LOW-SPEED MODEM SYSTEM DESIGN USING THE MC6860

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This application note describes modem systems for full duplex originate only, automatic answer-answer only, and answer/originate with automatic answer. Described are the peripheral circuits such as limiters and bandpass filters that surround the MC6860 to make it a 100 Series compatible modem system.

MOTOROLA *Semiconductor Products Inc.*

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GENERAL

Low-speed modem designers will find that the MC6860 MOS LSI Modem with its built-in modulator, demodulator, and supervisory control will allow the design of a high performance, low cost 100 Series type modem. The designer, by selecting from different filter configurations and some surrounding support circuitry, may design either an originate only, answer only, or automatic answer/originate modem system.

It is the purpose of this note to cover in some detail these surrounding building blocks that comprise the total system. To familiarize the reader with the MC6860 chip operation, a general overview will be included with a more detailed description to be obtained from the MC6860 data sheet.

BASIC MC6860 CIRCUIT OPERATION

As illustrated in Figure 1, the MC6860 Modem contains a digital modulator, demodulator, and a supervisory control section to handle line disciplines for full duplex originate, auto-answer, and auto-disconnect operations.

Modulator

The modulator section converts serial digital data into analog frequencies for output to the telephone network. The analog output from the modem is a digital synthesized sinewave having one of four possible frequencies as listed in Figure 2. The modulation scheme used is frequency shift keying (FSK), where a logic "0" (space) is the lower frequency and a logic "1" (mark) is the upper or higher
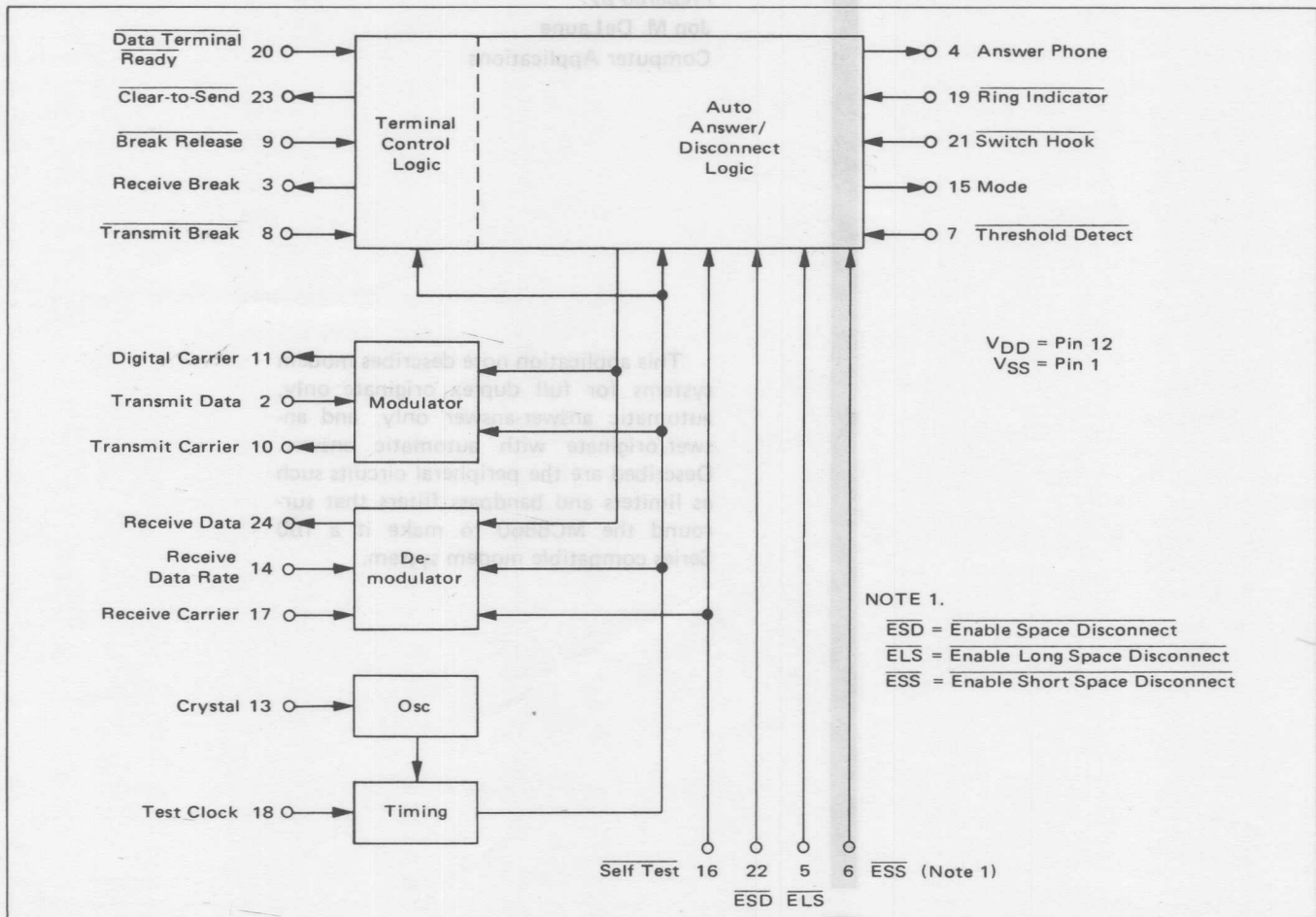

FIGURE 1 — MC6860 Modem

Output	Originate	Answer
Mark	1270 Hz	2225 Hz
Space	1070 Hz	2025 Hz

FIGURE 2 — Output Frequency Shift Keying Pairs

frequency of either the originate or answer frequency pairs. The analog signal output level from the modulator is typically 350 millivolts (rms) into a load of 100 k ohms; therefore, for the MC6860 to interface into a 600 ohm line system such as the telephone network with the necessary signal magnitude, an external transmit buffer will be required.

