

Transient behaviour of a data transmission system using FSK-modulation and coupled inductive loops

KARL HEINZ KRAFT

Faculty of Electrical Engineering, Ostfalia University of Applied Sciences,
Salzdahlumer Str. 46-48
D-38302 Wolfenbüttel,
GERMANY

Abstract: - Safety and efficiency of modern transportation systems require the intensive use of data transmission devices. They should operate in both directions between on-board and wayside equipment. Especially, high-speed applications require a suitable data transmission rate. The design of the entire transmission path has to consider the input and output signals of all the involved components, investigated here for the modulation method FSK. It is demonstrated how the transient behaviour can be simulated using SPICE subcircuits. The main components to be modelled are the following: modulator, a fixed and a movable inductive loop, connected to an input and an output filter, amplifier and limiter, and demodulator.

Key-Words: - Moving vehicles, data transmission, FSK signals and their spectra, modulation and demodulation, inductive loops, band-pass filters, circuit simulation, SPICE-models and subcircuits, transient analysis.

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1 Introduction

The importance of data transmission equipment as part of traffic control systems is well-known, for example, with a view to high-speed railways or urban traffic. Similar applications lie in the field of RFID and NFC, e.g., [1], [2]. This contribution deals with a transmission system based on coupled inductive loops and the modulation method FSK (Frequency Shift Keying; e.g., [3], [4], and with a listing of application fields in [5], [6]. From a historical point of view, [7] is interesting, dealing with early basics of data transmission, e.g., AC telegraphy. Here, various aspects and components have to be investigated in detail: The transmitter with data-input and generation of FSK signals, discussion of their frequency spectra, and a band-pass filter for limitation of the used bandwidth; the coupled inductive loops with a mutual inductance, depending on the position and the time, because of the moving vehicle; a band-pass on the receiving end for reduction of noise and interference; the receiver with demodulator and data-output. The filter design has to consider the properties of coupled loops, as discussed in detail in [8]. The influence of all the mentioned components on the output signal is essential for the system design.

2 Problem formulation

As described above, each component has to be investigated to define a circuit model or subcircuit. Figure 1 shows a block diagram containing all the components to be modelled.

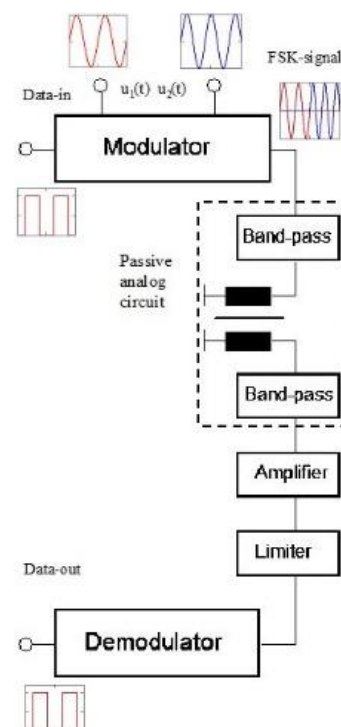


Fig. 1: Block diagram of the system
Source: created by the author.

The investigated transmission system starts with a FSK signal generated in principle by the multiplication of an analog sinusoidal signal by a binary code signal. The modulated signal is transmitted via two coupled inductive loops, which are integrated into a transmitting band-pass and a receiving band-pass. It should be emphasized that one of the two loops is movable. The received band-limited signal has to be amplified and demodulated, e.g., by a product demodulator.

A simulation of the entire circuit would be advantageous; the use of a program such as SPICE offers itself. An important task will be to organize the collaboration of the mentioned components. While the properties of the components are at least partly known, lies the main problem in the complexity of the system. This contribution aims to show that it is possible to simulate the transient behaviour of the entire system.

3 FSK spectra

The properties of the modulation method of Frequency Shift Keying (FSK) have been investigated by many projects and have been reported in numerous papers. However, in this field, it seems to be plausible to repeat and outline some basic relations because the matter is rather complex and has several features. For the design of the transmission system, some important properties should be taken into account. As a first step, the calculation of the frequency spectrum presents itself. The generated signal is given by (1) or (2):

$$u_T(t) = \begin{cases} \hat{u}_T \cdot \sin(2 \cdot \pi \cdot f_1 \cdot (t + T_{bit})) & \text{for } -T_{bit} \leq t < 0 \\ \hat{u}_T \cdot \sin(2 \cdot \pi \cdot f_2 \cdot (t - T_{bit})) & \text{for } 0 \leq t \leq T_{bit} \end{cases} \quad (1)$$

as part of a periodical 1-0-1-0-sequence with the period $T_p = 2 \cdot T_{bit}$. It can be written alternatively by

