SFTS from WPT for EV was conducted by a working group (WG) for WPT rule-making in the Ministry of Internal Affairs and Communications (MIC), Japan. The WG consisted of technology experts and representatives in the related fields including WPT industries, intended incumbent radio systems, EMC, radio wave exposure and academia. The study results were incorporated into Japanese radio regulation and guidelines for WPT operation; and then, the new rule became effective in March 2016.

WPT devices whose radiated emission are lower than the emission limits described in Table 14 will not cause harmful interference, which is defined by *C/I* derived from the minimum receiver sensitivity of the radio-controlled clock/watch devices using SFTS in agreed use cases. Separation distance of 10 m was agreed and used to assess the impact to those devices. In addition, operation time range of the device to receive the SFTS which is not overlapping with WPT operation, diversity of SFTS wave propagation direction, and expecting receiver performance improvement in the future of those devices were taken into assessment. Consequently, the impact of WPT systems to radio-controlled clock/watch devices has been confirmed to be small enough not to cause harmful interference.

TABLE 14

Radiated emission limits of WPT systems for EV using 79-90 kHz in Japan's study

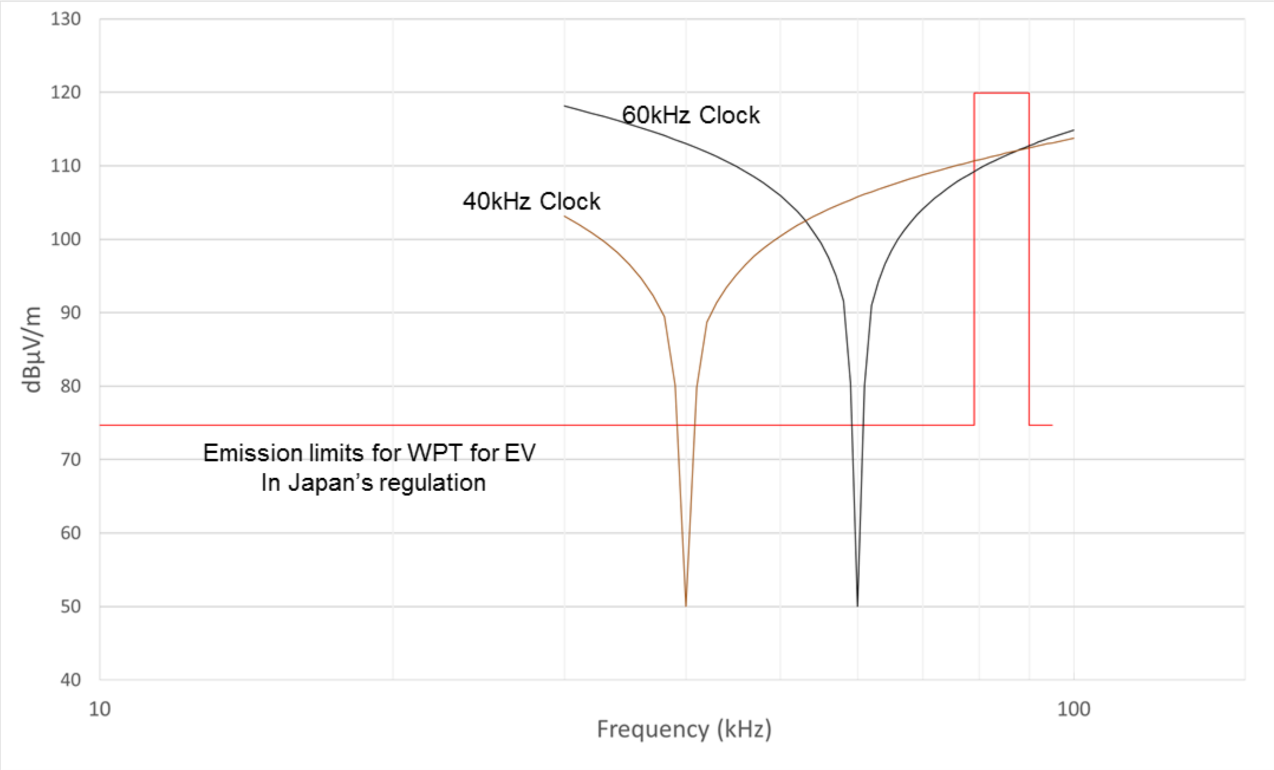
	Radiated emission limits
WPT frequency range (frequency range used for power transmission), 79-90 kHz	68.4 dBμA/m @ 10 m for 3 kW Tx Power 72.5 dBμA/m @ 10 m for 7.7 kW Tx Power
Frequency range from 526.5-1 606.5 kHz (MF broadcasting services frequency range)	−2.0 dBμA/m @ 10 m
Other frequency ranges under 3 MHz expect for 526.5-1 606.5 kHz	23.1 dBμA/m @ 10 m

Figure 35 plots emission limit of WPT for EV and allowable interference field strength of SFTS receiver in the interested frequency range. Here, the receiver is assumed to receive SFTS at field strength 50 dBμV/m (i.e. at the maximum receiver sensitivity), and the receiver performance has been derived from measurement results of commercial radio clock/watch devices receiving the 40/60 kHz transmitted SFTS waveforms. From this figure, it is expected that SFTS waveforms at

40/60 kHz received with a field strength of 50 dBμV/m or less might suffer harmful interference by emission of WPT for EV in 79-90 kHz received at higher field strength than 110 dBμV/m. The lower frequency band assigned for WPT for EV is not going to be permitted for use in Japan.

FIGURE 35

Emission limit for WPT for EV and allowable interference field strength of SFTS receiver



Thus, some types of widely used commercial radio-controlled clock/watch devices might come up against a problem caused by low immunity against interference signals and poor frequency selectivity of receivers.

6.3.2 Discussion of estimated required charging time of EV

An Estimation of Charging Time was studied.

The time required for charging an EV can be assumed, for example, as follows.

– The current EV has a battery capacity of 20 to 40 kWh. In the fully charged state, it is possible to travel 400 km. The charging time until the battery is fully charged requires about 5 hours by the 7.7 kW class charger (This charger corresponds to the class of MF-WPT2 of IEC Standard 61980-3).

– However, with daily use of passenger cars it will not run until its empty.

– According to Japanese passenger car usage time statistics surveyed by the National Institute of Technology and Evaluation (NITE)[*1)], survey results on the weekly men's driving/operating hours are 1.3 hours median and 1.8 hours average. When this is multiplied by the average vehicle speed of 24.4 km/h in the fuel economy driving mode JC 08, the average travel distance per day is estimated to be 44 km.

– To that extra charge requires the addition of 4.4 kWh, the required time is less than 40 minutes. If one charges at home and destination, the required time can be half of them.

*Reference*1): NITE:' Life and behaviour pattern information - About basic attribute and statistical information of investigation object', Section 4.1: Driving time of car, Revised 6th edition (in Japanese), July 2017.*

6.3.3 Probability of overlap with clock synchronization and charging of WPT-EV

SFTS receiving timing distribution for time adjustment was studied.

Radio-controlled clock/watch devices receive SFTS data automatically to keep own time adjusted to the reference time. Table 15 shows scheduled time distribution for automatic time adjustment of several commercial products where the time of day is given in Japan Standard Time. To receive data certainly every day, all companies' watches/clocks receive data during 2:00am – 5:00am.

TABLE 15

Scheduled timing distribution for automatic time adjustment

		Time to start receiving time data signal																							
		12	13	14	15	16	17	18	19	20	21	22	23	0	1	2	3	4	5	6	7	8	9	10	11
Watch 1	Company A															○	△	△							
Watch 2																○		△							
Watch 3																○	△								
Watch 4	Company B															○	△	△							
Watch 5																○		△							
Watch 6	Company C													○	△	△	△	△	△						
Watch 7																○	△	△							
Watch 8															○	○	△	△	△						
Clock 9	Company D	○	○	○	○	○	○	○	○	○	○	○	○	○	○	○	○	○	○	○	○	○	○	○	○
Clock 10			○	○	○	○	○	○	○	○	○	○			○	○	○	○	○	○	○	○	○	○	
Clock 11				○				○				○				○	○			○				○	
Clock 12	Company E			○			○			○			○			○			○			○			○
Clock 13																○	△	△	△	△					
Clock 14		○	○	○	○	○	○	○	○	○	○	○	○	○	○	○	○	○	○	○	○	○	○	○	○
Clock 15		Every 3 hours																							
Clock 16	Company F	△	△	△												○	△	△							
Clock 17		○		○		○		○		○		○		○		○		○		○		○		○	
Clock 18			○	○	○	○									○	○	○	○							

Note:
"circle" means "primary scheduled adjustment timing"
"triangle" means "secondary adjustment timing for backup".

6.3.4 Considerations on the impact to the clock/watch devices receiving SFTS from WPT for EV

1) Potential impact to SFTS receiver performance and mitigation measure

Receiver performance of radio-controlled watch/clock devices receiving SFTS might be degraded by receiver blocking caused by WPT emission in its operating frequency range due to insufficient sensitivities of SFTS receiving devices.

It should be noted that the event can be observed only when planned SFTS receiving timing falls into wireless EV charging period. Thus, harmful interference event may not continue beyond the overlapped period. WPT charging time alignment programs must thrive on timing adjustment to solve SFTS receiver blocking issues.

2) Compatibility framework agreed between the technologies of WPT for EV and radio-controlled watch/clock devices

In the WG (see § 6.3.1), leaders from WPT for EV proponents and representatives from the radio-controlled watch/clock device industry reached a consensus on compatibility framework on the two technologies. The baseline was that WPT for EV with the proposed limits of 68.4 dBμA/m @

85 kHz band (79-90 kHz) can be used while the radio-controlled watch/clock devices for 40/60 kHz are in use practically throughout Japan. The following points were carefully considered and agreed.

– Minimum received field strength 50 dBμV/m may be relaxed by about 10 dB;
– Wireless EV charging period is not always overlapped with the SFTS reception timing of radio-controlled watch/clock devices;
– SFTS arriving direction having maximum field strength at the receiver device may not always be same to the main direction of the WPT device;
– In the WPT device manual or on the WPT product, the following instruction or equivalent should be indicated: “Possible harmful electro-magnetic interference to the radio-controlled watch/clock devices receiving SFTS.”

6.3.5 Summary of impact study for WPT for EV in Japan

Based on the above consideration, in Japan, the use of WPT for EV in consideration of SFTS is controlled as follows:

WPT for EV devices shall not cause harmful interference defined by *C*/*I* derived from the minimum receiver sensitivity of the radio-controlled watch/clock devices in agreed use cases. Separation distance of 10 m should be used as a coexistence criterion. Additional measures on operation time non-overlapping between WPT and the radio-controlled watch/clock, radio propagation direction variation, and possible performance improvement were taken into consideration.

TABLE 16

Radiated emission limits of WPT systems to protect SFTS services, specifically for JJY

Frequency range of fundamental frequency	Protection criteria	Comments
10-79 kHz	The radiated emission limit for WPT for EV device is 23.1 dBμA/m measured at a distance of 10 metres.(1) Any WPT-EV device should make an individual application to the Minister of Internal Affairs and Communications and use it only if it is authorized.	WPT operation is not permitted at 40 kHz and 60 kHz where SFTS (JJY) services are operated.
79-90 kHz	The radiated emission limit for WPT for EV device (maximum power output 7.7 kW) is 68.4 dBμA/m measured at a distance of 10 metres.	In the WPT device manual or on the WPT product, the following instruction or equivalent should be indicated: “Possible harmful electro-magnetic interference to the radio-controlled watch/clock devices receiving SFTS.”
90-150 kHz	The radiated emission limit for WPT for EV device is 23.1 dBμA/m measured at a distance of 10 metres. Any WPT for EV device should make an individual application to the Minister of Internal Affairs and Communications and use it only if it is authorized.	

(1) This emission level is the same as for ‘Industrial facilities emitting radio waves’ in Japan.

7 Information on the Chinese SFTS services

7.1 BPC

BPC Time Signal Emission BPC is the call sign of the Chinese Low-Frequency Time-Code Radio Station at 68.5 kHz. It is located at 34.457°N, 115.837°E in Shanqiu of Henan Province, please see Fig. 36. BPC was completed in 2007 and its time signal is traced to UTC (NTSC) which is realized by National Time Service Center, the Chinese Academy of Sciences (NTSC) in China.

FIGURE 36

BPC transmitter location

Some BPC important parameters are as follows:

Antenna structure: Single tower umbrella antenna

Transmitter power: 100 kW

Carrier frequency: 68.5 kHz

Emission bandwidth: ±1 kHz

Signal coverage range: 3 000 km for sky wave and 1 000 km for ground wave

Modulation mode: pulsed negative polarity keying provided by code modulation unit

Timing accuracy: 1 ms for sky wave and 0.1 ms for ground wave.

Receiver sensitivity: 40 dBμV/m.

7.2 BPL time signal emission

7.2.1 Overview of BPL service

The BPL station, which is maintained by NTSC, is located in PuCheng and its coordinates are 34°56' north, 109°32' east. The BPL station was completed in 1979 and started service in 1983. It is the primary and legal means of dissemination for Chinese Standard Time (CST). BPL could cover the continent and the coastal area of China and provides timing precision of micro-second. It disseminates standard frequency (carrier frequency 100 kHz) and time-of-day information as an infrastructure service of the Chinese government. The time and frequency information are traced to UTC (NTSC), which generated by the timekeeping laboratory in LinTong which is approximately 70 km away from PuCheng.