Demodulator

The demodulator section receives either the lower or upper (answer or originate modem) frequency tone pairs, and by a technique of digital half-cycle detection determines the presence of a mark or a space frequency and will output at the Receive Data pin either a digital logic "1" or "0" to the terminal or computer equipment. The incoming analog signal from the line should be bandlimited (filtered) and limited (amplified/clipped) prior to the demodulator carrier input to remove interfering signals and system noise. The limited input signal presented to the demodulator input should be at 50% duty cycle (±4%) over the full input signal dynamic range and be at a TTL compatible input level in order to maintain low bit-error-rate performance.

Supervisory Control

The supervisory control section of the MC6860 contains the necessary logic to provide initial inter-modem handshaking as well as operational protocol, such as automatic answer, originate only, initiate disconnect, and automatic disconnect. A graphical illustration of these control operations provided by the MC6860 is shown in Figures 3, 4, 5, and 6. Signals provided by the MC6860 for interfacing between a data terminal and either a CBS or a CBT telephone network data coupler are shown at the top right of Figure 1. Switch Hook (SH), Ring Indicator (RI), and Answer Phone (An Ph) signals will interface directly with a CBT data coupler, or with a CBS data coupler when RS-232 interface circuits are used. Both of these data coupler interface methods will be illustrated in later system implementation examples.