$$u_T(t) = \hat{u}_T \cdot \sin[2 \cdot \pi \cdot (f_c - \Delta f + 2 \cdot \Delta f \cdot c_o(t))] \quad (2)$$

with the code signal $c_o(t)$ and the value “0” or “1”. The length T_{bit} of a transmitted bit should be suited to the numbers N_1 and N_2 of complete periods of the generated frequencies f_1 and f_2 , $N_1 \cdot 1/f_1 = N_2 \cdot 1/f_2 = T_{bit}$, to avoid unnecessary voltage steps. Furthermore, the following definitions and parameters have to be mentioned: centre frequency $f_c = (f_1 + f_2)/2$, frequency deviation $\Delta f = f_c - f_1 = f_2 - f_c$, data transmission rate $r_D = \text{bit}/T_{bit}$, modulation index $h = 2 \cdot \Delta f \cdot T_{bit} = \Delta f / T_p$. The choice of the modulation index is essential for the shape of the discrete amplitude spectrum and, therefore, as well, for the required bandwidth; (already mentioned in [7]). The general approach is the Fourier integral

$$\underline{A}(k) = \frac{2}{T_p} \cdot \int_{-T_{bit}}^{T_{bit}} u_T(t) \cdot e^{-j \cdot 2 \cdot \pi \cdot k \cdot t / T_p} dt \quad (3)$$

The magnitude spectrum $A(k) = |\underline{A}(k)|$ can, of course, be calculated numerically, but an analytical solution is possible with some manipulations. The variable k ($k=0, 1, 2, \dots, f_c \cdot T_p \dots$) may be replaced by $m = k \cdot f_c \cdot T_p$ (normalization, $m = \pm 3, \pm 2, \pm 1, 0$; the product $f_c \cdot T_p$ is supposed to be an integer number). This substitution is reasonable, because $k=0$ means $f=0$, which is not relevant here, while $m=0$ means $f=f_c$, $k=23$ with $f_c=2.3$ MHz, $\Delta f=100$ kHz, and lies in the centre of the used frequency range. The evaluation of (3) gives the intermediate result (4), (5) after some extensive manipulations.

$$\text{Re}(\underline{A}(m)) = \frac{\hat{u}_T}{\pi} \cdot \frac{h}{|h^2 - m^2|} \cdot (\cos(m \cdot \pi) - \cos(h \cdot \pi)) \quad (4)$$

$$\text{Im}(\underline{A}(m)) = \frac{\hat{u}_T}{\pi} \cdot \frac{h}{|h^2 - m^2|} \cdot (-\sin(h \cdot \pi)) \quad (5)$$

The rather simple formula (6) can be determined from (4) and (5) by calculating the magnitude:

$$A(m, h) = \frac{\hat{u}_T}{\pi} \cdot \frac{2 \cdot h}{|h^2 - m^2|} \cdot \left| \sin\left(\frac{\pi}{2} \cdot (h - m)\right) \right| \quad (6)$$

Alternatively, (7) or (8) may be used.

$$A(m, h) = \frac{\hat{u}_T}{\pi} \cdot \frac{2 \cdot h}{|h^2 - m^2|} \cdot \left| \sin\left(\frac{\pi}{2} \cdot h\right) \right| \quad \text{if } m=0, \pm 2, \dots$$

$$A(m, h) = \frac{\hat{u}_T}{\pi} \cdot \frac{2 \cdot h}{|h^2 - m^2|} \cdot \left| \cos\left(\frac{\pi}{2} \cdot h\right) \right| \quad \text{if } m=\pm 1, \pm 3, \dots \quad (7)$$

$$A(m, h) = \frac{\hat{u}_T}{\pi} \cdot \frac{h}{|h^2 - m^2|} \cdot |\cos(m \cdot \pi) - \cos(h \cdot \pi)|, \quad m \neq h, \text{ with } A(m=h)=0.5. \quad (8)$$

A very often used case is $h \approx 1$, the range of applications lies between $h=0.5$ and $h=1.1$. A complete example is given by the following values: $T_p=10 \mu\text{s}$, $f_p=1/T_p=100$ kHz, $T_{bit}=5 \mu\text{s}$, $r_D=200$ kbit/s, $f_c=2.3$ MHz, $f_1=2.2$ MHz, $f_2=2.4$ MHz, $\Delta f=100$ kHz, $h=1$, see also, [8]. The spectrum, depending on the variable m , is shown in Figure 2, together with the plots for $h=0.6$ and $h=1.2$. The discrete spectrum for $h=1$ has been supplemented by the filter curve $F(f)$ of a fictional band-pass of 4th order to estimate the required bandwidth, which should be $B \approx 200$ kHz for the selected example. The absolute frequency is $f=f_c+m \cdot f_p$. The mentioned transfer function can be derived, for example, via a normalized reference low-pass (Butterworth) with $F(p_L) = b_0 / (p_L^2 + \sqrt{2} \cdot p_L + 1)$ and use of the frequency transformation rule $p_L \rightarrow Q \cdot (\tilde{p} + 1 / \tilde{p})$ with $\tilde{p} = p / \omega_c$

and $Q=f_c/B$; here $Q=11.5$; b_0 is a suitable DC-factor.