FIGURE 37

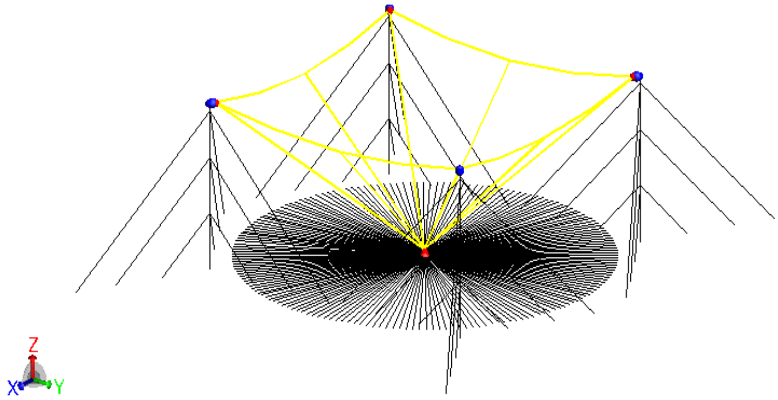
BPL transmitter location

7.2.2 Technical description of BPL

The BPL transmitter effective radiated power is 2 000 kW and the antenna structure is four-tower inverted cone antenna. It adopts the Loran system of signals structures and follows Recommendation ITU-R M.589-3 – Technical characteristics of methods of data transmission and interference protection for radio navigation services in the frequency bands between 70 and 130 kHz.

FIGURE 38

Schematic diagram of four tower inverted cone antenna



Pulse group structure analysis

The BPL signal is a phase modulated pulse with a carrier frequency of 100 kHz and its bandwidth is 20 kHz. The standard current waveform at the bottom of the transmitting antenna is strictly defined as:

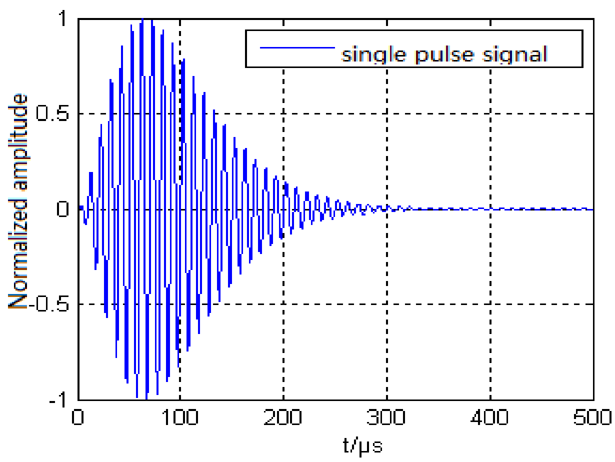
$$s_0(t) = \begin{cases} 0, & t < \tau \\ A(t-\tau)^2 exp\left[\dfrac{-2(t-\tau)}{65}\right] \sin[\omega_0 t + p_c(m)], & \tau \le t \le 65+\tau \\ Undefined, & t > 65+\tau \end{cases}$$

where:

A :	a constant related to the peak current
t :	time, µs
τ :	envelope to Cycle Difference, µs
ω_0 :	carrier frequency
$p_c(m)$:	phase encoding, 0 or π.

FIGURE 39

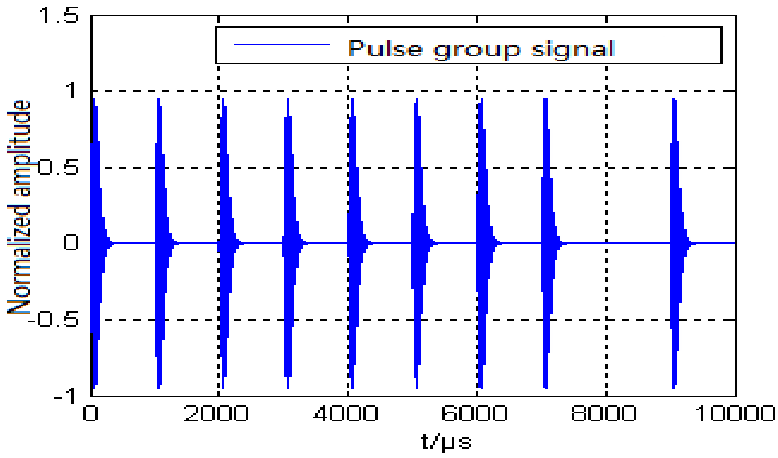
Single pulse signal diagram



BPL emits nine pulses per pulse group, the first eight pulses are separated by 1000 µs, and the eighth and ninth pulses are separated by 2000 µs. The time interval between two sets of adjacent pulse group signals which is called the Group Repetition Interval (GRI) amounts to 60000 µs in case of BPL.

FIGURE 40

Pulse group signal diagram



Data coding characteristics

The BPL system uses the pulse group signal as a carrier to establish a data communication system. This method is called Eurofix. Modulation method is 3-level plus position modulation.

BPL firstly arranges and encodes time information to be transmitted. After then, modulates it onto the BPL pulse group signal by means of 3-level plus position modulation, and then broadcast it to the user. BPL receiver could receive and decode time information. Finally, the RS error correction and the CRC cyclic redundancy check are performed.

There are 30 GRIs in one frame for Eurofix digital modulation. 8 GRIs are used for data information, and other 22 GRIs are occupied by error correction.

RS error correction coding parameters

Primitive polynomial: $\mathrm{P}(x) = x^7 + x^3 + 1$

Code generator polynomial: $\mathrm{g}(x) = (x - a)(x - a^2)(x - a^3)\cdots(x - a^{20})$

CRC cyclic redundancy check:

Information code polynomial: $M(x) = a_{55}x^{55} + a_{54}x^{54} + \cdots + a_1x^1 + a_0x^0$

Generate polynomial: $\mathrm{G}(x) = x^{14} + x^{13} + x^7 + x^5 + x^4 + 1$

TABLE 17

Modulation pattern and data correspondence

Decimal data	Hexadecimal data	Modulation pattern	Decimal data	Hexadecimal data	Modulation pattern
0	0	--00++	64	40	+-0-+0
1	1	--0+0+	65	41	+-00-+
2	2	--0++0	66	42	+-00+-
3	3	--+00+	67	43	+-0+-0

TABLE 17 (*cont.*)

Decimal data	Hexadecimal data	Modulation pattern	Decimal data	Hexadecimal data	Modulation pattern
4	4	--+0+0	68	44	+-0+0-
5	5	--++00	69	45	+-+-00
6	6	-0-0++	70	46	+-+0-0
7	7	-0-+0+	71	47	+-+00-
8	8	-0-++0	72	48	+0--0+
9	9	-00-++	73	49	+0--+0
10	A	-00+-+	74	4A	+0-0-+
11	B	-00++-	75	4B	+0-0+-
12	C	-0+-0+	76	4C	+0-+-0
13	D	-0+-+0	77	4D	+0-+0-
14	E	-0+0-+	78	4E	+00--+
15	F	-0+0+-	79	4F	+00-+-
16	10	-0++-0	80	50	+00+--
17	11	-0++0-	81	51	+0+--0
18	12	-+-00+	82	52	+0+-0-
19	13	-+-0+0	83	53	+0+0--
20	14	-+-+00	84	54	++--00
21	15	-+0-0+	85	55	++-0-0
22	16	-+0-+0	86	56	++-00-
23	17	-+00-+	87	57	++0--0
24	18	-+00+-	88	58	++0-0-
25	19	-+0+-0	89	59	++00--
26	1A	-+0+0-	90	5A	-0000+
27	1B	-++-00	91	5B	-000+0
28	1C	-++0-0	92	5C	-00+00
29	1D	-++00-	93	5D	-0+000
30	1E	0--0++	94	5E	-+0000
31	1F	0--+0+	95	5F	0-000+
32	20	0--++0	96	60	0-00+0
33	21	0-0-++	97	61	0-0+00
34	22	0-0+-+	98	62	0-+000
35	23	0-0++-	99	63	00-00+
36	24	0-+-0+	100	64	00-0+0
37	25	0-+-+0	101	65	00-+00
38	26	0-+0-+	102	66	000-0+
39	27	0-+0+-	103	67	000-+0
40	28	0-++-0	104	68	0000-+
41	29	0-++0-	105	69	0000+-

TABLE 17 (*end*)

Decimal data	Hexadecimal data	Modulation pattern	Decimal data	Hexadecimal data	Modulation pattern
42	2A	00--++	106	6A	000+-0
43	2B	00-+-+	107	6B	000+0-
44	2C	00-++-	108	6C	00+-00
45	2D	00+--+	109	6D	00+0-0
46	2E	00+-+-	110	6E	00+00-
47	2F	00++--	111	6F	0+-000
48	30	0+--0+	112	70	0+0-00
49	31	0+--+0	113	71	0+00-0
50	32	0+-0-+	114	72	0+000-
51	33	0+-0+-	115	73	+-0000
52	34	0+-+-0	116	74	+0-000
53	35	0+-+0-	117	75	+00-00
54	36	0+0--+	118	76	+000-0
55	37	0+0-+-	119	77	+-+-+-
56	38	0+0+--	120	78	-+-+-+
57	39	0++--0	121	79	+-+--+
58	3A	0++-0-	122	7A	-+-++-
59	3B	0++0--	123	7B	+--+-+
60	3C	+--00+	124	7C	-++-+-
61	3D	+--0+0	125	7D	+--++-
62	3E	+--+00	126	7E	-++--+
63	3F	+-0-0+	127	7F	+0000-

TABLE 18

Timing message type I

Bit	Length	Information	Unit	Range
1-4	4	Message type	1	16
5-6	2	Message subtype	1	4
7-8	2	System status	1	4
9-13	5	Hour	h	32 (>24)
14-19	6	Minute	min	64 (>60)
20-25	6	Second	s	64 (>1 min)
26-31	6	Year	year	64 (2000-2063 year)
32-35	4	Month	month	16 (>12)
36-40	5	Day	day	32 (>31)
41-47	7	Leap-second	s	128
48-56	9	Broadcast deviation	10 ns (resolution)	±2.56 µs

TABLE 18 (*end*)

Bit	Length	Information	Unit	Range
57-70	14	CRC		
71-210	140	RS		

TABLE 19

Timing message type II

Bit	Length	Information	Unit	Range
1-4	4	Message type	1	16
5-6	2	Message subtype	1	4
7-8	2	System status	1	4
9-13	5	Hour	h	32 (>24)
14-19	6	Minute	min	64 (>60)
20-25	6	Second	s	64 (>60)
26-35	10	Millisecond	ms	1024 (>1000)
36-45	10	Microsecond	µs	1024 (>1000)
46-52	7	Ten nanosecond	10 ns (resolution)	1.28 µs (>1.00 µs)
53-56	4	Station logo	1	16
57-70	14	CRC		
71-210	140	RS		

Message description:

(1) Message type

The message type occupies 4 bits and indicates the type of information. International Recommendations have assigned 1, 2 and 5 to differential GPS, differential GLONASS and text information. Therefore, the timing message is tentatively set to 4, and the corresponding binary is "0100".

(2) Message subtype

Message subtype occupies 2 bits. The timing message type I and II is tentatively set to "01", the Timing message type is defined as "10", and the remaining two subtypes are used as timing relay messages.

(3) Year

Year occupies 6 bits, and the maximum can be 64. Here only the last two digits of the year byte are taken. Tentatively starting from 2000, the year number can be expressed until 2063.

(4) Month

Month occupies 4 bits, represents 12 months from January to December in a year.

(5) Day

Day occupies 5 bits, represents 31 days from 1 to 31 in 1 month.

(6) Time

Time occupies 5 bits, represents 24 hours from 0 to 23 in a day.

(7) Minute

Minute occupies 6 bits, represents 60 minutes from 0 to 59 in 1 hour.

(8) Seconds

Seconds occupies 6 bits, represents 60 seconds from 0 to 60 in 1 minute.

(9) Millisecond

Millisecond occupies 10 bits. It is from 0 to 999.

(10) Microsecond

Microsecond occupies 10 bits. It is from 0 to 999.

(11) Ten nanoseconds

Ten nanoseconds occupies 7 bits. It is from 0 to 99, the range is from 0.00 to 0.99 microseconds.

(12) Leap-second

Leap-second=TAI-UTC and it occupies 7 bits. The maximum can represent 128.

(13) System status

Indicates the system working status, which are ‘normal’, ‘warning’ and ‘alarm’, and it takes up 2 bits.

(14) Broadcast deviation

Broadcast deviation occupies 10 bits, the highest bit is the sign bit, and the resolution is 10 ns, indicating range of −2.56 microseconds to +2.56 microseconds.

(15) Station logo

Station logo occupies 4 bits.

(16) CRC

Cyclic redundancy check occupies 14 bits.

