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DATA SET FOR LOW-SPEED REMOTE TERMINALS

by

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1. INTRODUCTION

Data transmission over telephone lines involves the acceptance of serialized binary data, modulation of the data to a signal suitable to the bandwidth of the telephone channel, transmission of the signal over the line, reception at the receiving end and demodulation of the signal to reproduce the original binary data. One typical application of data communication is between a low-speed terminal and a computer. In such a case, full-duplex operation is often desirable (i.e., transmission of information on the line in both directions simultaneously). The data set described in this report operates full-duplex and provides for channelization of the telephone line for the simultaneous use of several data sets. The data set can either be coupled to leased telephone lines or acoustically coupled to any telephone.

2. DESIGN CONSIDERATIONS

In determining design specifications, it is illuminating to consider statistically the nature of the binary information. Assuming that the transmitted binary data is ergodic, that the two states have the same probability of occurrence and that the modulated signal to the receiver is noncoherent, the autocorrelation of the binary data can be shown to be as given in Fig. 1a. The power-density spectrum of the autocorrelation is shown in Fig. 1b. In a frequency-shift-keyed (FSK) system, one binary level produces a frequency f_1 and the other level produces a frequency f_2 . This moves the power-density spectrum to f_1 and f_2 when f_1 and f_2 , respectively, are keyed ON (see Fig. 2).

It was desired that each data set use as narrow a bandwidth as possible so the telephone line's available bandwidth might be channelized for simultaneous use by several data sets. For a rate of 200 bits per second (the IBM 1050 system operates at about 150 bits per second), the optimum separation for the two frequencies f_1 and f_2 was found experimentally to be 150 cps. The combined spectrum around f_1 and f_2 can thus be band-limited to 300 cps (Fig. 3) for telephone line transmission.

Continuing the statistical approach to the detection of the signal, we note that the receiver must "decide" whether the signal of f_1 or of f_2 was sent. As has been shown,^[1,2] the decision may be made by taking the cross-correlation of the signals of f_1 and f_2 with the actual received wave form, and then selecting the larger cross-correlation. This may be accomplished by passing the received signal through matched filters and comparing the envelopes of the outputs (see Fig. 4) to see which is larger. This method is ideal for a noncoherent FSK system because the variable phase angle of the received signal is of little importance and the timing can be obtained from the envelope detector output.

FIGURE 1a. AUTOCORRELATION OF BINARY DATA

FIGURE 1b. POWER-DENSITY SPECTRUM OF BINARY DATA WITH τ EQUAL TO BIT LENGTH

FIGURE 2. POWER-DENSITY SPECTRUM FOR FSK

FIGURE 3. POWER-DENSITY SPECTRUM FOR 200 BITS/SECOND

FIGURE 4. MATCHED FILTER DETECTION FOR FSK

3. DATA SET GENERAL DESCRIPTION

In the data set pair shown in Fig. 5, a zero voltage level input produces a frequency f_1 and a negative voltage input produces a frequency f_2 . The data set is coupled to the line in such a way that the signal energy from the transmitter is sent to the line and not to the receiver (i.e., the energy from the transmitter of data set A does not enter the receiver of the set, but is sent through the line to the receiver of data set B).

In the receiver, the amplifier-limiter produces filtering in addition to that obtained by the line coupling filter and yields a constant amplitude output. The discriminator delivers a positive level for the signal of f_1 and a negative level for f_2 , and its output is squared to recover the original binary data.

3.1 Transmitter Description

The modified Class C Hartley oscillator for this data set requires fewer components than the free-running multivibrator conventional in FSK transmission. The circuit schematic of the transmitter is given in the Department of Computer Science Circuits Book as symbol F_{sx} and Circuit 415-175. To produce the desired frequency shift, part of the tank circuit inductance is switched into or out of the circuit. When the input voltage is zero, the input transistor is OFF and the oscillator operates with the total inductance in the tank circuit. When the input is -4 to -12 volts the input transistor saturates, switching out part of the inductance and raising the frequency. Resistance is added to keep the load on the tank circuit constant, regardless of whether part of the inductance is switched out or not. A current transient is not produced when switching from one frequency to the other. The amplitude of the output voltage of the transmitter is constant for both frequencies and the transition is virtually free from distortion.