The result can be seen in (9):

$$F(\tilde{p}) = \frac{b_0 / Q^2 \cdot \tilde{p}^2}{\tilde{p}^4 + \sqrt{2} / Q \cdot \tilde{p}^3 + (2 + 1 / Q^2) \cdot \tilde{p}^2 + \sqrt{2} / Q \cdot \tilde{p} + 1} \quad (9).$$

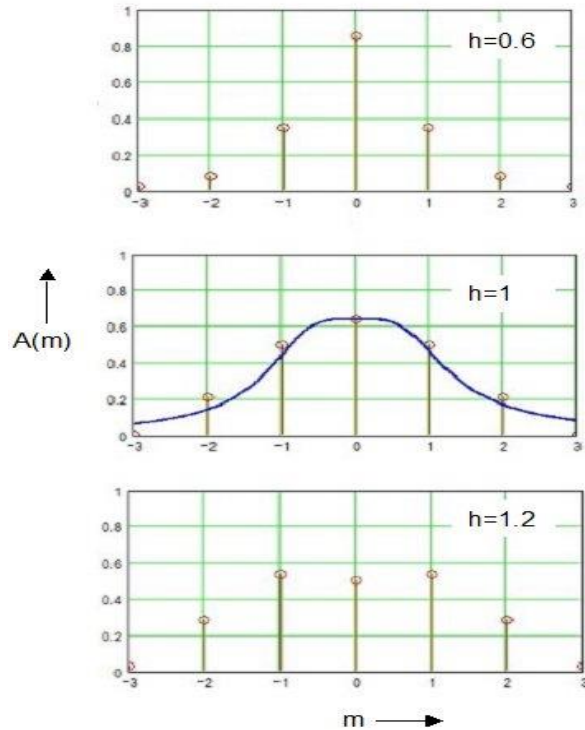


Fig. 2: Amplitude spectrum for a periodical FSK signal with various modulation indices
Source: created by the author.

4 Band-pass filter

The design of the two band-pass filters for the transmitter and the receiver has been treated in detail in [8], where the properties of the coupled inductive loops have been taken into account as well. The example mentioned above has also been used for the filter design. Figure 3 shows an equivalent circuit for the analog part of the transmission system.

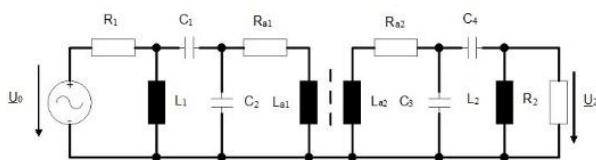


Fig. 3: Schematic of the coupled filters
Source: created by the author.

The calculated values of the circuit elements are:
 $L_1=L_2=10.5 \mu\text{H}$, $C_1=C_4=586 \text{ pF}$, $C_2=C_3=2.07 \text{ nF}$ for the band-pass elements, $L_{a1}=L_{a2}=1.8 \mu\text{H}$, $R_{a1}=R_{a2}=0.25 \Omega$ and the max. mutual inductance $L_{12}=0.09 \mu\text{H}$ for the coupled inductive loops (each

square loop with the length $a=40 \text{ cm}$, their distance $d=25 \text{ cm}$). The max. coupling factor is $k_0=0.05$, the dependency $k(x)$, $k(t)$ respectively, has to be discussed in this paper. The resulting filter curve can be seen in Figure 4 (taken from [8], the realized bandwidth is $B \approx 200 \text{ kHz}$ with $R_1=R_2=50 \Omega$).

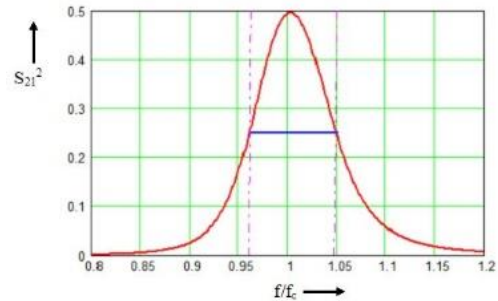


Fig. 4: Frequency-dependent power transfer ($f_c=2.3 \text{ MHz}$, $S_{21}^2=P_2/P_{1\text{max}}$).
Source: created by the author.