(17) RS

Reed-Solomon code occupies 140 bits

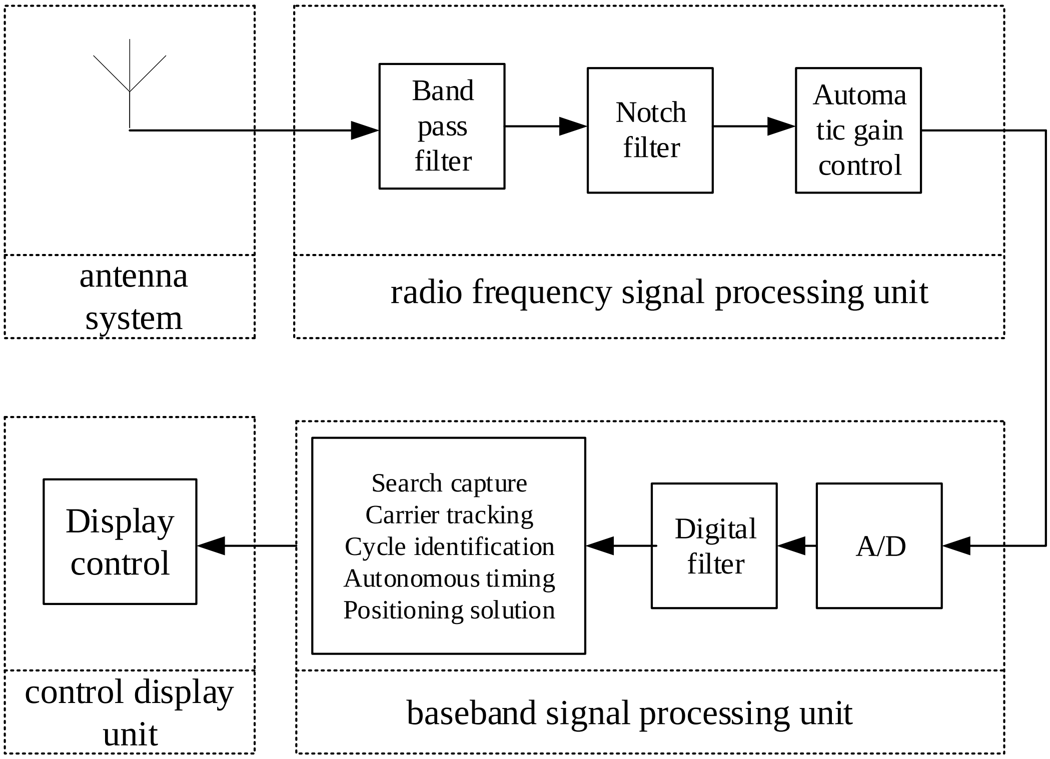
(18) Additional instructions

(19) Signed data information, the most significant bit is the sign bit, and the others are data bits. Sign bit ‘1’ means ‘-‘ and ‘0’ means ‘+’. The highest bit of the data bit is ‘Left’ and the lowest bit is ‘Right’ with ‘Right’ alignment. When the information data bit length is less than the data bit length allocated by the message, fill in the ‘left’ side of the data to fill ‘0’. When the information data bit length is greater than the data bit length allocated by the message, fill in the data bit with "1".

7.2.3 BPL receivers

The BPL receiver is mainly composed of antenna system, radio frequency signal processing unit, baseband signal processing unit and control display unit, please see Fig. 41.

FIGURE 41
BPL receiver diagram



BPL receiver could be designed according to the Chinese national standard GB12752, which is 'General specification for marine Loran-C receiving equipment', and the general conditions for the normal operation of the receiver are as follows:

(1) Signal level: 25~110 dBμV/m

(2) Envelope to cycle difference (ECD): -2.4~2.4 μs

(3) Noise level: 12~75 dBμV/m

(4) Signal-to-noise ratio (SNR): SNR ≥ −10 dB

The receiver should be able to achieve to accuracy of 0.3 μs or better. The receiver sensitivity should be better than 25 dBμV/m. The lock time is no more than 7.5 min. The receiver should be able to enter the locking state under the condition of sky wave interference with a delay of 37.5~60 μs and a relative level of 12~26 dB.

7.3 BPM Time Signal Emission

7.3.1 Overview of BPM service

BPM is operated by NTSC in China. It is located at 35°00′N 109°31′E in Pucheng of Shaanxi. In 1970, Chinese government began to build the BPM. In 1980, BPM provided time and frequency service at 2.5 MHz, 5 MHz, 10 MHz and 15 MHz alternatively 24 hours and it can cover the whole country with precision of 1 ms.

FIGURE 42

BPM short-wave time transmission station

7.3.2 BPM broadcast programme

TABLE 20

BPM broadcast programme

Time slot	Start time within the hour mm:ss	Stop time within the hour mm:ss	Data content
1	59:00	00:00	Station identification
2	00:00	10:00	UTC time signal
3	10:00	15:00	Carrier only (No modulation)
4	15:00	25:00	UTC time signal
5	25:00	29:00	UT1 time signal
6	29:00	30:00	Station identification
7	30:00	40:00	UTC time signal
8	40:00	45:00	Carrier only (No modulation)
9	45:00	55:00	UTC time signal
10	55:00	59:00	UT1 time signal

7.3.3 Station identification format

Morse call sign（—••• •——• ——） is for 40 seconds, and then voice announcement "BPM 标准时间标准频率发播台" ("BPM standard time and frequency transmission station") is twice for 20 seconds.

7.3.4 Time signals format

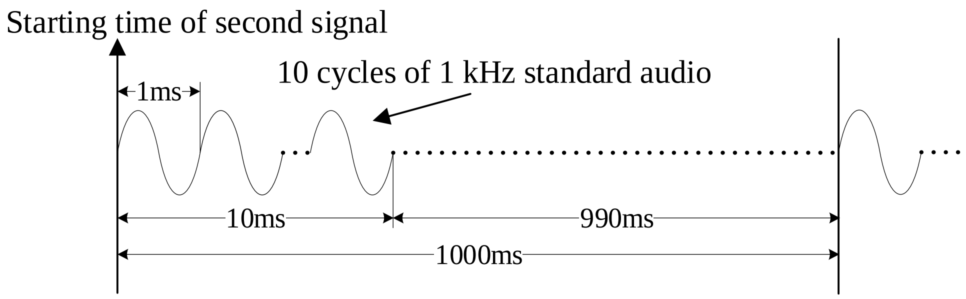
UTC time signals

Include UTC second and minute information.

1) UTC second: Modulated by 1 kHz standard frequency with 10 ms duration. The starting phase of the first period of the sinusoidal wave signal is zero, and its time is the starting time of the UTC second signal.

FIGURE 43

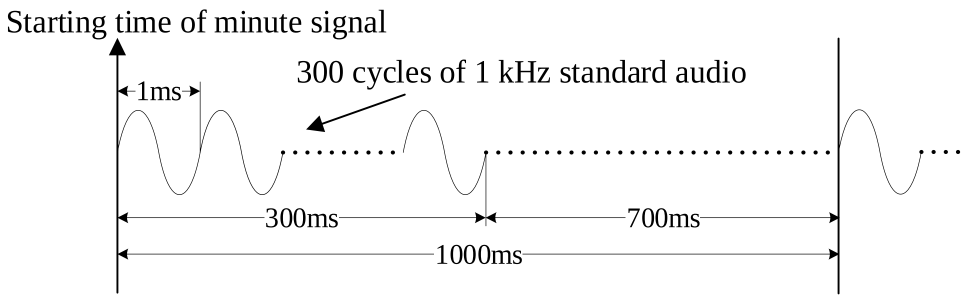
UTC second signal waveform



2) UTC minute: Modulated by 1 kHz standard frequency with 300 ms duration. The starting phase of the first period of the sinusoidal wave signal is zero, and its time is the starting time of the UTC minute signal

FIGURE 44

UTC minute signal waveform



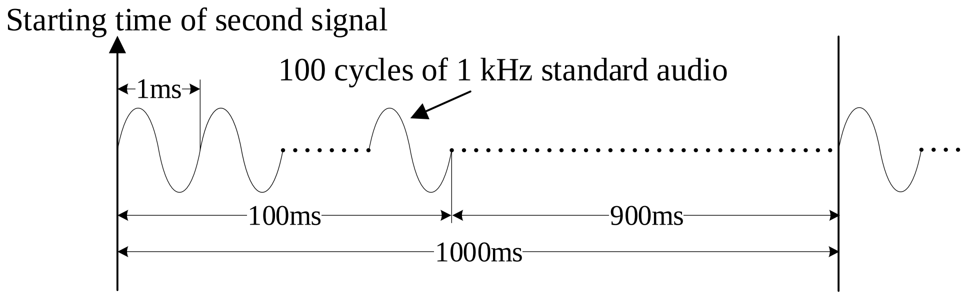
UT1 time signals

Include UT1 second and minute information.

1) UT1 second signal: Modulated by 1 kHz standard frequency with 100 ms duration. The starting phase of the first period of the sinusoidal wave signal is zero, and its time is the starting time of the UT1 second signal.

FIGURE 45

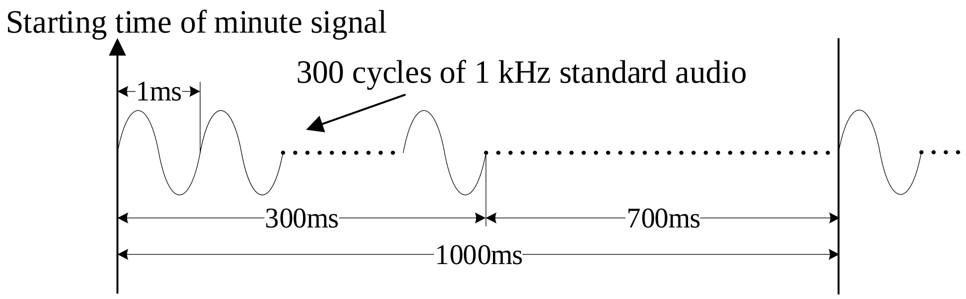
UT1 second signal waveform



2) UT1 minute signal: Modulated by 1 kHz standard frequency with 300 ms duration. The starting phase of the first period of the sinusoidal wave signal is zero, and its time is the starting time of the UT1 minute signal.

FIGURE 46

UT1 minute signal waveform



No modulated carrier

Emission of only carrier frequency signal only, without modulation signal.

7.3.5 Broadcast time, Carrier power and antenna type

Some key parameters of BPM are showed in Table 21.

TABLE 21

Broadcast time, broadcast power and antenna type

Carrier (MHz)	Broadcast time		Carrier power (kW)	Antenna type
	UTC	CST (Chinese Standard time)		
2.5	07:30-01:00	15:30-09:00	1	Horizontal dipole angle cage
5	00:00-24:00	00:00-24:00	10	Horizontal dipole angle cage
10	00:00-24:00	00:00-24:00	10	Horizontal dipole angle cage
15	01:00-09:00	09:00-17:00	10	Horizontal dipole angle cage

8 Information on the Russian SFTS services

The reference signals of frequency and time are a means to transfer of the unit of time and time scales and consists of carrier oscillations modulated in amplitude, phase or the frequency of the signals, containing the timestamps of the time scale, and information about current values of time, date and other additional information.

The reference signals of frequency and time are designed to transmit the unit of time and frequency and the coordinated time scale UTC(SU) from the state standard to the reference and working measuring instruments to ensure uniformity of measurements of time and frequency inside the country. To transmit reference signals, the State Service of Time and Frequency and Earth Rotation Parameters Determination (in short SSTF) uses an extensive transmission network, which includes:

- Long waves specialized radio stations (RBU 66.6 kHz, RTZ 50.0 kHz)
- Short waves specialized radio stations (RWM 4 996 kHz, 9 996 kHz, 14 996 kHz)

8.1 RBU and RTZ Service

RBU Location – 56° 44'N 37° 40'E.

The carrier frequency – 66.(6) kHz. The effective radiated power is 50 kW.

RTZ Location – 52° 26'N 103° 41'E.

The carrier frequency – 50.0 kHz. The effective radiated power is 10 kW.

Relative uncertainty of the carrier frequency – 2×10^{-12} for RBU and RTZ.

8.1.1 RBU protection limits

No specific protection limits exceeding those of other services of that kind which have been identified in ERC Recommendation 70-03, Table 9bis, page 28:

TABLE 22

RBU protection limits

Station	Carrier frequency	Protection bandwidth	Maximum field strength at 10 m
RBU	66.6 kHz	±750 Hz	42 dBµA/m

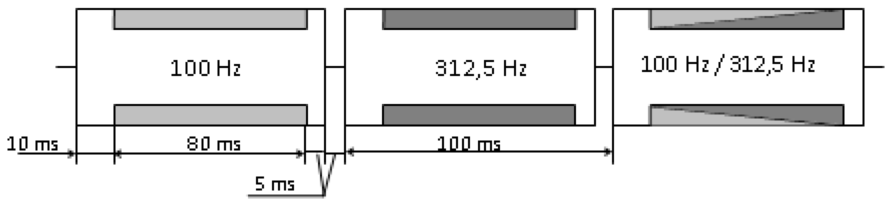
8.1.2 Form of the reference frequency and time signals

The RBU and RTZ radio stations use the signals of the DXXXW type for transmitting the sizes of the time and frequency units.