The oscillator is tuned by a tuning slug in the pot core. Since the frequency of oscillation is inversely proportional to the number of turns of the coil, the desired frequency shift of 150 cps can be fixed in the design of the coil. Both frequencies are set to the proper value by tuning to either one. Data set A transmits frequencies $f_{1A} = 1275$ cps and $f_{2A} = 1425$ cps; data set B transmits frequencies $f_{1B} = 2175$ cps and $f_{2B} = 2325$ cps. The frequencies

FIGURE 5. OPERATION OF DATA SET PAIR

selected are odd harmonics of 75 cps to avoid 60-cps harmonics of the primary power.

The oscillator is connected to the line coupling through a transistor. This transistor, connected in common collector configuration, puts a very small load on the oscillator and isolates it from line disturbances. The transmission level can be varied from -15 to 0 db by changing a potentiometer.

3.2 Line Coupling Description

The problem of line coupling (terminating both transmitter and receiver to a single wire pair) is an old one to the telephone industry. The standard termination is a form of a Wheatstone bridge known as a "hybrid," which generally consists of two transformers. The receiver accepts energy from the line, but energy from the transmitter is cancelled by the two opposing identical windings.

However, in this data set, a composite hybrid was designed which incorporated band-pass filters of 300 cps on the transmitter output and the receiver input. This not only filters energy from the line, but doubles the normal hybrid rejection of the data set's own transmitter energy as compared to the spectrum to which the receiver is tuned.

When several data sets are used simultaneously on one line pair in a channelized fashion, the line couplings are connected in series and the balancing resistor is appropriately reduced.

3.3 Receiver

In the receiver circuit (see card descriptions 415-175 and 415-176, symbols F_{sx} and F_{SR}), the three-transistor amplifier-limiter uses interstage tuned coupling, with standard transistor biasing. The line coupling and the amplifier form, in effect, a band-pass filter with a bandwidth of 300 cps and cutoff attenuation of 36 db per octave. The amplifier-limiter delivers constant amplitude for a 50-db range of received signal amplitude fluctuation. A transistor is used in common collector configuration for impedance matching the amplifier-limiter to the discriminator.

The discriminator has two tuned filters--one for the signal of f_1 and one for the signal of f_2 . These filters have approximately the same shape as that of the signal spectrum, but are intentionally mismatched to allow for frequency shifts caused by the carrier systems and variations due to temperature changes and the aging of components. The signal of each tuned filter is taken by a secondary winding to a bridge rectifier. The two rectifiers produce output voltages (or signal envelopes) of opposite polarity. These outputs are summed and passed through a low-pass filter with cutoff at 300 cps and 18 db per octave attenuation, which is also designed to match the output stage to the discriminator. The positive and negative voltage pulses from the low-pass filter, which represent the original binary data, are squared by a Schmitt trigger and confined between zero and any voltage between -4 and -12 volts as determined by the clamping diode for a load of 1000 ohms. Since the timing is derived solely on a bit-to-bit basis from the envelope of the low-pass filter output, a slight amount of distortion or "jitter" occurs. The symmetry between the two states of the binary data can be adjusted by a potentiometer.

4. PERFORMANCE

4.1 Distortion

The distortion of binary data may be defined as the distance the leading or lagging edge of a bit is shifted, divided by the original bit length, or

$$\text{Distortion (per cent)} = \frac{d}{T_0} \times 100$$

where d = shift and T_0 = original bit length. Besides whatever distortion the telephone line might introduce, the data set itself introduces two types of distortion caused by the noncoherent nature of the system and by changes in operating temperature and supply voltage.

The bit rate is not synchronized with the transmitting frequencies f_1 and f_2 , nor are these frequencies exact multiples of the bit rate. The discriminator disregards the phase angle, and the timing is derived from the envelope detector. For a transmission rate of 200 bits per second and a carrier frequency of 1275 cps, this noncoherent system theoretically causes, at worst, two per cent distortion.^[3] The distortion measured for this data set was at worst three per cent.