While methods of the frequency domain are necessary for the filter design, the transient behaviour of the whole system is to be analysed, with respect to the mentioned filter design.

5 Time-dependent coupling factor

Usually, one loop will be fixed, while the other one is movable. The coupling factor k , as well as the mutual inductance L_{12} , depends, therefore on the position of the moving loop, with the consequence that the parameter L_{12} is time-dependent. For transient analysis, it is necessary to determine the functions $L_{12}=f(x)$, $x=v_0 \cdot t$, and finally $L_{12}=f(t)$ with the speed v_0 , e.g., of a moving vehicle. The mutual inductance L_{12} can rather easily be determined by using a simple measurement set-up (Figure 5). The voltage transfer function $F(p)=U_2(p)/U_0(p)$ (10) with the measured voltages $u_0(t)$ and $u_2(t)$ delivers L_{12} with the known element values R_1 , R_2 , L_1 , L_2 and the frequency f .

$$F(p) = \frac{p \cdot L_{12} / R_1}{p^2 \cdot \frac{L_1 \cdot L_2 - L_{12}^2}{R_1 \cdot R_2} + p \cdot \left(\frac{L_1}{R_1} + \frac{L_2}{R_2} \right) + 1} \quad (10)$$

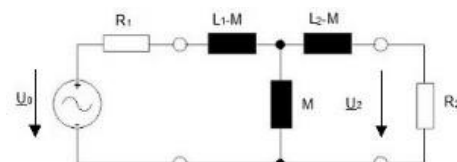


Fig. 5: Equivalent circuit for the coupling of two inductive loops
Source: created by the author.

With $R=R_1=R_2$, $L=L_1=L_2$ and $L_{12}^2 \ll L^2$, the simple result is $L_{12}=2 \cdot L \cdot U_2(f_0)/U_0$, using the frequency $f_0=R/(2 \cdot \pi \cdot L)$. With $a=40$ cm, $h=25$ cm, and with various positions x , $L_{12}(x)$ will be found as shown in Figure 6.

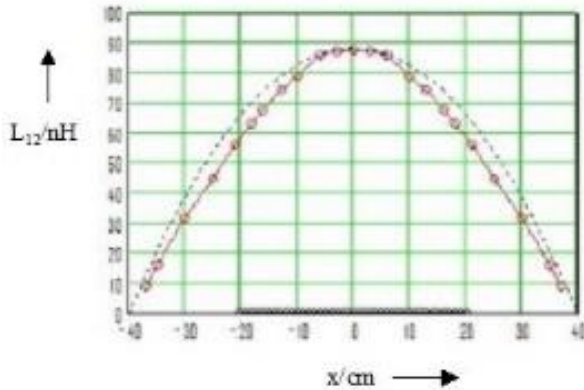


Fig. 6: The mutual inductance depending on the position x ; the centre at $x=0$, $a=40$ cm, $h=25$ cm
Source: created by the author.

A very simple approximation (11) would be

$$L_{12}(x) \approx L_{12m} \cdot \left(1 - \left(\frac{|x|}{a}\right)^2\right) \quad (11)$$

with the maximum value L_{12m} at the centre $x=0$

(dotted line in Figure 6). Three typical configurations are depicted in Figure 7. The ranges $x < -a$ and $x > a$ are neglected here, because the induced magnetic field has only a very small effect.

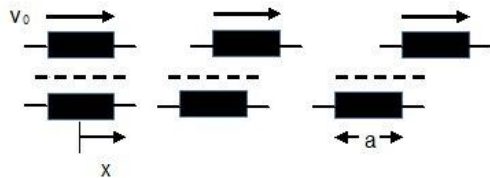


Fig. 7: Three symbolic loop configurations
Source: created by the author.

The time domain analysis of the circuit of Figure 5 gives us two differential equations (12), (13) using the input current $i_1(t)$ and the output current $i_2(t)=u_2(t)/R_2$.

$$R_1 \cdot i_1 + L_1 \cdot \frac{di_1}{dt} - \frac{dL_{12}}{dt} \cdot i_2 - L_{12}(t) \cdot \frac{di_2}{dt} = u_0(t) \quad (12)$$

$$L_{12}(t) \cdot \frac{di_1}{dt} + \frac{dL_{12}}{dt} \cdot i_1 - R_2 \cdot i_2 - L_2 \cdot \frac{di_2}{dt} = 0 \quad (13)$$

The influence of the term $dL_{12}/dt \cdot i_1$ can be estimated by comparison with $L_{12}(t) \cdot di_1/dt$ and assuming a certain current $i_1(t)=i_{1p} \cdot \sin(\omega \cdot t)$. A numerical analysis shows that the named term with dL_{12}/dt

can be neglected here. The function $L_{12}(t)$ has to be integrated into the SPICE-simulation of the passive analog part of the transmission system.