DXXXW signals (Fig. 47) are sinusoidal form carrier with a frequency f_H, interrupted during every 100 ms for a time of 5 ms; in 10 ms after interruption, carrier oscillations during 80 ms undergo narrow-band phase modulation by sinusoidal signals with subcarrier frequency of 100 or 312.5 Hz and a modulation index of 0.698. Signals with a subcarrier frequency of 312.5 Hz are used for marking units in binary code when transmitting information on time scales, as well as for marking second and minute marks. Signals with a frequency of 100 Hz are used to mark the zeros in the binary code when transmitting information on time scales, as well as to fill all the remaining 80 ms intervals free from the transmission of any information.

FIGURE 47

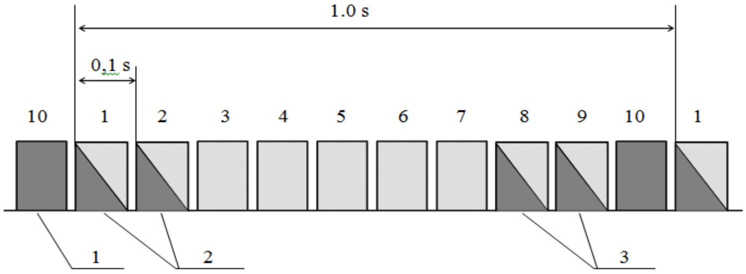
Fragment of DXXXW type signals



Second marks are identified by marking of the preceding 0.1 s interval with signals 312.5 Hz. Minute labels are identified by additional marking with signals 312.5 Hz of two 0.1 s intervals preceding the second marker (Fig. 48).

FIGURE 48

Signal information structure
1 – second marker, 2 – information signals, 3 – minute marker



8.1.3 Information coding, transmitted as part of DXXXW signals

The time code transmitted as constituent of the reference signals of frequency and time is based on two types of codes:

– the code for the transmission of DUT1 as described in Annex 2 of R-REC-TF.460-6;

– binary-decimal with parity check to transmit the rest of the information.

The full code format contains 120 elements (60 elements in the first 0.1-second interval and 60 elements in the second 0.1-second interval) and is transmitted in a cycle of 1 min. The beginning of the minute cycle (minute mark) is identified by additional marking of the eighth and ninth 0.1 second intervals.

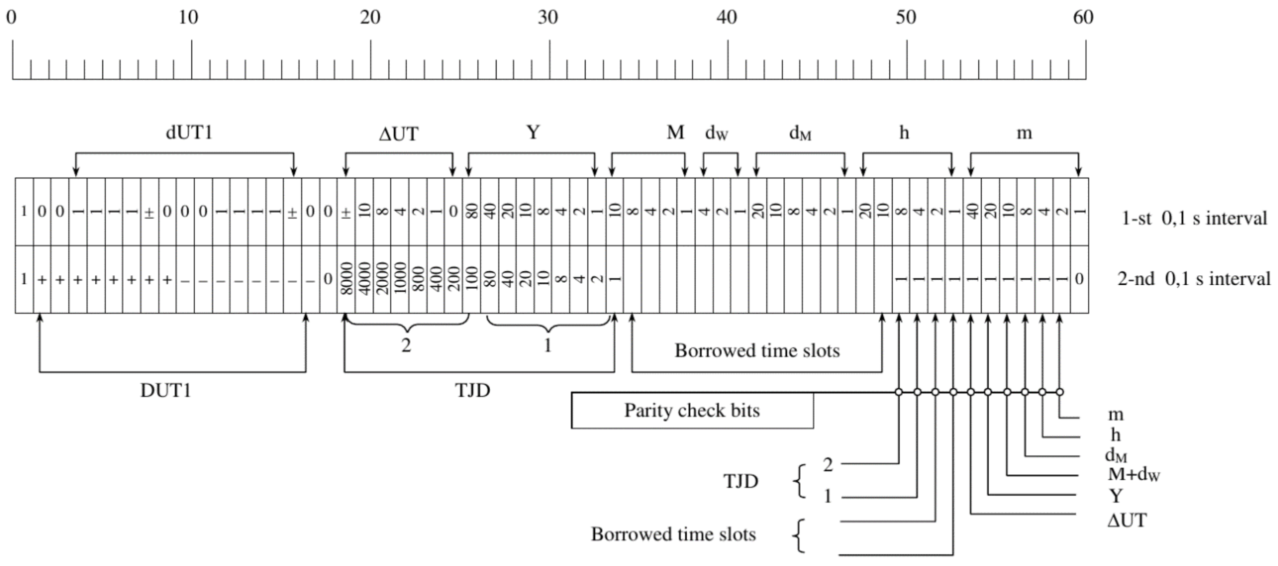
The RBU and RTZ radio stations transmit, as constituent of the frequency and time reference signals information on the difference between Universal Time (UT1) and Coordinated Universal Time, UT1-UTC = DUT1 + dUT1. In this case, the values of DUT1 are transmitted according to the data of the International Earth Rotation and Reference Systems Service (IERS), and the values of dUT1 according to the State Service of Time and Frequency data.

The format of the time code and the content of the transmitted information are presented graphically in Fig. 49.

Information about the current values of the time of day is presented in hours (h) and minutes (m), transmitted in the scale of Moscow time with the offset ΔUT amendment relative to universal coordinated time added. Information on calendar dates includes: the value of the year of the century (Y), the value of the month of the current year (M), the value of the day of the month (d_m) and the value of the day of the week (d_w). Information on the Julian date includes a shortened Julian date (TJD), which is the four lower-orders of the numerical value of the modified Julian date (MJD).

FIGURE 49

Time code format



8.2 RWM service

RWM Location: 56° 44'N 37° 38'E.

The carriers frequency: 4996.0 kHz, 9996.0 kHz, 14996.0 kHz.

The effective radiated power is 10 kW.

Relative uncertainty of the carrier frequency: 5×10^{-12}, averaging for one day or longer.

8.2.1 Form of the reference frequency and time signals

The RWM radio station uses the signals of the NON-type for transmitting the sizes of the time and frequency units, and the signals A1X and A1N – for transmitting the time scales. To transmit time marks signals with a repetition frequency of 1 Hz and 10 Hz are used.

The duration of the A1X signals with a repetition frequency of 1 Hz (second signals) amounts to 100 ms. Signals occurring at the beginning of each minute are extended to 500 ms. The duration of A1N signals with a repetition frequency of 10 Hz is 20 ms. Signals occurring at the beginning of every second extend to 40 ms, and of every minute – to 500 ms.

The characteristic point for A1X and A1N signals is the beginning of the rising edge of radio signal.

The RWM radio stations transmit, as constituent of the frequency and time reference signals information on the difference between Universal Time and Coordinated Universal Time UT1-UTC = DUT1 + dUT1. In this case, the values of DUT1 are transmitted according to the data of the IERS, and the values of dUT1 – according to the SSTF data.

The numerical value of the difference UT1-UTC is denoted DUT1 and rounded to 0.1 s and is transmitted through the radio stations as described in Annex 2 of R-REC-TF.460-6. Besides the DUT1 information, the SSTF radio stations transmit additional information dUT1, specifying the UT1-UTC difference values to 0.02 s.

The information DUTl + dUTl is transmitted after each minute time signal by marking the corresponding second signals with additional impulses.

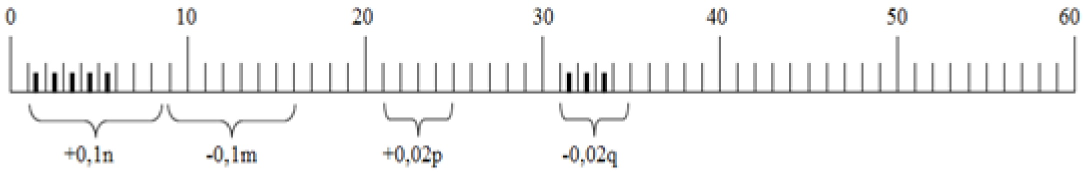
A positive value of DUT1 is transmitted by marking **n** second signals in the interval between the 1st and the 8th second, so DUT1 = +0.1 n seconds. A negative value of DUT1 is transmitted by marking the **m** second signals in the interval between the 9th and the 16th second, so that DUT1 = −0.1 m seconds. The absence of marked signals in the interval between the minute signal and the 16th second means that DUT1 = 0.0 s.

A positive dUTl value is transmitted by marking **p** seconds signals in the interval from 21 to 24th seconds, so dUTl = +0.02 p seconds. A negative value - by marking **q** second signals in the interval between the 31st and the 34th second, so dUTl = −0.02 q seconds.

An example of DUTl + dUTl information coding is shown in Fig. 50.

FIGURE 50

Example of DUT1+dUT1 coding:
n=5; q=3; DUT1+dUT1 = 5 · 0.1 c - 3 · 0.02 c = +0.44 s



8.2.2 The hourly programme of the radio station RWM

TABLE 23

Hourly programme of the radio station RWM

Signal transmission time, hh:mm		**Type of signals**
start	**end**	
00:00	07:55	carrier signal
30:00	37:55	
08:00	09:00	no signal
38:00	39:00	
09:00	10:00	radio identification signals
39:00	40:00	
10:00	19:55	A1X signals, containing second signals, minute signals and information DUT1 + dUT
40:00	49:55	
20:00	29:55	A1N signals with a repetition rate of 10 Hz
50:00	59:55	

* Time signals of 56, 57, 58, 59th seconds following 9, 14, 19, 24, 29, 39, 44, 49, 54 and 59th minutes are skipped.

9 Information on the French SFTS service

9.1 Introduction

The Allouis ALS162 (formerly TDF) site is a long-wave broadcasting centre commissioned in 1938. Initially, the technical device was designed for the sole broadcast of a radio program. This site with an area of 0.95 km² is located near the town of Vierzon (47° 10’ 05” N, 02° 12’ 02” E). The long-wave signals propagate mostly as ground wave on the French territory, and on site the emission efficiency is enhanced by an earth network with an area of 0.85 km². This site currently broadcasts the time signal in phase modulation (G2B mode) on the frequency of 162 kHz with a power transmitted to the antenna input of 800 kW, 24 hours a day (except every Tuesday from 8 h to 12 h in French legal time).

The authorities responsible for the Allouis radio signal are:

– ANFR (Agence nationale des fréquences), 78 avenue du général de Gaulle, 94704 Maisons-Alfort, France;

– LNE (Laboratoire national de métrologie et d’essais), 1 rue Gaston Boissier, 75724 Paris Cedex 15, France;

– France Horlogerie (formerly CFHM, Chambre française de l’horlogerie et des microtechniques), 22 avenue Franklin Roosevelt, 75008 Paris, France.

The signal form is a phase modulation of the carrier by +1 and −1 rad in 0.1 s every second except the 59th second of each minute. This modulation is doubled to indicate binary 1. The numbers of the minute, hour, day of the month, day of the week, month and year are transmitted each minute from the 21st to the 58th second, in accordance with the French legal time scale. In addition, a binary 1 at the 17th second indicates that the local time is two hours ahead of UTC (summer time); a binary 1 at the 18th second indicates that the local time is one hour ahead of UTC (winter time); a binary 1 at the 14th second indicates that the current day is a public holiday (Christmas, 14 July, etc.); a binary 1 at the 13th second indicates that the current day is a day before a public holiday. The relative uncertainty of the carrier frequency is 2×10^{-12}.

A detailed description of the technical specifications of the Allouis radio signal transmitter is laid down in the French standard NF C90-002 (1988) – Broadcasting and telecommunication: Data broadcasting system compatible with AM sound broadcasting (version available only in French).

In the following, the results of a recent study are reported which was done in order to determine the protection criteria through sensitivity measurements of radiosynchronization receivers using as reference Allouis time signal at 162 kHz.

9.2 Description of the testing facility

Determination of the protection limits for the ALS162 signals requires precise knowledge of the sensitivity of radiosynchronization receivers currently deployed on the French territory. The following report aims to recall the measurement methodology used to characterize these receivers under conditions representative of their actual use and to present the results of the measurement campaign carried out at the TDF Antenna Measurement Centre in Liffré, France.

The measurements are performed in a parallel plate cell, specially optimized to characterize miniature antennas in VLF, LF, MF and HF bands. This cell has in particular the property of generating a quasi-TEM wave over a very wide frequency range (from 10 kHz to 150 MHz), covering 162 kHz. The flatness of the generated field and the associated wave impedance close to 120 π (≅ 377 Ω) make it possible to approximate the far field conditions encountered on site. The homogeneity of the field is guaranteed with a variation of the order of 1 dB in a volume of 60 cm ×

60 cm × 60 cm positioned in the centre of the cell. The list of equipment used for the measurements is given in Annex 1.

The parallel plate cell is itself placed in an anechoic chamber exhibiting effective shielding higher than 110 dB (plane waves) which enables to be free of any uncontrolled external interference that may disturb the measurement (including the time signal transmitted from Allouis).

The measurement campaign is carried out on a sample of nine receivers representative of the fleet currently deployed throughout the national territory.

Finally, it is important to stress that the measurements results should be used only to define the level of sensitivity of the receivers. The defined sensitivity must in no case be equated with the level of the useful signal to determine the coverage area of Allouis.