Additional control signals that are provided for data terminal control are: Data Terminal Ready (DTR), Clear-to-Send (CTS), Receive Break (Rx Brk), Transmit Break (Tx Brk), and Break Release (Brk R). The Mode output is a control function that is system oriented for the surrounding filter block. This output can be used to control switchable filters to provide a full automatic answer/originate modem system. A logic low level at the Mode output pin indicates the demodulator is in the answer mode of operation and will demodulate 1070 Hz and 1270 Hz incoming signals. When the Mode output is in a high state, the frequencies demodulated will be 2025 Hz and 2225 Hz. A design example using switchable filters will be illustrated in a later section.
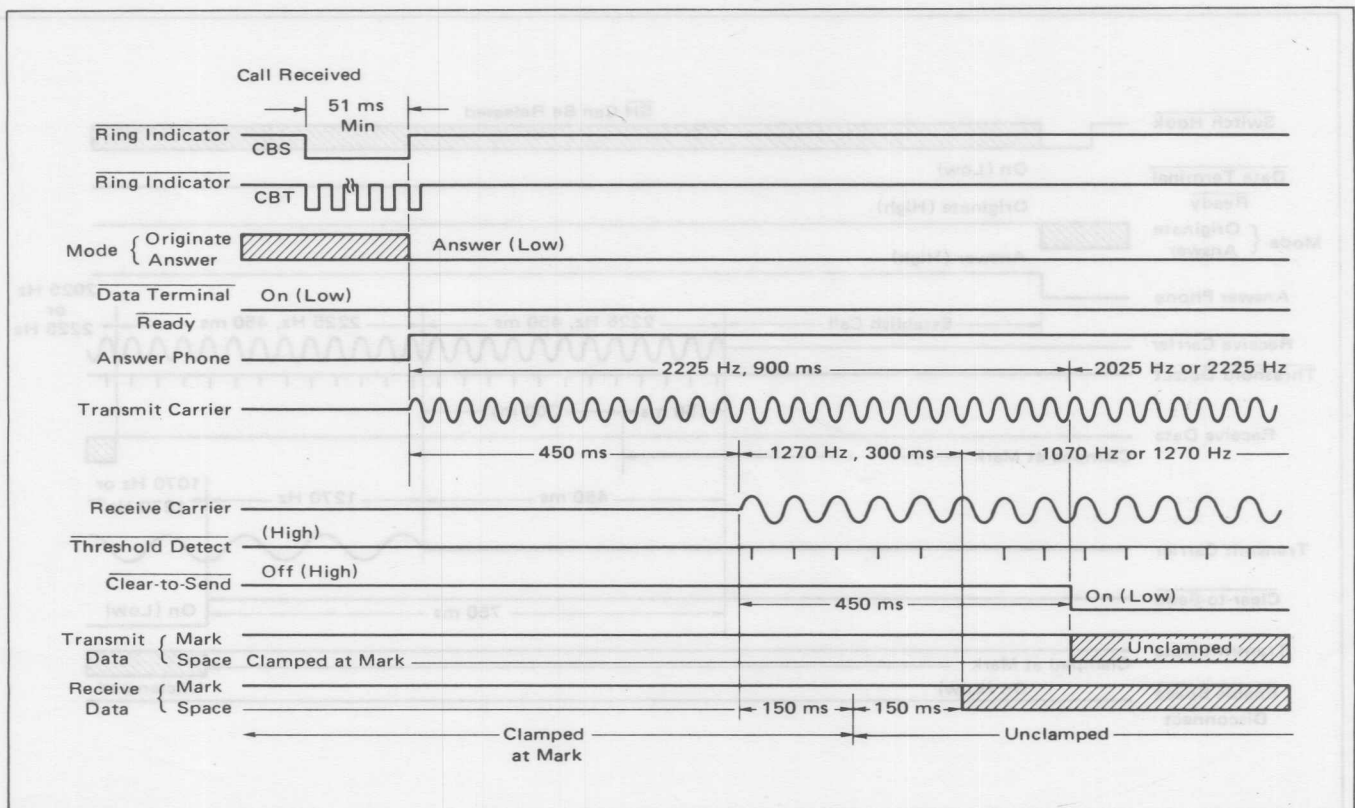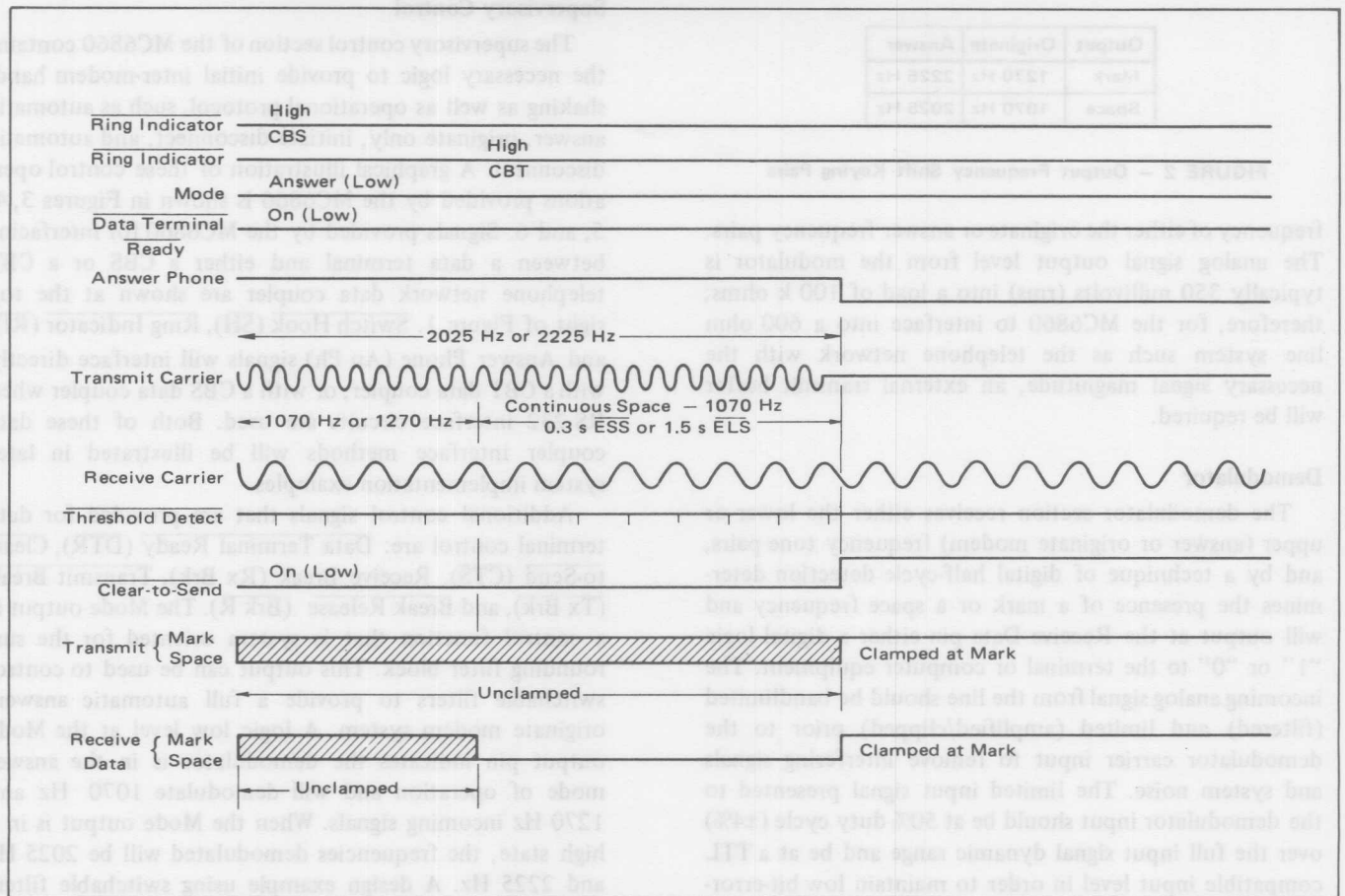