6 Subcircuits

6.1 Generation of a FSK signal

This task is easy, only two rectangular signals $u_{R1,2}(t)$ with period $T_p=2 \cdot T_{bit}$ and two sinusoidal signals $u_{S1,2}(t)$ with frequencies f_1 and f_2 are needed.

Their combination (14) with $u_{S1}(t)=u_{Sp} \cdot \sin(\omega_1 \cdot t)$ and $u_{S2}(t)=u_{Sp} \cdot \sin(\omega_2 \cdot (t-T_{bit}))$ produces the signal $u_{FSK}(t)$:

$$u_{FSK}(t) = u_{R1}(t) \cdot u_{S1}(t) + u_{R2}(t) \cdot u_{S2}(t) \quad (14)$$

For the first period, there is $u_{R1}=1$, $u_{R2}=0$ for $0 \leq t < T_{bit}$ and $u_{R1}=0$, $u_{R2}=1$ for $T_{bit} \leq t < 2 \cdot T_{bit}$. The condition $u_{S1}(T_{bit})=u_{S2}(T_{bit})$ is necessary to avoid signal distortions, therefore $N_1/f_1=N_2/f_2=T_{bit}$ with $N_{1,2}$ as periods of the sinusoidal signals. An example which will be used later, as well: $r_D=200$ kbit/s, $T_{bit}=5$ μ s, $f_1=2.2$ MHz, $f_2=2.4$ MHz, $N_1=11$, $N_2=12$.

Eq. (14) is the key for modelling a subcircuit as shown in Figure 8, with external pulse sources and pins 1 to 4, and using two voltage-controlled sources (type EVALU) for the sinusoidal signals with frequencies f_1 and f_2 . Each signal is multiplied by the belonging rectangular voltage, defined above; u_{R2} is inverted to u_{R1} ($u_{R2} = \bar{u}_{R1}$). The output voltages of E1 and E2 are added up using the Element ESUM.

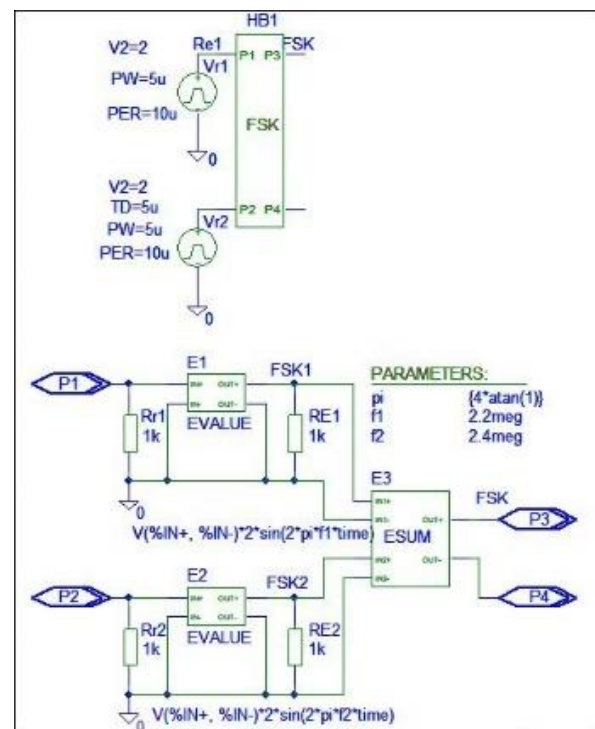


Fig. 8: Subcircuit of the FSK-generator
Source: created by the author.

The result can be seen in Figure 9, presenting the desired FSK signal for one period, here with a “1”-“0” bit combination. The crucial time step at the transition from $f=f_1$ to $f=f_2$ has been marked.

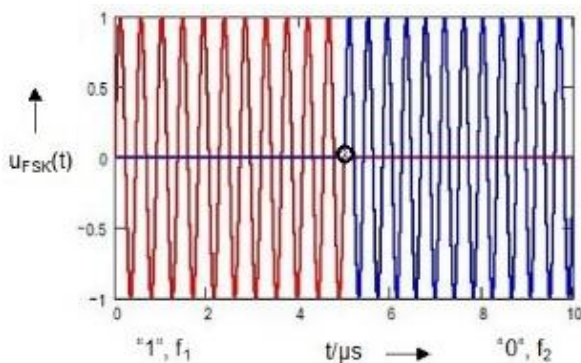


Fig. 9: Generated FSK signal $u_{FSK}(t)$
(1-0-combination, with the marked crucial point).
Source: created by the author.