9.3 Measurement protocol and results

9.3.1 Step 1 – Calibration of the TEM cell at 162 kHz with a reference antenna

A 50 Ω generator (HP 8648D) injects an arbitrary power of +20 dBm, at 162 kHz, at the input of the parallel plate cell (Fig. 51). The characteristic impedance of the cell being 150 Ω, the latter is equipped at its input with an impedance adaptor (balun) which also ensures the symmetrisation of ports, by balancing the currents on the two plates. At the end of the cell, a 150 Ω load allows to recover the energy transmitted to the device and thus ensures a matched system over a very wide frequency range. A reference antenna (ETS-Lindgren 6502 active loop) is positioned in the centre of the cell, in the area of flatness and homogeneity of the EM field, and the voltage received is recorded using a spectrum analyser (Agilent E4446A).

The amplitude of the generated electric field E in the cell may be deduced according to the formula:

$$E\ (\text{dB}\mu\text{V/m}) = U\ (\text{dB}\mu\text{V}) + AF\ (\text{dB/m})$$

where U denotes the voltage measured at the terminals of the reference antenna and AF the (electric) antenna factor on 50 Ω (Table 24).

FIGURE 51

Calibration of the TEM cell at 162 kHz, in an anechoic chamber (20 m × 9 m × 9 m) with highly effective shielding of 110 dB

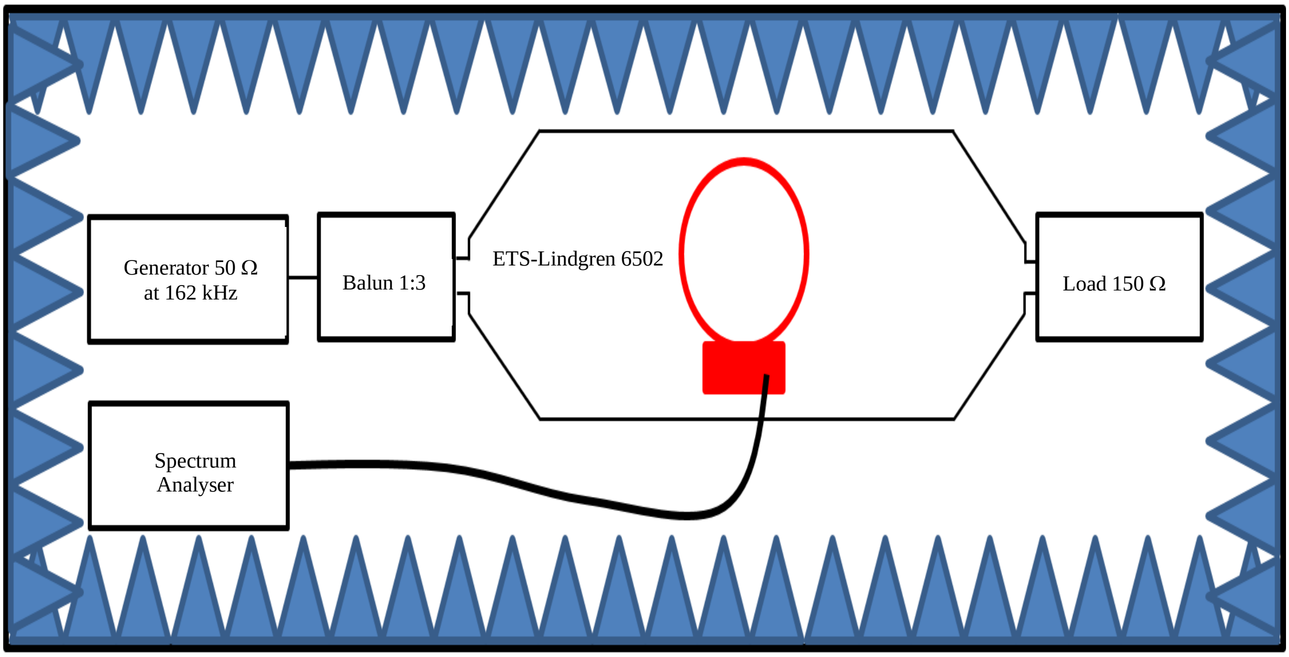


TABLE 24

Amplitude of the electric field generated at 162 kHz in the TEM cell, from the measurement of the voltage at the terminals of the loop antenna

Antenna model	Power injected (dBm)	Voltage (dBµV)	Antenna factor (dB/m)	E-field (dBµV/m)
ETS-Lindgren 6502	20	122.3	11.6 (1)	133.9

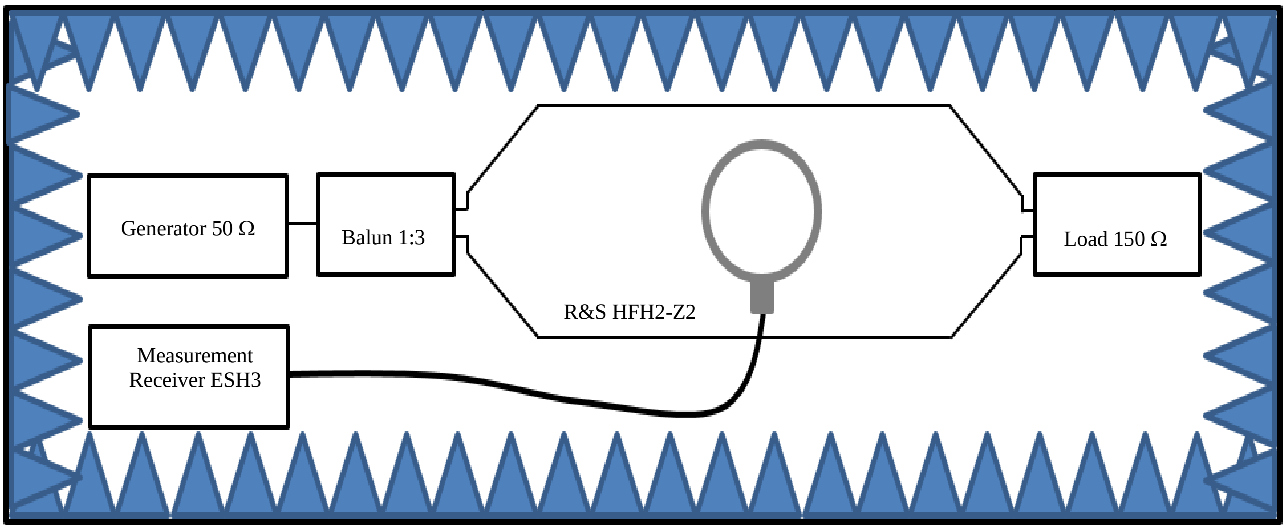
(1) The certified antenna factor value is provided by the manufacturer.

9.3.2 Step 2 – Verifying the linearity of the measurement device

The measuring device being entirely passive, and the balun being used at low power levels which do not allow the generation of significant non-linearities, the linearity must be maintained over a wide range of power injected by the generator. A first check is made at high power levels on the reference antenna used in Step 1 (ETS-Lindgren 6502) as well as on another reference antenna, the R&S HFH2-Z2 active loop (see Fig. 52). The latter being associated with an R&S ESH3 measurement receiver, the field strength measurement is a direct measurement, the *AF* being taken into account by the measuring device.

FIGURE 52

Testing the linearity of the measurement system using an active reference loop antenna



The measurement being carried out at high levels initially, a check is also made with a reference fieldmeter, the Narda EHP-200 (see Fig. 53).

FIGURE 53

Verifiying the linearity of the measurement system with a fieldmeter

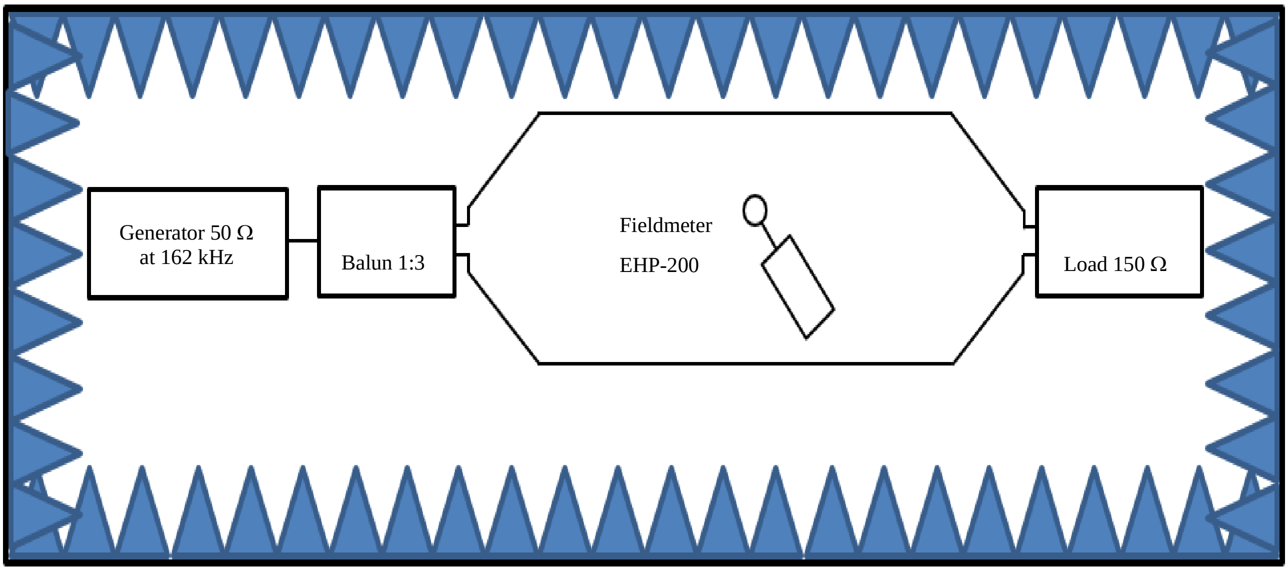


TABLE 25

Measurements of the amplitude of the electric field for different power levels emitted at 162 kHz

Reference antenna	Power injected (dBm)	Power received (dBm)	Voltage (dBμV)	Antenna factor (dB/m)	E-field (dBμV/m)	E-field (V/m)
ETS-Lindgren 6502	10	5.3	112.3	11.6	123.9	1.56
ETS-Lindgren 6502	20	15.3	122.3	11.6	133.9	4.95
HFH2-Z2 + ESH3	10				124.1	1.60
HFH2-Z2 + ESH3	20				133.9	4.95
EHP-200	10				124.0	1.58
EHP-200	20				133.4	4.68

The measurements carried out make it possible to confirm the linearity of the device at high power levels but also the good agreement of the electric field values generated in the cell. At 10 dBm and 20 dBm, the measurement difference noted between the three devices does not exceed 0.5 dB (Table 25).

In order to check the linearity of the device over a much larger range of electric field values, the output level of an HP 8648D generator is adjusted from +20 dBm to −90 dBm with 10 dB steps and the generated electric field is measured with the measuring receiver + loop antenna system ESH3 – HFH2-Z2 (Table 26).

TABLE 26

Linearity of the system by measuring the amplitude of the electric field induced in the TEM cell as a function of the power transmitted by the HP 8648D generator at 162 kHz

Frequency (kHz)	Power transmitted (dBm)	E-field induced (dBμV/m)	E-field induced (mV/m)
162	20	133.7	4 841.724
162	10	123.8	1 548.817
162	0	113.8	489.779
162	−10	103.8	154.882
162	−20	93.9	49.545
162	−30	83.8	15.488
162	−40	73.9	4.955
162	−50	63.8	1.549
162	−60	53.6	0.479
162	−70	43.7	0.153
162	−80	33.6	0.048
162	−90	23.7	0.015

As the differences between two consecutive values vary between 9.9 dB and 10.2 dB for 10 dB expected, the observed linearity is completely within the measurement tolerance. This measurement further allows to confirm the wide dynamic range of the measurement bench, including for weak electric fields (and here limited by the sensitivity of the active antenna (around 15 dBµV/m at nominal frequency)).

The same protocol is then reproduced with the use of a passive antenna. A loop or a ferrite antenna, could be used, here a TDF loop antenna was used whose antenna factor has been deduced, in the cell, by substitution with the ETS-Lindgren 6502 reference antenna and an Agilent E4446A spectrum analyser (see Fig. 54). Again, the measured electric field must be directly proportional to the power injected via the generator.