FIGURE 3 — Automatic Answer
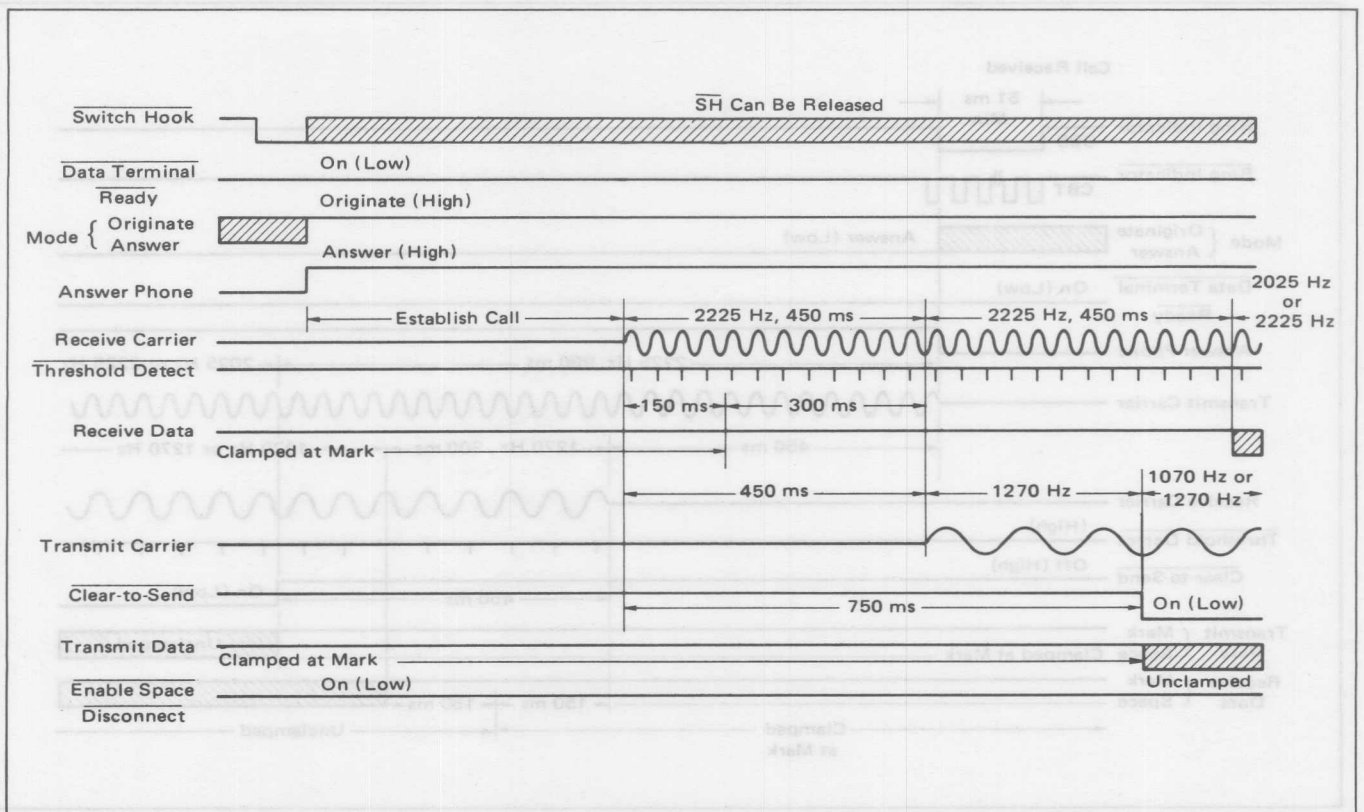
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FIGURE 4 — Automatic Disconnect — Long or Short Space

Signal labels (top to bottom):
- Ring Indicator CBS — High
- Ring Indicator CBT — High
- Mode — Answer (Low)
- Data Terminal Ready — On (Low)
- Answer Phone
- Transmit Carrier — 2025 Hz or 2225 Hz
- Receive Carrier — 1070 Hz or 1270 Hz / Continuous Space — 1070 Hz / 0.3 s ESS or 1.5 s ELS
- Threshold Detect
- Clear-to-Send — On (Low)
- Transmit { Mark / Space } — Clamped at Mark / Unclamped
- Receive Data { Mark / Space } — Clamped at Mark / Unclamped