When the data set was tested in an operating range from 25° to 55°C, with supply voltage fluctuations of ten per cent, two types of distortion were observed: transmitter frequency drift, which was less than 0.5 per cent and caused maximum distortion of three per cent, and discriminator frequency drift (drift of the center frequency of the tuned filters), causing a maximum distortion of three per cent.

Assuming the unlikely condition that all sources of distortion are at their worst, the maximum possible distortion would be nine per cent. Under normal operating conditions the distortion is less than three per cent.

4.2 Selectivity

In the frequency response curve of the discriminator of data set B (Fig. 6), the output voltage from the low-pass filter is plotted versus the

FIGURE 6. DISCRIMINATOR OUTPUT VOLTAGE VS. FREQUENCY OF APPLIED LINE SIGNAL (DATA SET B. LINE SIGNAL = -6 dbm; LINE ATTENUATION = 10 db)

frequency of the applied line signal. The arrows point to the frequencies received from the transmitter of data set A ($f_{1A} = 1275$ cps and $f_{2A} = 1425$ cps).

Figure 7 is a composite plot of the discriminator output voltage of data sets A and B for each of the four frequencies with harmonic noise superimposed on the line. In the test, each discrete frequency of the harmonic noise was the same magnitude as the signal. The discriminator output voltage shifts from the level established by the transmitted signal only where the harmonic noise is of the other frequency to which the discriminator is sensitive. No error is caused until the voltage passes a "decision" level established by the Schmitt trigger. The composite plot shows the part of the total telephone channel of 300-3000 cps which is used for a data set pair. As can be seen, the bandwidth requirement for each frequency pair is only 300 cps. Thus the frequency band could be channeled easily for more than one data set pair per telephone line.

DATA SET A RECEIVES f_{1B} (2175 cps) AND f_{2B} (2325 cps) ---
 DATA SET B RECEIVES f_{1A} (1275 cps) --- AND f_{2A} (1425 cps) —

FIGURE 7. DISCRIMINATOR OUTPUT VOLTAGE VS. HARMONIC NOISE SUPERIMPOSED ON TRANSMITTED SIGNAL (COMPOSITE PLOT FOR DATA SETS A AND B. HARMONIC NOISE = -6 dbm; LINE ATTENUATION = 10 db; TRANSMITTED SIGNAL = -6 dbm)

5. ACOUSTICAL COUPLING

Outputs are available from the 415-175 cards so the data set may either be connected directly to the telephone line or to acoustical couplers for dial-up transmission. For the acoustical coupling, a jumper between pin 2 of TX2 and pin 2 of TX3 is removed as well as the balancing resistor between pins 4 and 1 of TX2 and TX3, respectively. A 2-1/4 inch P.M. speaker of 100 ohms is connected between logic points C and M, and a 1-inch magnetic microphone of 1000 ohms with a frequency response of 400 to 4000 cps is connected between logic points B and F. These are placed in contact with the speaker and earpiece of the telephone after line connection has been established by dialing.

6. POWER SUPPLIES

The data set is designed to use two voltage supplies, +12 volts and -12 volts. These are the supplies which are normally available with IBM remote terminals. For a teletype remote terminal and for the central, the available supplies have greater magnitudes so the voltages of +12 volts are derived with zener diode clamps on the 415-175 and 415-176 cards.

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APPENDIX. DESIGN EQUATIONS FOR DATA SET

A.1 Input Stage

The input stage was designed in the same manner as a standard inverter except that the added collector current from the tank circuit was taken into consideration. The design of the data set employed the Taylor worst-case design criteria.

A.2 Frequency Shift Oscillator

The Class C operation was used for the oscillator so that the transistors could be interchanged, and leakage current (I_{co}) had little significance since it merely provides additional bias margin at high temperatures. The ac equivalent circuit is shown in Fig. 8.