6.2 The passive analog circuit with band-passes and inductive loops

The values of the band-pass elements can be taken directly from Ch. 4 and Figure 3, but the simulation of the time-dependent mutual inductance is complicated. Figure 10 shows a test circuit with the named components. The subcircuit called $L_{12}(t)$, based on the schematics version, contains several controlled sources (types F, EMULT, ETABLE) and a pulse source required for the time-voltage transformation, producing a linear dependency between time and voltage. The function $L_{12}(x)$ is realized by ETABLE with several pairs of values; the other input of the multiplier is fed by the reference inductance, which equals here the value L_{12m} , and the mirrored input current from the main circuit. Both resistors are required for formal reasons. If the term $dL_{12}/dt \cdot i_2$ (see (12)) should be necessary, then an additional resistor $R_1(t)=dL_{12}/dt$ in series to $L_{12}(t)$ can be used, built by the same method as for $L_{12}(t)$, e.g., replacing $M=L_{12}$ in Figure 5.

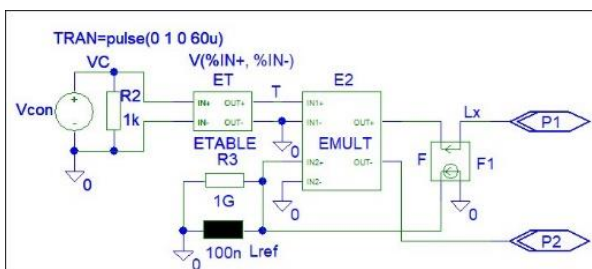


Fig. 10: Subcircuit for $L_{12}(t)$
Source: created by the author.

A schematic built with SPICE elements is presented in Figure 11, and the simulation results

(transient analysis can be seen in Figure 12, using the values from Ch. 4, 5, and 6.1, about Figure 11. This intermediate result serves the purpose to check the modelling of the FSK-generator in combination with the passive part of the transmission system, including $L_{12}(t)$.

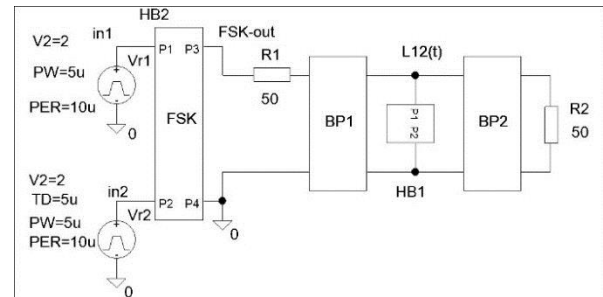


Fig. 11: Test circuit with SPICE subcircuits
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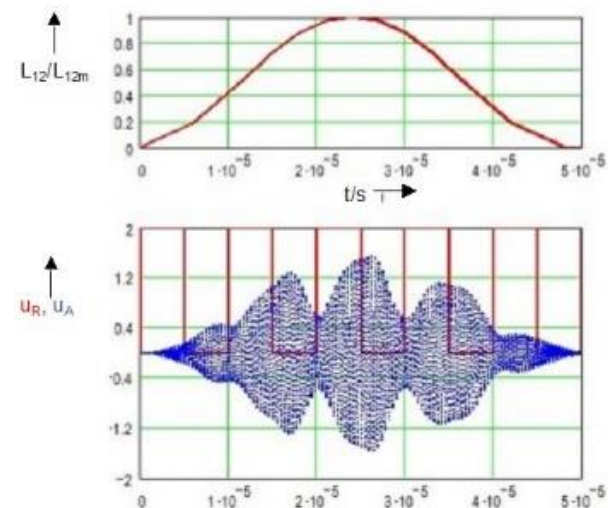


Fig. 12: Normalized mutual time-dependent inductance, data-input $u_R(t)$ and output-signal $u_A(t)$
Source: created by the author.

The test is successful, as can be seen in Figure 12 with three important signals. The transient time at the output (BP 2) is about 5 μs , in accordance with the bandwidth of $B=200$ kHz. A general treatment of time-bandwidth relations can be found, e.g., in [9]. The next question should be whether the demodulator can handle the shown analog output signal $u_A(t)$.