FIGURE 5 — Originate Only

Signal labels (top to bottom):
- Switch Hook — SH Can Be Released
- Data Terminal Ready — On (Low)
- Mode { Originate / Answer } — Originate (High) / Answer (High)
- Answer Phone
- Receive Carrier — Establish Call — 2225 Hz, 450 ms — 2225 Hz, 450 ms — 2025 Hz or 2225 Hz
- Threshold Detect
- Receive Data — Clamped at Mark — 150 ms — 300 ms
- Transmit Carrier — 450 ms — 1270 Hz — 1070 Hz or 1270 Hz
- Clear-to-Send — 750 ms — On (Low)
- Transmit Data — Clamped at Mark
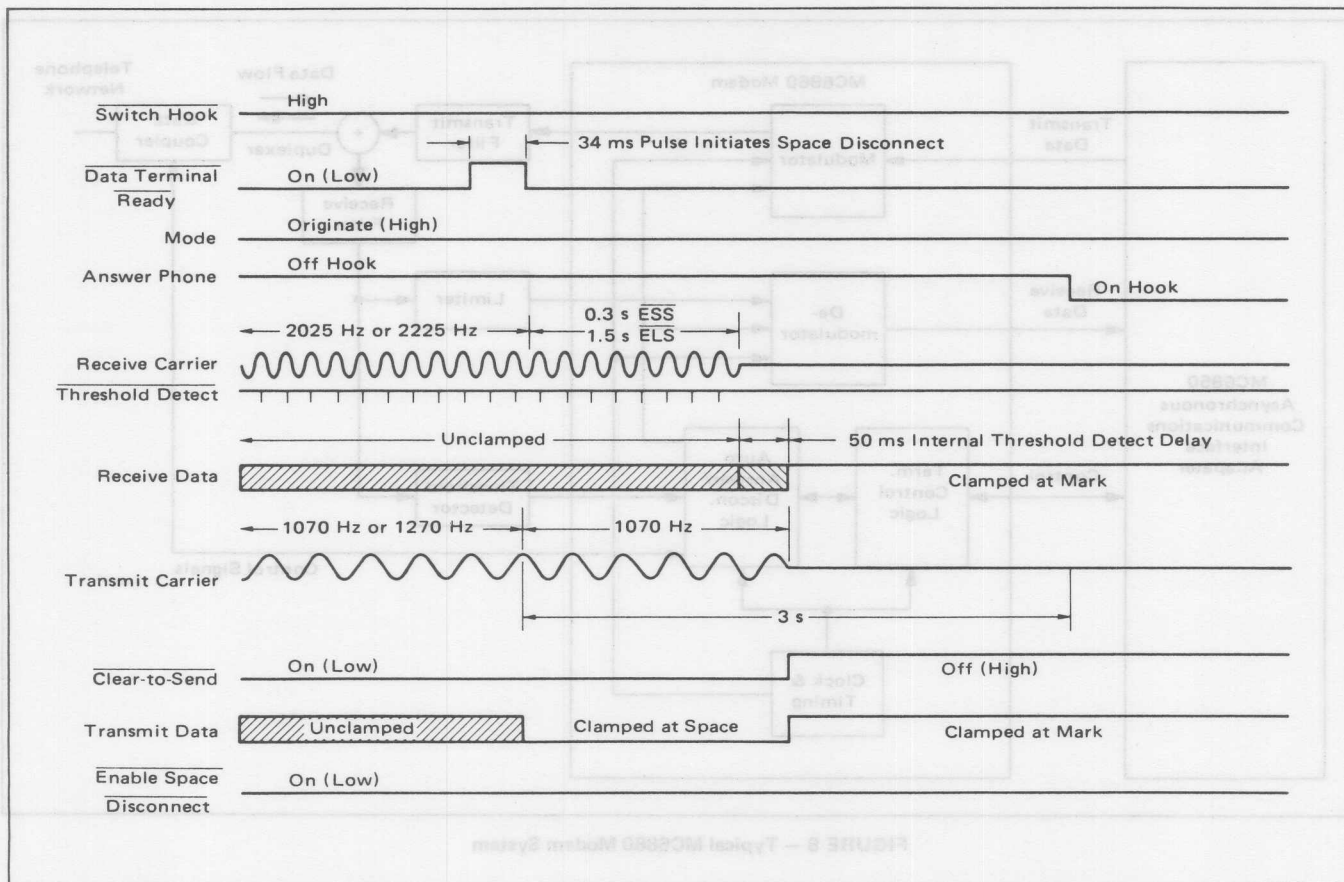- Enable Space Disconnect — On (Low) — Unclamped

4

FIGURE 6 — Initiate Disconnect

A self test feature is included in the MC6860 for testing the modulator/demodulator sections. When a low logic level is applied to the Self Test (ST) input pin, the demodulator is switched to detect the modulator transmitted frequency pair. Channel establishment obtained during initial handshaking is not lost, with only the Mode output changing state during initiation of self test as shown in Figure 7. This test feature allows the modulator, demodulator, and interval timer circuitry to be checked for proper operation during diagnostic system test.

$\overline{ST}$	$\overline{SH}$	$\overline{RI}$	Mode
H	L	H	H
H	H	L	L
L	L	H	L
L	H	L	H