FIGURE 54

Testing the linearity of the measurement system using a TDF passive reference loop antenna

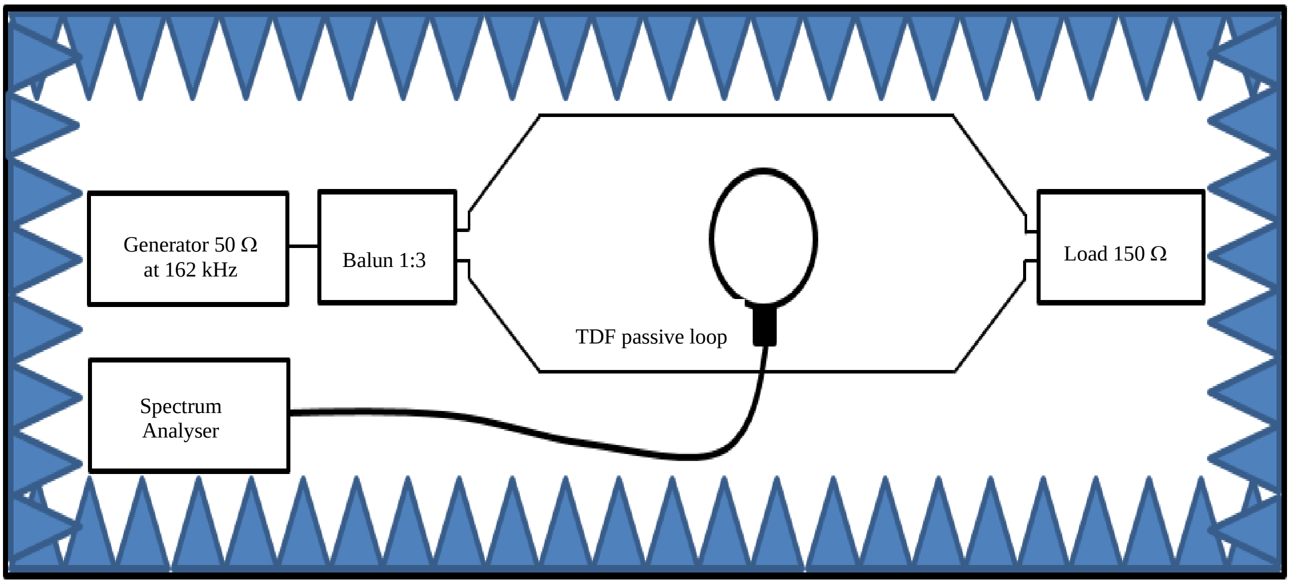


TABLE 27

Linearity of the system in measuring the voltage induced on the passive loop antenna and calculating the electric field induced in the TEM cell w.r.t the power transmitted by the HP 8648D generator at 162 kHz

Frequency (kHz)	Power transmitted (dBm)	Voltage induced (dBµV)	Antenna factor passive loop (dB/m)	E-field induced (dBµV/m)	E-field induced (mV/m)
162	20	71.9	61.8	133.7	4 841.724
162	10	62.2	61.8	124.0	1 548.893
162	0	51.9	61.8	113.7	484.172
162	−10	42.1	61.8	103.9	156.675
162	−20	32.2	61.8	94.0	50.119
162	−30	22.1	61.8	83.9	15.668
162	−40	12.3	61.8	74.1	5.070
162	−50	2.2	61.8	64.0	1.585

The same observation as with the active loop is performed, the linearity remains here also within the measurement tolerance (between 9.8 dB and 10.3 dB for 10 dB expected). The electric field values obtained for the same transmitted power are moreover identical to that obtained with the active loop, with a maximum deviation of 0.2 dB (Table 27).

Taking into account the very good effective linearity of the system, it is therefore possible to generate with high accuracy electric fields from a few µV/m to several V/m at 162 kHz.

9.3.3 Step 3 – Determination of the synchronisation time of radiosynchronization receivers in a noise-free environment and in the absence of misalignment

After checking that the time signal from Allouis was not received in the anechoic chamber, a reference time signal is injected into the parallel plate cell (see Fig. 55) via a suitable signal generator provided by the company Bodet (GENE-ALS162). A variable attenuator makes it possible to precisely control the output level of the generator. The generator output power is set to its maximum (approximately −18 dBm), at 162 kHz, to reach a signal level of 80 dBµV/m in the cell. For each device tested, the maximum synchronization time to the time signal is recorded, which is the reference synchronization time in an ideal environment (noise-free and in the absence of misalignment of the receiving antennas). For the following tests, the maximum time considered is the reference synchronization time to which a time lapse of 3 min is added for each device under test.

FIGURE 55

Determination of the synchronisation time of radiosynchronization receivers in a noise-free environment and in the absence of misalignment

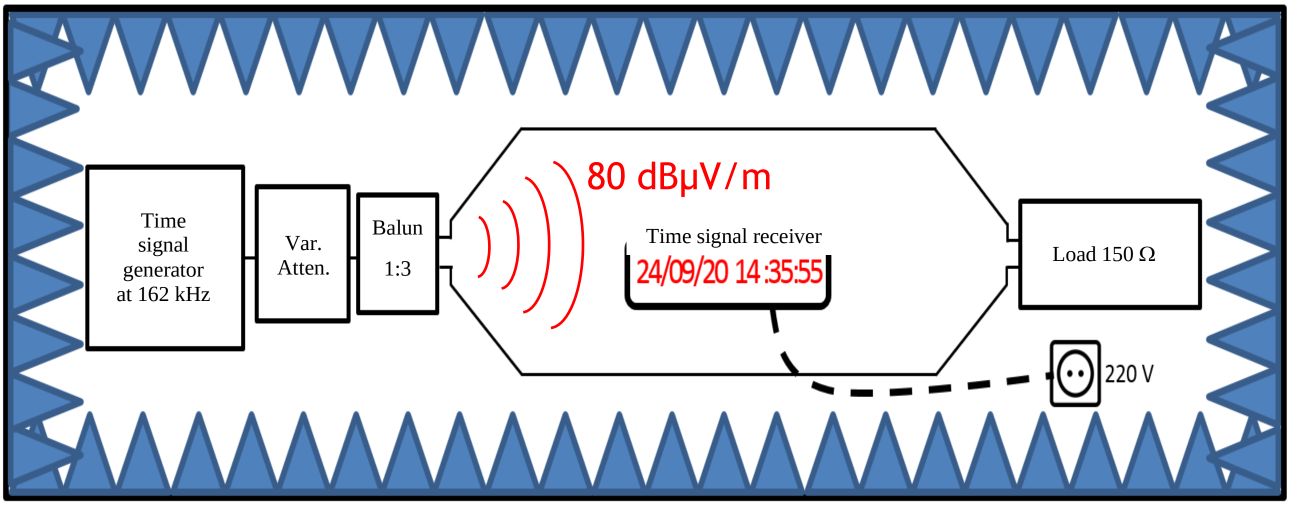


Table 28 summarizes the synchronization times obtained for all the devices tested.

TABLE 28

**Synchronization times for all the devices tested
Maximum synchronization times range from 5 min to 8 min**

Device under test	Reference synchronization time in ideal environment (80 dBµV/m) TO	Maximum considered synchronization time TO + 3 min
Style 5S	2 min	5 min
ALS1 62 demonstrator	2 min	5 min
Style 10SD	2 min	5 min
RCH 218	2 min	5 min
Profil 930	3 min	6 min
Sigma	3 min	6 min
Radiolite 200	3 min	6 min
Cristalys Date	4 min	7 min
Radiolite 110	5 min	8 min

9.3.4 Step 4 – Sensitivity test of radiosynchronization receivers in a noise-free environment and in the absence of misalignment

For each device tested, the minimum detection threshold of the time signal is sought w. r. t. the maximum synchronization time set in the previous step (Table 28), which is the reference sensitivity in an ideal environment (noise-free and in the absence of misalignment of the receiving antennas).

After checking that the noise brought back was insignificant in the useful range and did not risk affecting the measurements, the same procedure is carried out for battery and mains equipment (see Fig. 56).

FIGURE 56

Sensitivity test of radio-synchronisation receivers in a noise-free environment and in the absence of misalignment

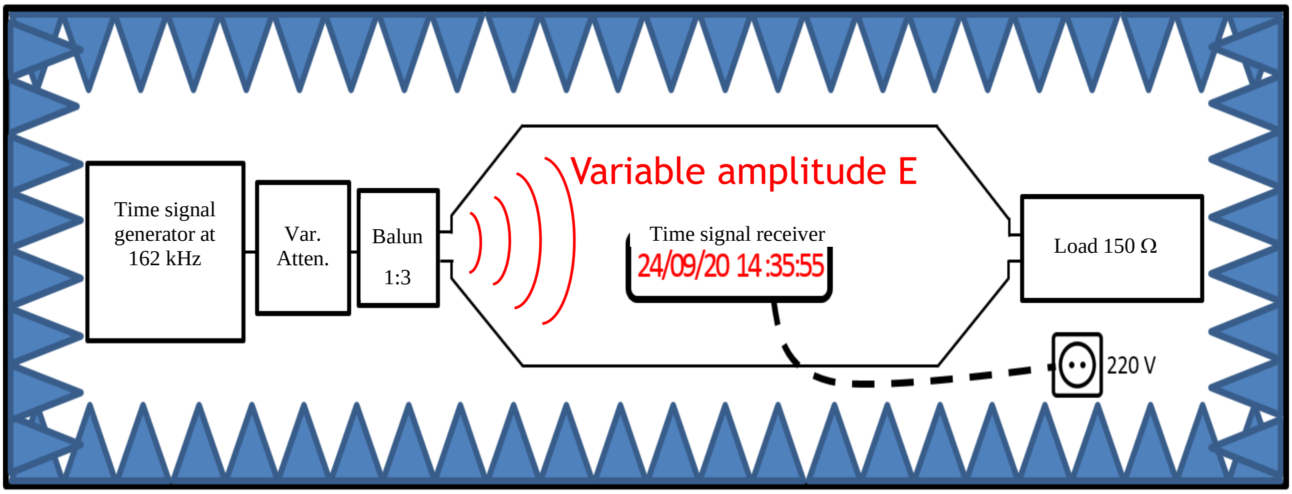


Table 29 summarizes the results obtained for all the devices tested.

TABLE 29

Measurement of the sensitivity threshold for all the devices tested

Device under test	**Sensitivity threshold at TO + 3 min (dBµV/m)**
Sigma	36.3
Radiolite 110	39.7
Radiolite 200	40.9
Profil 930	49.0
Cristalys Date	48.2
Style 5S	50.2
ALS1 62 demonstrator	51.0
RCH 218	68.9
Style 10SD	71.1

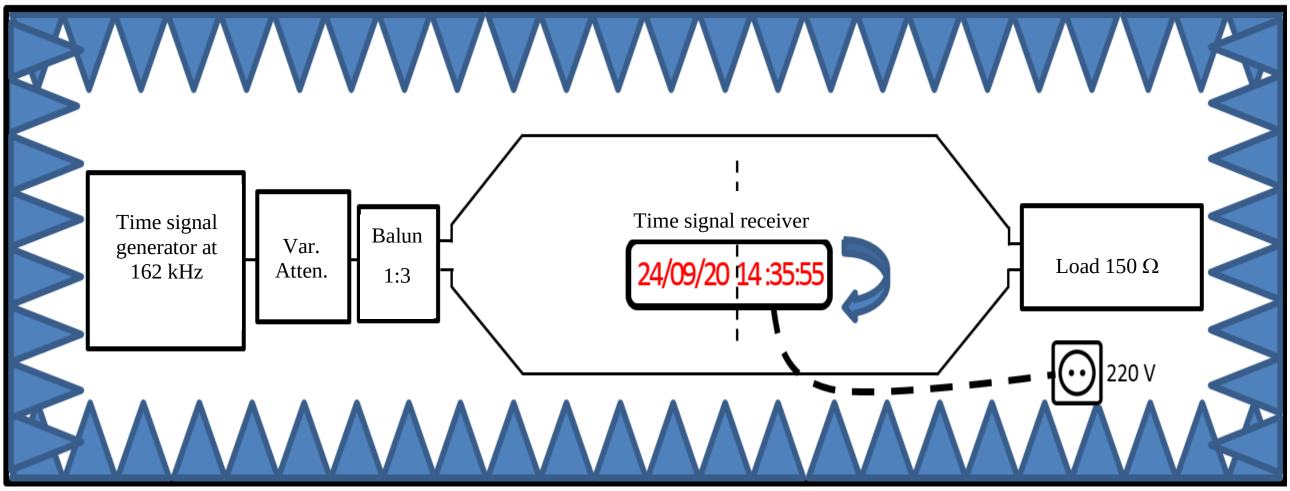
The sensitivity thresholds obtained are highly variable from one receiver to another, with values varying between 36.3 dBµV/m and 71.1 dBµV/m. Among the nine receivers tested, three receivers are significantly more sensitive than the others with a sensitivity less than or equal to 41 dBµV/m, four receivers have an average sensitivity between 48 dBµV/m and 51 dBµV/m and the last two devices have a much lower sensitivity. If all receivers are taken into account, the average sensitivity value is 50.6 dBµV/m (median value of 49 dBµV/m). By discarding the two devices which are clearly the least sensitive, this average sensitivity value drops to 45 dBµV/m (median value at 48.2 dBµV/m).

9.3.5 Step 5 – Sensitivity test of radiosynchronization receivers in a noise-free environment and in the presence of misalignment

The procedure is the same as in the previous step but with a misalignment on the receiver under test (see Fig. 57). Since ferrite antennas are very small relative to wavelength, their radiation pattern is easily predictable and can be approximated fairly accurately by the sine of the azimuth angle. The greater the misalignment, the lower the voltage across the receiving antenna, proportionally degrading the sensitivity of the associated receiver. This step being very time-consuming, a check of the theory / measurement adequacy is simply carried out on a single antenna equipment (ALS162 demonstrator) and on an orthogonal antenna equipment (Cristalys Date). To do this, each device is successively placed in the optimum position in the cell and its sensitivity threshold is recorded. The device under test is then rotated in 10° steps between 0° and 90°, with an additional position at 85° (necessary to accurately assess the radiation dip of the single antenna). A 360° projection is then carried out on the basis of these measurements.