6.3 Demodulator, data-output, and additional measurement results

The aim here is not to create a novel demodulator circuit, but to show how SPICE subcircuits for system simulation can be modelled. The function of the demodulator will be modelled in the sense of ABM (Analog Behaviour Modelling; e.g., [10]). Each component can be represented by a standard SPICE element. The main components of the SPICE-option

ABM are the controlled sources E and G, for example, with the extensions ELAPLACE, EVALUE, and ETABLE. It is easy to model an ideal amplifier, and the simulation of devices with special frequency- or time-dependent behaviour is possible and is presented here. Necessary for our solution are amplifiers, limiters, a multiplier, a diff, an active low-pass and a phase-shifting circuit. Figure 13 shows the complete subcircuit with all the mentioned components. Note: $p=j\omega$ or $s=j\omega$ are equivalent.

Starting with a preamplifier and a limiter, a phase-shifting circuit (a combination of low-pass and band-pass, each of 2nd order) follows, dedicated to the application parameters mentioned above. Its transfer function is given by (15) with the parameters $\omega_c=1/\sqrt{L_{ph} \cdot C_{ph}}$ and $Z_0=\sqrt{L_{ph}/C_{ph}}$, $R_{ph}/Z_0=B/f_c=11.5$ with $L_{ph}=40 \mu H$, $C_{ph}=120 pF$ and $R_{ph}=50 \Omega$.

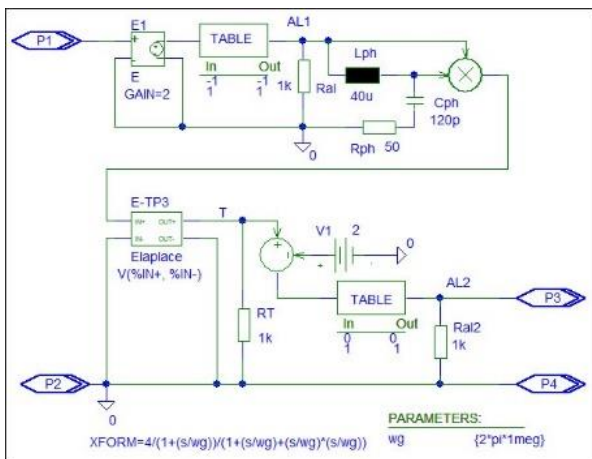


Fig. 13: Modelled subcircuit for an FSK-demodulator

Source: created by the author.

$$F(p) = \frac{1 + \frac{p}{\omega_c} \cdot \frac{R_{ph}}{Z_0}}{\left(\frac{p}{\omega_c}\right)^2 + \frac{p}{\omega_c} \cdot \frac{R_{ph}}{Z_0} + 1}, \quad (15)$$

As an alternative, the values $L_{ph}=4 \mu H$, $C_{ph}=1.2 nF$, and $R_{ph}=5 \Omega$ may be used. The smaller L_{ph} is probably better in view of a practical solution, as shown below in a realized example. The frequency-dependent phase can be derived from (15) and is shown in Figure 14, referring to Fig.11 and eq. 16:

$$\varphi(f) = \arctan\left(\frac{f}{f_c} \cdot \frac{R_{ph}}{Z_0}\right) - \arctan\left(\frac{\frac{f}{f_c} \cdot \frac{R_{ph}}{Z_0}}{1 - (f/f_c)^2}\right) \quad (16)$$

The output voltage is multiplied by the input voltage to separate the two FSK frequencies, see (17). The phase-shifter delivers an angle of $\phi=-40^\circ$ for $f_1=2.2$ MHz and $\phi=-130^\circ$ for $f_2=2.4$ MHz, so that the

multiplication (see (17)) gives a suitable signal.

$$\sin(\omega \cdot t) \cdot \sin(\omega \cdot t + \varphi) = \frac{1}{2}(\cos \varphi - \cos(2\omega \cdot t + \varphi)) \quad (17)$$

The marked difference $\cos(\phi_1)-\cos(\phi_2)=1.41$ (Figure 14) is finally relevant; the term with $2 \cdot \omega$ has to be removed by a suitable low-pass filter.

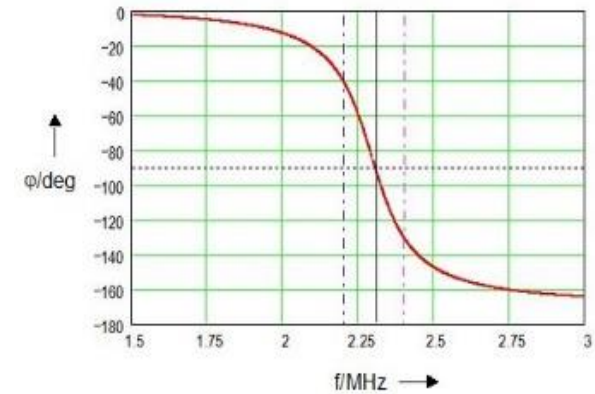


Fig. 14: Frequency-dependent phase

Source: created by the author.