The oscillation frequency may be obtained by considering the active and passive networks connected in series.^[4] The circuit equations, given in "h" parameters, are

active:

$$v_1 = h_{11_a} i_1 + h_{12_a} v_2$$

$$i_2 = h_{21_a} i_1 + h_{22_a} v_2$$

passive:

$$-v_1 = h_{11_p} i_1 + h_{12_p} (-v_2)$$

$$i_2 = h_{21_p} i_1 + h_{22_p} (-v_2)$$

Combining the equations to eliminate v_1 and i_2 yields

$$0 = (h_{11_a} + h_{11_p}) i_1 + (h_{12_a} - h_{12_p}) v_2$$

$$0 = (h_{21_a} - h_{21_p}) i_1 + (h_{22_a} + h_{22_p}) v_2$$

If the determinate of the equations is set equal to zero, the imaginary part may be solved for the oscillation frequency.

The frequency of oscillation for the low frequency is given by

FIGURE 8. OSCILLATOR ac EQUIVALENT CIRCUIT

$$\omega^2 = \frac{A + \frac{BD - EF' - E'F}{C}}{G}$$

where

$$A = \frac{R_{23} + R_{12} + R_2}{R_3} + 1 + h_{22a}(R_{23} + R_{12} + R_2)$$

$$B = Cl(R_{23} + R_{12} + R_2) + \frac{L_{13}}{R_3} + h_{22a}L_{13}$$

$$C = L_{23}(R_{12} + R_2) + L_{12}R_{23} + L_{13}R_1 + h_{11a}L_{13}$$

$$D = R_{23}(R_{12} + R_2) + R_{23}R_3 + (R_{12} + R_2)R_3$$

$$E = h_{12a}L_{13} - L_{23} - M_{23-12}$$

$$E' = h_{21a}L_{13} + L_{23} + M_{23-12}$$

$$F = h_{12a}(R_{23} + R_{12} + R_2) - R_{23}$$

$$F' = h_{21a}(R_{23} + R_{12} + R_2) + R_{23}$$

and

$$G = L_{13}Cl$$

and where

L_{ij} = inductance of winding between points i and j

M_{ij-kl} = mutual inductance between L_{ij} and L_{kl}

and

R_{ij} = resistance of winding between i and j

All other components are shown in the original circuit schematic.

For the high-frequency case, the following substitutions are made:

replace:	with:
L_{13}	L_{34}
L_{12}	L_{24}
M_{23-12}	M_{23-24}
$R_{12} + R_2$	R_{24}

To keep the amplitude of the oscillator the same for both frequencies, the loading of the tank circuit must be constant. The ac equivalent circuits for the tank circuit are shown in Fig. 9. The loading produced by the tank itself is $\omega_1 L_{12} Q$ for the low-frequency case and $\omega_2 L_{24} Q'$ for the high-frequency case, where

$$Q = \frac{\omega_1 L_{13}}{R_{13} + R_2}$$

and

$$Q' = \frac{\omega_2 L_{34}}{R_{34}}$$

Then equating the total loading of the two cases,

$$\omega_1 L_{12} Q // \text{effect of } R_3 = \omega_2 L_{24} Q' // n^2 R_2 // \text{effect of } R_3$$

or

TRANSISTOR T1 OFF (LOW-FREQUENCY CASE)

TRANSISTOR T1 ON (HIGH-FREQUENCY CASE)

FIGURE 9. EQUIVALENT CIRCUITS FOR TANK CIRCUIT

$$\frac{1}{\omega_1 L_{12} Q} + \frac{1}{\text{effect of } R3} = \frac{1}{\omega_1 L_{24} Q} + \frac{1}{n^2 R2} + \frac{1}{\text{effect of } R4}$$

where

$$n^2 = \frac{L_{34}}{L_{14}} = \left(\frac{\text{turns from 3 to 4}}{\text{turns from 1 to 4}} \right)^2$$

Considering the effect of R3 to be nearly the same for both cases, the result is

$$\frac{R_{13} + R2}{\omega_1^2 L_{12} L_{13}} = \frac{R_{34}}{\omega_2^2 L_{24} L_{34}} + \frac{1}{n^2 R2}$$

For a proper conduction angle of the transistor, the peak base-to-emitter voltage should be approximately nine-tenths of the supply voltage, which gives an additional voltage loop condition (refer to Fig. 8).