FIGURE 57

Sensitivity test of radiosynchronization receivers in a noise-free environment and in the presence of misalignment



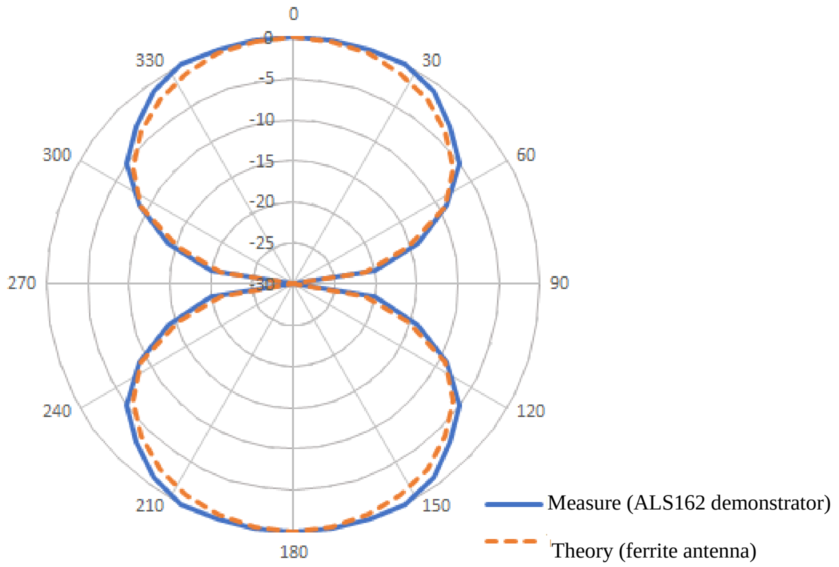
9.3.5.1 Misalignment on a device with a single antenna (ALS162 demonstrator)

Figure 58 represents the measured sensitivity attenuation relative to optimal antenna alignment (ferrite antenna placed perpendicular to the TEM cell). For comparison, the theoretical attenuation of a very small magnetic antenna is also shown in dotted lines.

The correlation observed in theory and measurement is quite acceptable given the fact that the attenuation values tested were varied in 1 dB steps only. The theoretical sinus attenuation, which is a characteristic of magnetic antennas of very small dimensions relative to wavelength, can therefore be used to estimate the receiver desensitization induced by misalignment. The behaviour will be more or less the same for all the rod type ferrite antennas (only the depth of zero can be possibly modified). For this antenna configuration, the receiver desensitization will remain less than or equal to 3 dB for a misalignment less than or equal to 45 degrees. In case of misalignment from the main direction, desensitization can quickly reach several tens of dB. Among all the devices tested, only the Cristalys Date clock is not affected by this topology.

FIGURE 58

Sensitivity attenuation relative to optimal antenna alignment (dB)

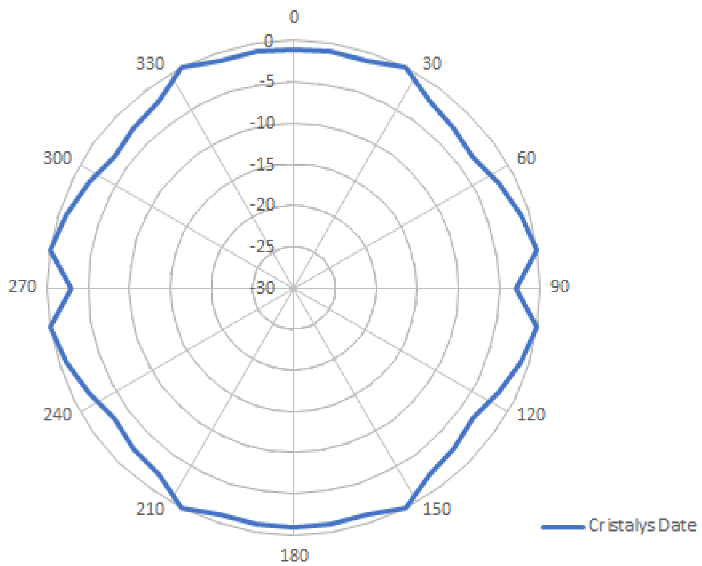


9.3.5.2 Misalignment on devices with orthogonal antennas (Cristalys Date)

The same protocol is applied for a receiver equipped with two orthogonal ferrite antennas. The pattern noted in this case is quasi-omnidirectional, the attenuation noted systematically remaining less than or equal to 3 dB in all directions (Fig. 59). In practice slight differences may be observed from one device to another depending on the accuracy of orthogonality of the two ferrites, the relative phase of the two ports and the signal processing strategies associated with the ports. In any case, the attenuation induced by any misalignment in this case would not exceed 3 dB.

FIGURE 59

Sensitivity attenuation relative to optimal antenna alignment (dB)



9.3.6 Step 6 – Calibration of the noise field with a reference antenna

In order to assess the sensitivity of receivers in a noisy environment (see Fig. 60), the reference time signal is combined with a Gaussian white noise using a noise generator (Noisecom NC6109A). To do so, a combiner is added to the measurement bench allowing simultaneous injection of the time signal and the noise at the input of the TEM cell. With all the measuring bench equipment connected, the noise generator is first switched *on* in the absence of a signal emitted by the time signal generator. The active reference loop thus makes it possible to directly correlate the amplitude of the noise available at the output of the noise generator, with the field level generated in the cell. By adjusting the level of attenuation at the output of the noise generator, the noise field is calibrated to obtain the noise field levels used in the study. The two levels retained (26 dBµV/m and 35.5 dBµV/m) are determined based on Recommendation ITU-R P.372 [1] in Annex 2.

FIGURE 60

Calibration of the noise field with a reference antenna

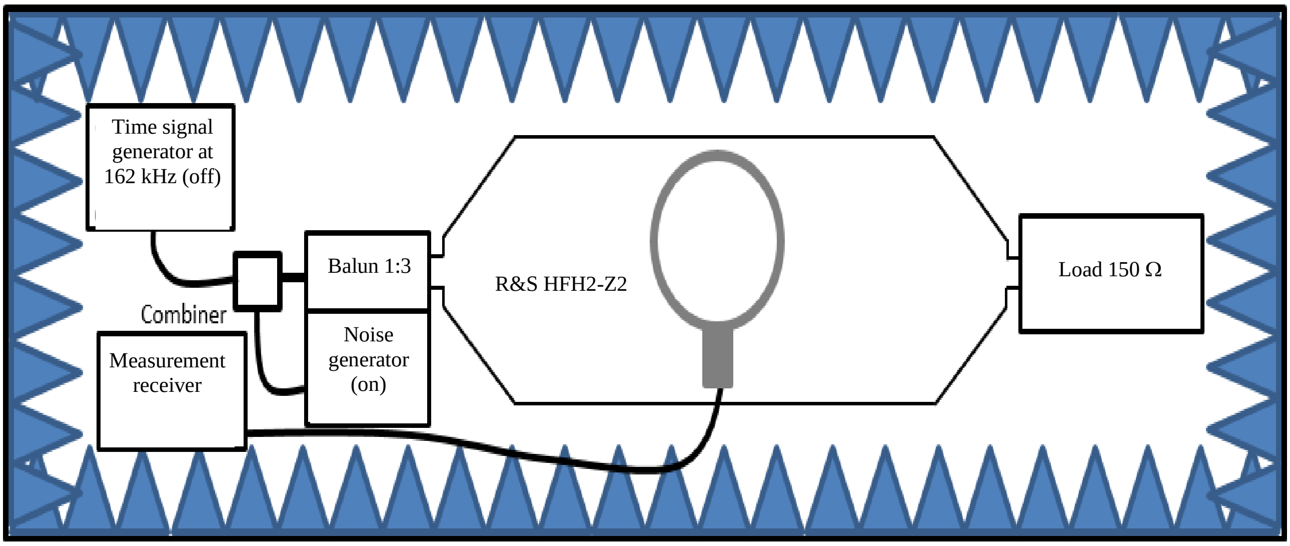


TABLE 30

Calibrated levels of the noise field induced in the TEM cell in 250 Hz bandwidth

Frequency (kHz)	**Attenuation (dB)**	**Noise level (dBm/Hz)**	**Noise field (dBµV/m) induced in the TEM cell over 250 Hz**
162	27	−108.0	26.5
162	17	−98.6	35.7

The variable attenuation of the system being adjusted to the nearest dB, the levels of the noise field given in Table 30 are the values closest to the reference values obtained. These values take into account the bandwidth of the analysis filter ESH3 (200 Hz). As the Gaussian white noise generated is wideband, a 1 dB correction was therefore applied in order to satisfy these levels over 250 Hz.

9.3.7 Step 7 – Sensitivity test of radiosynchronization receivers in a noisy environment and in the absence of misalignment

Once the noise field has been calibrated, the minimum hooking threshold of the time signal is determined for each level of noise emitted by the noise generator (see Fig. 61). In order to make the correlation between dropout thresholds and amplitude of the useful signal obvious with the new measurement bench, a calibration of the useful signal is also carried out (Table 31), while the noise field generator is *off* (in practice, in order to avoid a significant mismatch when the generator is *off*, the equipment is *on* with maximum attenuation at its output (100 dB)).

FIGURE 61

Sensitivity measurement system to all devices under test

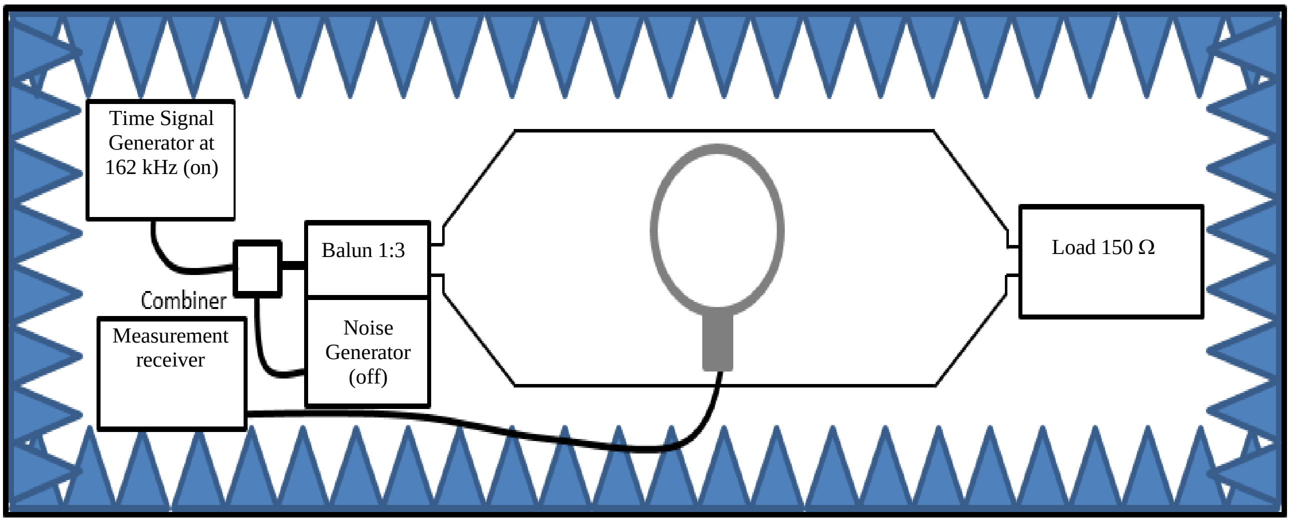


TABLE 31

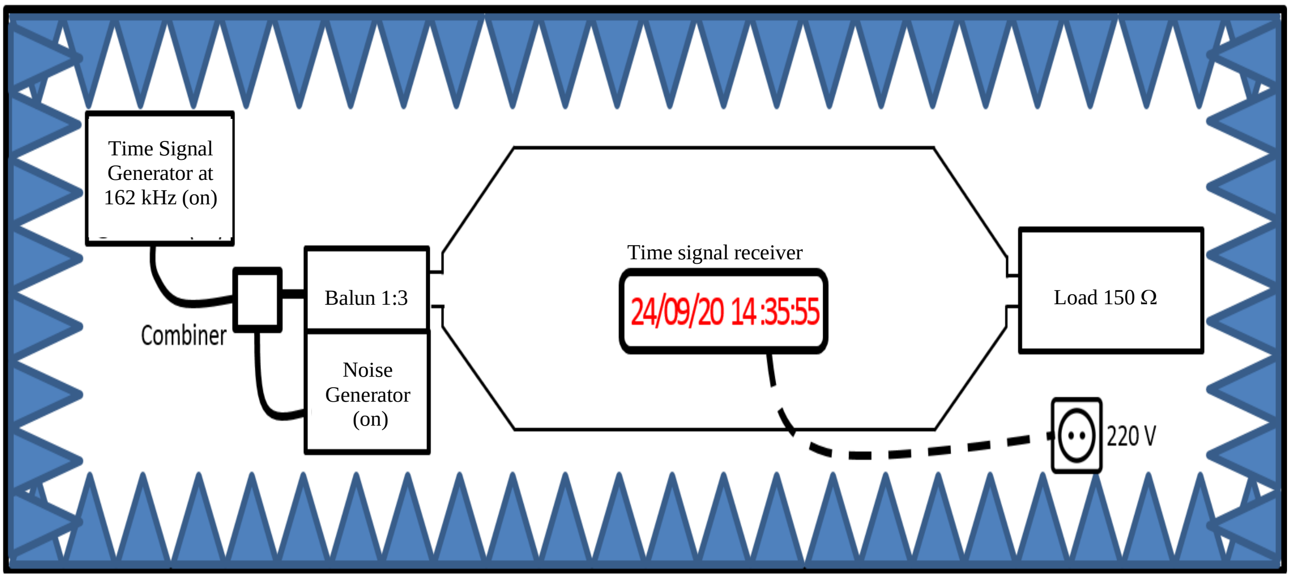
Calibration values of the useful signal in the absence of the noise field (noise generator off)

Frequency (kHz)	Attenuation (dB)	Maximum signal level (dBm/Hz)	E-field (dBμV/m) induced in the TEM cell over 250 Hz
162	0	−49.8	85.9
162	5	−54.7	80.4
162	10	−60.2	75.0
162	15	−65.3	69.9
162	20	−69.9	65.3
162	25	−75.0	59.9
162	30	−80.2	54.8
162	35	−85.3	50.0
162	40	−90.4	44.9
162	45	−95.5	39.6
162	50	−100.4	35.0

Once the calibrations have been carried out, the sensitivity of each receiver under test is measured, for the two selected noise field thresholds, the time signal and noise generators being *on* simultaneously (see Fig. 62).