For comparison, a practical result can be seen in Figure 15 with the clock signal and transmitter signal and the appertaining input and output voltages of the phase shifting circuit, which is connected to the complete analog part of the transmission system. The clock signal is not the same as the modulating signal, which is here pseudo-random because of the used signal generator, [11]. Furthermore, the transmission delay has to be considered with a view on the input signal of the phase-shifting circuit. The measured phase difference with relation to the output signal and the signal frequency $f_1=2.2$ MHz is here $\Delta\varphi \approx -55^\circ$. The deviation from the theoretical value should be mentioned. The difference between the signal frequencies f_1 and f_2 cannot be seen here with the naked eye, because of the small frequency deviation. The result for $f_2=2.4$ MHz is similar; we find $\Delta\varphi \approx -110^\circ$.

An active low-pass (LP3, Butterworth type) with the transfer function (18) and with $b_0=4$, cut-off frequency $f_g=1$ MHz, and another limiter are building the output circuit.

$$F(p) = \frac{b_0}{\left(1 + \frac{p}{\omega_g}\right) \cdot \left(\left(\frac{p}{\omega_g}\right)^2 + \frac{p}{\omega_g} + 1\right)}, \quad (18)$$

Finally, the output signal of the entire transmission system can be valued, see Figure 16. It becomes clear that the values of the system parameters have been chosen correctly, and the data-output signal has no distortions.

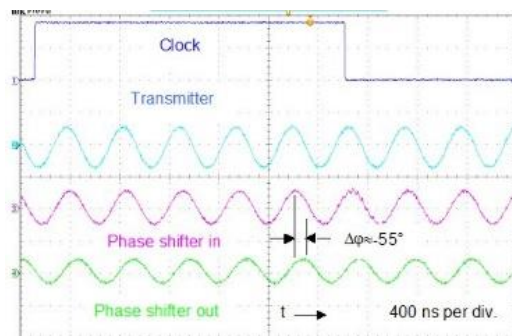


Fig. 15: Selected measurement results (screenshot)
Source: created by the author.

Discussing the application aspects of high-speed traffic, the loop occupation time has to be taken into account; see, [3]. With the loop length of $l=40$ cm and a train speed of $v_0=360$ km/h=100 m/s, we have to deal with $T_0=4$ ms. In Figures. 10 and 13, a value of $T_0=50$ μ s has been chosen to limit the number of simulation data with consideration of the time step of 10 ns concerning the transient analysis.

The main aspect of the simulation lies in the possibility to change and improve all the mentioned parameters for an optimised system design.

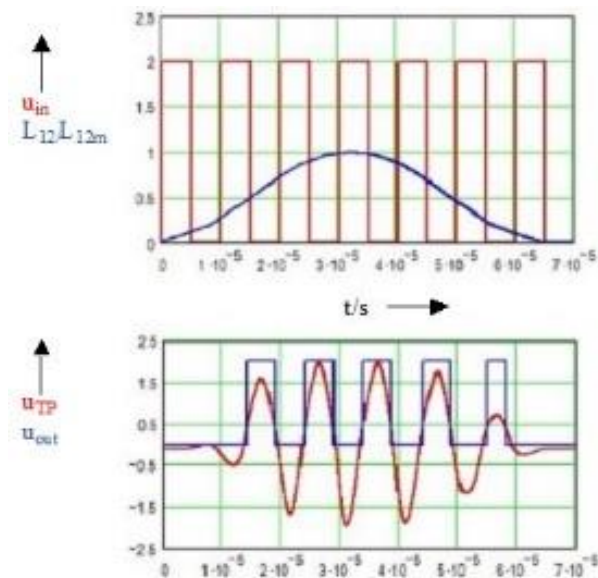


Fig. 16: Final simulation results (data-in, data-out, $L_{12}(t)/L_{12m}$, $u_{TP3}(t)$ (reduced))
Source: created by the author.

7 Conclusion

The modulation method FSK has been in practice for several decades, and there are theoretical basics as well. These should be used for the design of a data transmission system, but accompanied by powerful

software tools. Using the program SPICE for the system simulation, it is necessary to create several

subcircuits, including the modulator, the demodulator with its components, and especially the classical analog part of the transmission path.

Besides the transmitting and receiving band-passes, however, the transient behaviour of the coupled inductive loops has to be determined, where one of

them is in motion, as we find in traffic-related applications. Laboratory tests of high-speed applications are difficult, so the simulation of parts and parameters, such as time-dependent coupling factors, is very helpful. It is shown how all the essential components can be simulated as subsystems separately and in cooperation, and the time-dependent signals are available for the design process of the entire system.

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