Point 2 of TX 1	Across R5	V_{BE} Across T2	Point 3 of TX 1
--------------------	-----------	-----------------------	--------------------

$$E + \frac{(E)(R1)}{R1 + R_{12} + \omega_1 L_{12} + R2} + 0.9E = aE$$

where

$$a = \frac{\text{turns from 1 to 3}}{\text{turns from 1 to 2}}$$

and

$$E = \text{supply voltage}$$

Therefore

$$\frac{R1}{R1 + R2 + R_{12} + \omega_1 L_{12}} = a - 1.9$$

The coupling capacitor should be chosen to give the time constant ($R3C_c$) a value of 10 to 50 oscillation periods.

The design of the frequency shift oscillator must meet the conditions of frequency of oscillation, tank circuit load and conduction angle of the transistor.

A.3 Tuned Circuit Synthesis

The hybrid, the amplifier-limiter interstage and the discriminator filter circuits were all designed using the same theory. The basic circuit and its equivalent circuit are shown in Fig. 10, and the latter can be analyzed^[5] to yield the following design equations:

$$B' = \frac{B}{f_0}$$

$$Q = \frac{1}{(1 - \sqrt{\eta_{\max}}) B'}$$

$$L = \frac{B' Q - 1}{4\pi f_0 Q g_1}$$

$$a^2 = \frac{g_2}{g_1}$$

$$g'_2 = \frac{g_2}{a^2}$$

$$L_2 = \frac{L}{a}$$

$$C = \frac{1}{4\pi^2 f_0^2 L_1}$$

where

FIGURE 10. TUNED CIRCUIT AND EQUIVALENT CIRCUIT

B = bandwidth

f_0 = center frequency

B' = normalized bandwidth

Q = Q of the coil

$\eta_{\max}$ = power-transfer efficiency when $g'_2 = g_1$

L_1 = inductance of the primary

L_2 = inductance of the secondary

g_1 = conductance of input

g_2 = conductance of output

a = turns ratio

C = capacitance

g_p = equivalent parallel conductance of the tank circuit

A.4 Amplifier-Limiter

The amplifier-limiter is designed in the conventional manner, using Taylor worst-case design criteria. The interstage is designed according to the design equations given in the preceding section.

A.5 Low-Pass Filter

The low-pass filter was designed to be a third-order Chebyshev filter with pole locations as shown in Fig. 11. The pole locations for this filter are given in the polynomial

$$s^3 + 2\sigma s^2 + (1.25\sigma^2 + 0.75)s + \sigma(0.25\sigma^2 + 0.75)$$

The circuit configuration (Fig. 12) has pole locations expressed by a polynomial of the same form:

$$s^3 + \left(\frac{1}{R_2 C_2} + \frac{1}{R_1 C_1} + \frac{R_L}{L} \right) s^2 + \left(\frac{1}{LC_1} + \frac{1}{LC_2} + \frac{1}{R_1 R_2 C_1 C_2} + \frac{R_L}{R_2 LC_2} + \frac{R_L}{R_1 LC_1} \right) s + \frac{R_1 + R_2 + R_L}{R_1 R_2 LC_1 C_2}$$

FIGURE 11. POLE LOCATIONS FOR THIRD-ORDER CHEBYSHEV FILTER

FIGURE 12. CIRCUIT FOR CHEBYSHEV FILTER WITH POLE LOCATION OF FIG. 11

The resistances are determined by the discriminator output impedance and the input impedance of the Schmitt trigger. The inductance and capacitances are then selected to satisfy the above equation.

A.6 Schmitt Trigger

The Schmitt trigger used for the output stage was designed by writing equations for the two stable conditions and applying the Taylor worst-case design criteria.