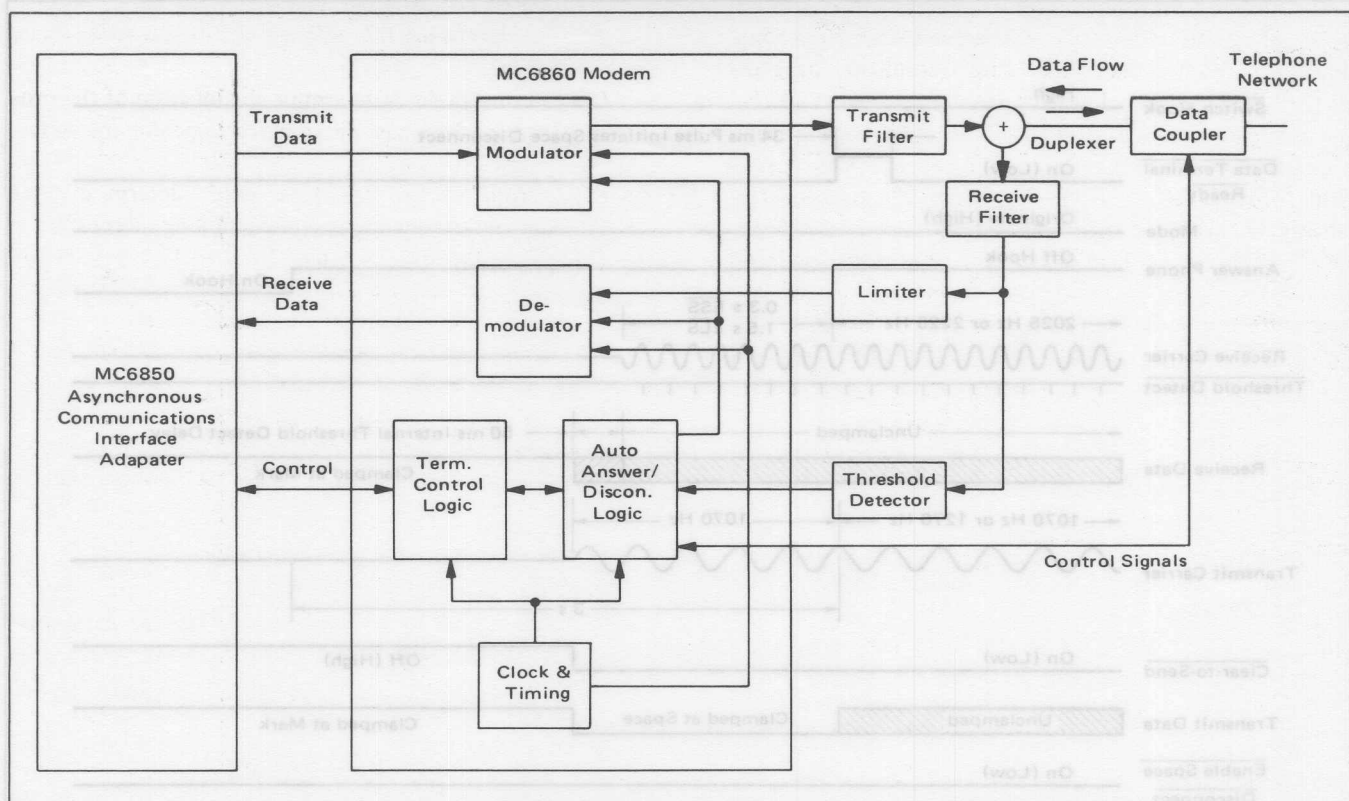**FIGURE 7 — Mode Control Truth Table**

MODEM FILTER DESIGN

Filter networks are among the most important surrounding element blocks in a modem system. As shown in Figure 8, a filter block is used in the receive carrier signal path and another filter block is used in the transmit carrier signal path. The transmit carrier filter may not be required in answer only modem designs but is required for originate mode operation.

The receive filter must provide sufficient adjacent channel rejection to provide good bit-error performance. During answer only operation, the filter must pass the receive frequencies of 1070 and 1270 Hz, but reject the adjacent channel local transmit frequencies of 2025 and 2225 Hz.

Typically, the receive carrier bandpass filter should provide greater than 35 dB attenuation to the adjacent channel. During full duplex originate operation, the local transmit signal produces second harmonic energy within the receive filter bandpass (2 x 1070 Hz = 2140 Hz). To reduce this frequency component in the receive filter passband, a transmit carrier filter must be included. This transmit filter may be either a low pass, a high pass, or a bandpass filter dependent upon the designed mode of operation of the modem: originate only, answer only, or auto answer/originate.

The filter design example presented is a bandpass configuration which could be used in either the transmit or receive signal paths with only component value changes. The transmit filter must have a pass frequency of 2025-2225 Hz when the modem is used as an answer only modem (receiving frequencies of 1070-1270 Hz). The opposite configuration is true when the modem is in the originate only mode of operation (transmit frequencies of 1070-1270 Hz and receive frequencies of 2025-2225 Hz).

A design example is presented, with design tables and equations to solve for the modem system bandpass filter

5

FIGURE 8 — Typical MC6860 Modem System

component values. A 6-pole answer filter is developed in detail in this application note, whereas a 6-pole originate filter has values tabulated only. Also tabulated are component values for 8-pole, 50-dB receive filters and 4-pole, 25-dB transmit filters.

A filter design may take one of many forms. The included design examples use a 0.5 dB ripple Chebyshev approximation. The filter element configuration used is a multiple feedback bandpass as shown in Figure 9. As indicated in Figure 10, the Chebyshev filter will provide a high degree of attenuation in the stop band, but with less phase linearity than a Butterworth or Bessel filter. Linear phase or group delay in the passband is an important design consideration for modem filter design. Error performance and demodulator phase/bias distortion of the modem system is affected by unequal delay of data frequencies within the filter passband. Therefore, it is important to provide filters that not only provide sharp stopband attenuation, but also provide some degree of phase linearity in
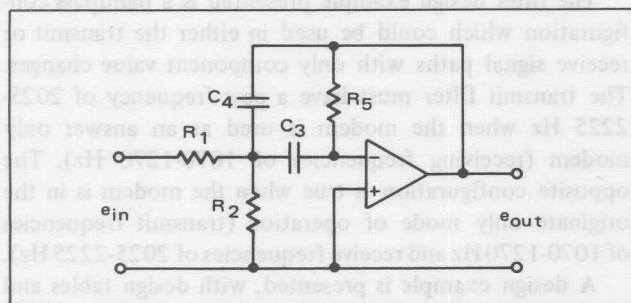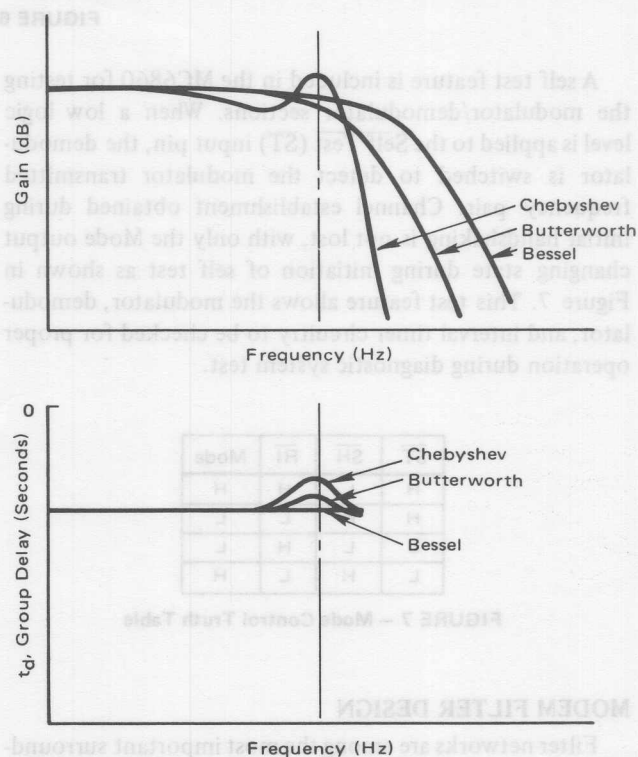