FIGURE 62

Sensitivity and SNR measurement system of the receivers under test



9.3.7.1 Reference noise field at 26.5 dBµV/m

In the presence of a noise field at 26.5 dBµV/m, the sensitivity threshold is significantly modified for the three most sensitive receivers (Table 32), namely: the master clock Sigma (Bodet) and the two Radiolite 110 and 200 (BH Technologies). The desensitization observed on other receivers is not significant; it remains less than or equal to 1 dB relative to the thresholds obtained in the absence of noise[3].

For the three desensitized receivers, the desensitization observed is sufficiently clear (≥ 3 dB) to be validated without ambiguity. In the presence of the noise field level considered here, the sensitivity thresholds vary between 42.9 dBµV/m and 44.9 dBµV/m against 36.3 dBµV/m to 40.9 dBµV/m in the absence of noise, with an average value and a median value equal to 43.9 dBµV/m. The measured SNR thus changes between 16.4 dB and 18.4 dB, with an average value and a median value equal to 17.4 dB.

TABLE 32

Sensitivity and SNR values of the receivers in the presence of a noise field at 26.5 dBµV/m

Device under test	Noise field in the TEM cell over 250 Hz (dBµV/m)	Useful signal field in the TEM cell at the limit of sensitivity (dBµV/m)	SNR (dB)	Sensitivity threshold at TO + 3 min (dBµV/m)
Sigma	26.5	42.9	16.4	36.3
Radiolite 110	26.5	44.9	18.4	39.7
Radiolite 200	26.5	43.9	17.4	40.9

[3] Depending on receivers, the repeatability of the sensitivity threshold is sometimes difficult to achieve at dB level.

TABLE 32 (*end*)

Device under test	Noise field in the TEM cell over 250 Hz (dBµV/m)	Useful signal field in the TEM cell at the limit of sensitivity (dBµV/m)	SNR (dB)	Sensitivity threshold at TO + 3 min (dBµV/m)
Profil 930	26.5	50.0	Not desensitized	49.0
Cristalys Date	26.5	40.0	Not desensitized	48.2
Style 5S	26.5	51.0	Not desensitized	50.2
ALS1 62 demonstrator	26.5	51.0	Not desensitized	51.0
RCH 218	26.5	68.9	Not desensitized	68.9
Style 10SD	26.5	71.1	Not desensitized	71.1

9.3.7.2 Reference noise field at 35.7 dBµV/m

In the presence of a noise field at 35.7 dBµV/m, clear desensitization is observed on all receivers except RCH218 and Style 10SD receivers (Table 33). Among the seven desensitized receivers, the sensitivity thresholds, in the presence of this noise level, vary between 51 dBµV/m and 53.8 dBµV/m against 36.3 dBµV/m and 51 dBµV/m in the absence of noise, with an average value of 53.1 dBµV/m and a median value of 53 dBµV/m. The measured SNR varies between 15.3 dB and 18.1 dB, with an average value and a median value equal to 17.3 dB.

TABLE 33

Sensitivity and SNR values of the receivers, in the presence of a noise field at 35.7 dBµV/m

Device under test	Noise field in the TEM cell over 250 Hz (dBµV/m)	Useful signal field in the TEM cell at the limit of sensitivity (dBµV/m)	SNR (dB)	Sensitivity threshold at TO + 3 min (dBµV/m)
Sigma	35.7	51.0	15.3	36.3
Radiolite 110	35.7	52.8	17.1	39.7
Radiolite 200	35.7	53.8	18.1	40.9
Profil 930	35.7	53.8	18.1	49.0
Cristalys Date	35.7	53.0	17.3	48.2
Style 5S	35.7	53.8	18.1	50.2
ALS1 62 demonstrator	35.7	53.0	17.3	51.0
RCH 218	35.7	68.9	Not desensitized	68.9
Style 10SD	35.7	71.1	Not desensitized	71.1

9.3.8 Step 8 – Sensitivity test of radiosynchronization receivers in a noisy environment and in the presence of misalignment

For the same reasons as mentioned in Step 5, this step was not carried out. To estimate the additional receiver desensitization induced by the misalignment, it is referred to the results presented in § 9.3.5.

9.4 Summary

A measurement campaign has been carried out on a sample of nine Allouis receivers, representative of the fleet currently deployed throughout France, to determine their sensitivity as well as their signal to noise plus interference ratio (SNIR).

The sensitivity of the tested Allouis receivers varies from 36.3 dBµV/m to 51 dBµV/m. Note that the two receivers having a sensitivity much lower than that of the rest of the receivers were not taken into account. Nevertheless, these two receivers are no longer or little deployed in France.

As for the SNR of the receivers, it varies from 16.4 dB to 18.4 dB for a noise level of 26.5 dBµV/m and, from 15.3 dB to 18.1 dB for a noise level of 35.7 dBµV/m respectively.

On the basis of these results, the following values are defined to guarantee the protection of the reception of the Allouis network deployed in France: The sensitivity and the minimum usable field strength for proper reception of the ALS162 signal is 50 dBµV/m. As the tolerable SNIR was found to 21 dB, a level of interference of about 30 dBµV/m is permitted. The SNIR of 21 dB consists of 18 dB of typical value plus 3 dB of margin to compensate for the cumulative measurement error at the different measurement steps.

Finally, it is important to stress that the sensitivity defined above should in no case be equated with the level of the useful signal for determining the Allouis coverage area, since it does not include any coverage margin.

References

[1] Recommendation ITU-R P.372, *Radio Noise*, 08-2019.

[2] A.C Fraser-Smith, *The effective antenna noise figure Fa for a vertical loop antenna and its application to extremely low frequency/very low frequency atmospheric noise*, Radio Science, Vol. 42, 2007.

Annex 1
to Section 9

List of equipment used in measurements

TABLE 34

List of equipment used in measurements

Designation	Type
Time Signal Generator at 162 kHz	Bodet GENE-ALS162
Measurement Receiver	R&S ESH3
Spectrum Analyzer	Agilent E4446A
RF signal generator	HP 8648D
Fieldmeter	Narda EHP-200
Noise generator	Noisecom NC6109A
Active shielded magnetic loop	ETS-Lindgren 6502
Active magnetic loop	R&S HFH2-Z2
Passive loop antenna	TDF

Annex 2
to Section 9

Calculations used to determine the noise field strength in Step 7

The following calculations are based on the data from Recommendation ITU-R P.372 [1].

a) Noise field linked to keraunic activity

According to Figs 13a to 36b of Recommendation ITU-R P.372 [1], the median values of the mean power factor of atmospheric noise related to keraunic activity vary between 71 dB and 113 dB (Table 35), depending on the Time of the Day and Season, in Metropolitan France, at 162 kHz.

TABLE 35

Median values of the average power factor of radio noise related to keraunic activity in France, at 162 kHz and 1 MHz

	Winter		**Spring**		**Summer**		**Autumn**	
	***Fa* (dB) at 1 MHz**	***Fa* (dB) at 162 kHz**	***Fa* (dB) at 1 MHz**	***Fa* (dB) at 162 kHz**	***Fa* (dB) at 1 MHz**	***Fa* (dB) at 162 kHz**	***Fa* (dB) at 1 MHz**	***Fa* (dB) at 162 kHz**
0 h – 4 h	65/70	100/105	25/30	71/76	65/70	102/107	70/75	105/110
4 h – 8 h	60/65	95/100	35	77	45/55	90/100	60	98
8 h – 12 h	25/30	71/76	30	78	30/35	82/87	35	83
12 h – 16 h	35	77	35/40	80/85	35/45	83/93	40/45	87/92
16 h – 20 h	55/60	88/93	50/60	87/97	50/65	90/105	60/70	97/107
20 h – 24 h	60/65	94/99	65/70	100/105	65/75	103/113	75	109

The noise field calculation is performed in Recommendation ITU-R P.372 [1] by considering a vertical monopole antenna above a perfectly conducting plane or an isotropic antenna. In this study, the configuration used is not strictly in either of these two cases. The case of a magnetic antenna above a perfectly conducting plane is dealt with in [2] but does not however apply to directions perpendicular to the ferrite or to the orthogonal ferrite systems present in some devices (to avoid misalignment). In the absence of relevant data for the treated case, the calculations are limited to those proposed in [1] by considering the case of a short vertical monopole reference antenna placed above a perfectly conducting plane. It is this type of antenna that was used to establish the reference levels of atmospheric noise in Figs 13a to 36b.

As a first approximation, it can therefore be considered that the noise field associated with a given external noise factor Fa at the frequency f and for a bandwidth B = 10log (b) is such that:

$$En = Fa + 20 \log f(\text{MHz}) + B - 95.5 \qquad (\text{dB}\mu\text{V/m})$$

For a bandwidth of 250 Hz (Allouis signal bandwidth), this would therefore give a median noise field related to keraunic activity of approximately between −16 dBμV/m and 26 dBμV/m, for Fa values of 71 dB and 113 dB respectively.

If the values reached 10% of the time and not the median values are considered, from Figs 13c to 36c, an increase of about 9.5 dB can be expected with regard to the median values. In this case, a noise field between −6.5 dBμV/m and 35.5 dBμV/m is obtained.

b) Man-made noise field

Unlike atmospheric noise field, the man-made noise field is unfortunately difficult to predict in the low frequency bands. This is explained in particular by the fact that the sources of man-made noise are highly variable and constantly changing and that the associated noise field is very strongly dependent on the distance from the source.

However, at the latitudes concerned here, for the majority of season / time of the day slots, the man-made noise factor in a quiet environment (76 dB at 162 kHz, see Figs 13b to 36b of [1]) is clearly lower than the atmospheric noise linked to the keraunic activity at the frequency of use.

The man-made noise figure below 300 kHz is unfortunately not established in this recommendation for urban type environments (see Fig. 39 of [1]). But, as a first approximation, it can however be considered that this man-made noise figure would be at least 24 dB higher than the values observed in quiet rural areas (see also Fig. 39). Based on this assumption, the man-made noise figure in an urban environment is at least 100 dB, which is significantly lower than 113 dB observable at night in summer for atmospheric noise related to lightning.

10 Conclusion

In the Radio Regulations (No. **1.53**) a SFTS service has been defined as “a radiocommunication service for scientific, technical and other purposes, providing the transmission of specified frequencies, time signals, or both, of stated high precision, intended for general reception.” Provision **26.4** invites administrations to cooperate in reducing interference in the frequency bands to which the SFTS service is allocated.

In this Report, information on the various dedicated SFTS services in the frequency band 40 kHz – 162 kHz and other frequencies in China and the Russian Federation have been compiled. The services are provided by institutes entrusted with the dissemination of legal time in their respective countries and each of them serve millions of users. See Recommendation ITU-R TF.583-6 – Time Codes, for an up-to-date compilation of the codes used for time dissemination, and Recommendation ITU-R TF.768 – Characteristics of standard-frequency and time-signal emissions in allocated bands and characteristics of stations emitting with regular schedules with stabilized frequencies, outside of allocated bands, for a full list of SFTS services.

Very different kinds of equipment are utilized for the reception of the respective signals, from sophisticated receivers as part of industrial or scientific installations to low-cost receivers for private use. In this Report, the result of theoretical studies as well as experimental findings on reception conditions have been compiled. A matter of concern is the expected increase of WPT-EV installations which are proposed to use frequencies falling on or close to the frequencies used in the SFTS services. The field strength of signals transmitted from WPT-EV installations that are currently permitted to be 10 m from the installation will seriously disturb the reception and decoding of SFTS service. Such disturbance could be mitigated by restricting the transmission power of the WPT-EV charging installation and carefully selecting its operational frequency or by other mitigation techniques, e.g. interrupting the charging process at certain hours of the day.