FIGURE 10 — Filter Approximation Characteristics

the passband. By designing the Chebyshev filter to have a wider bandwidth than required for FSK (frequency shift keyed) data recovery, the designer can maximize phase linearity within the required passband.

FIGURE 9 — Multiple Feedback Bandpass Filter Element

Determining the minimum filter bandwidth comes by investigating the received signal characteristics. Data communication theory states that data transmitted by FSK can be recovered by detecting the data carrier and the first sidebands. At a data rate of 300 bits per second and a data format of alternate mark and space, the first sidebands occur ±150 Hz from the carrier which is located halfway between the mark and space frequencies. Therefore, the minimum bandwidth for the receive bandpass filter is 300 Hz. Typically, frequencies within this 300 Hz bandwidth should undergo no greater than 0.8 millisecond change in group delay. Group delay is defined by:

$$t_d = \frac{\triangle\phi}{\triangle F} \frac{1}{360^\circ/\text{cycle}}$$

where $\triangle\phi$ = change in phase in degrees
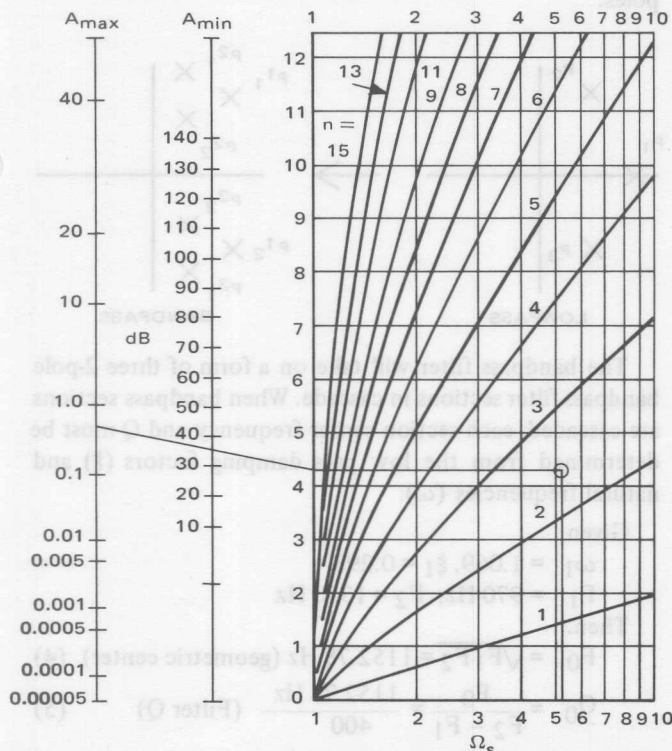 $\triangle F$ = change in frequency in Hz

To maintain less than 0.8 millisecond group delay at a data rate of 300 bits per second requires an overall filter bandpass of 400 Hz. This results in the low frequency pair (answer) filter passband being between 970 Hz and 1370 Hz (6-pole, 0.5 dB ripple Chebyshev).

Filter Design Steps

The modem bandpass filter examples will be designed using the following procedural steps:

(1)   Determine the required prototype low pass filter shape factor from the passband width and stopband attenuation.

(2)   Enter Table 1 with the shape factor, passband

ripple (A_{max}), and stopband attenuation (A_{min}), to determine the order of the prototype lowpass filter.

(3)   From Table 2, determine the location of the prototype low pass filter poles opposite the determined filter order.

(4)   From the low pass filter poles, determine their natural frequency (ω) and damping factor (ξ).

(5)   Transform the low pass filter section parameters to cascaded second order bandpass filter design section Q and center frequency values.

(6)   Determine the active element operational amplifier gain by solving for center frequency loss and system filter passband gain (A_{VO}).

(7)   Use each section Q, frequency, and gain to solve for the bandpass filter passive component values.

Step (1) — Filter Shape Factor

Figure 11 shows a design example for a typical 6-pole answer modem receive filter design. From this data, it is possible to calculate the filter shape factor (Ω_s) for the prototype filter.

$$\Omega_s = \frac{F_4 - F_3}{F_2 - F_1} = \frac{2225 - 115}{1370 - 970} \tag{1}$$

$$\Omega_s = \frac{2110}{400} = 5.28$$

TABLE 2 — Pole Locations and Quadratic Factors
$(s^2 + a_1 s + a_0)$ **for Chebyshev 0.5 dB Ripple Filter**

	0.5 dB Ripple		
Order	Poles	a_0	a_1
2	−0.71281 ± j 1.00404	1.51620	1.42562
3	−0.31323 ± j 1.02193	1.14245	0.62646
	−0.62646		
4	−0.17535 ± j 1.01625	1.06352	0.35071
	−0.42334 ± j 0.42095	0.35641	0.84668
5	−0.11196 ± j 1.01156	1.03578	0.22393
	−0.29312 ± j 0.62518	0.47677	0.58625
	−0.36232		
6	−0.07765 ± j 1.00846	1.02302	0.15530
	−0.21214 ± j 0.73824	0.59001	0.42429
	−0.28979 ± j 0.27022	0.15700	0.57959
7	−0.05700 ± j 1.00641	1.01611	0.11401
	−0.15972 ± j 0.80708	0.67688	0.31944
	−0.23080 ± j 0.44789	0.25388	0.46160
	−0.25617		
8	−0.04362 ± j 1.00500	1.01193	0.08724
	−0.12422 ± j 0.85200	0.74133	0.24844
	−0.18591 ± j 0.56929	0.35865	0.37182
	−0.21929 ± j 0.19991	0.08805	0.43859
9	−0.03445 ± j 1.00400	1.00921	0.06891
	−0.09920 ± j 0.88291	0.78936	0.19841
	−0.15199 ± j 0.65532	0.45254	0.30397
	−0.18644 ± j 0.34869	0.15634	0.37288
	−0.19841		
10	−0.02790 ± j 1.00327	1.00734	0.05580
	−0.08097 ± j 0.90507	0.82570	0.16193
	−0.12611 ± j 0.71826	0.53181	0.25222
	−0.15891 ± j 0.46115	0.23791	0.31781
	−0.17615 ± j 0.15890	0.05628	0.35230

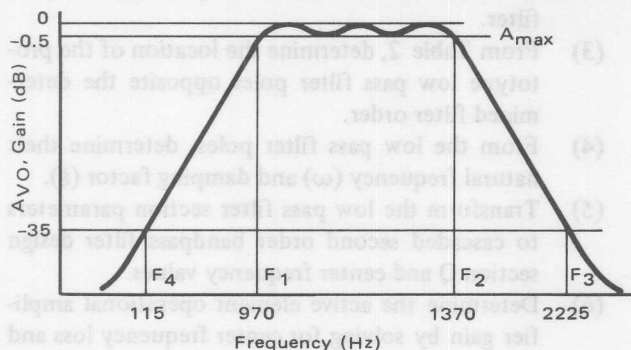**TABLE 1 — Complexity Nomograph for Chebyshev Filters (Zverev)**

FIGURE 11 — Answer Filter Design Goals

where:

F_1 = lower passband frequency in Hz
F_2 = upper passband frequency in Hz
F_3 = lower stopband frequency in Hz
F_4 = upper stopband frequency in Hz

NOTE:

F_1 and F_2 are ripple bandwidth frequencies, i.e., gain down 0.5 dB.

Steps (2) and (3) — Filter Order and Pole Location

The second step of the filter design process was to determine the complexity of the filter. To determine this complexity, the following information is required:

1. The passband ripple, A_{max}.
2. The minimum stopband attenuation, A_{min}.
3. The ratio of the ripple bandwidth and the first frequency of minimum attenuation, shape factor Ω_s.

With A_{max} = 0.5 dB, A_{min} = –35 dB, and Ω_s = 5.28 enter the nomograph in Table 1 to determine the filter complexity or order.

The nomograph is used by locating the passband ripple A_{max} and the minimum stopband attenuation A_{min} and drawing a line from A_{max} through A_{min} to the left-hand side of the graph. From this point, a horizontal line is drawn to an intersection of the vertical line value of Ω_s. The minimum complexity or order, n, will be the n curve that passes through or above this intersection. In our example, the order n equals 3. This implies that the low pass prototype filter will have 3 poles and, consequently, the final bandpass filter will have 3 pole-pairs.
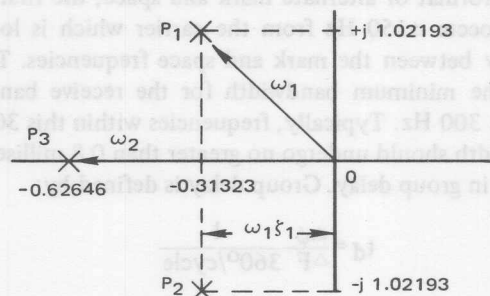
Table 2 gives the pole locations and quadratic factors for a third order 0.5 dB passband ripple Chebyshev low pass filter.

The values obtained from Table 2 are:

–0.31323 ± j1.02193	Complex conjugate pole
–0.62646 + j0	Real pole
a_0 = 1.14245	Characteristic of non s term
a_1 = 0.62646	Characteristic of s term

where the s term equation = $(s^2 + a_1 s + a_0)$

Step (4) — Lowpass Prototype Filter Natural Frequencies and Damping Factors

Using the following relationships, solve for the natural frequencies (ω) and damping factors (ξ):

$$\omega_1^2 = (1.02193)^2 + (-0.31323)^2 \qquad (2)$$
$$\omega_1 = 1.069$$

$$\text{also, } \omega_1 \xi_1 = 0.31323 \qquad (3)$$
$$\xi_1 = \frac{0.31323}{1.069}$$
$$\xi_1 = 0.293$$
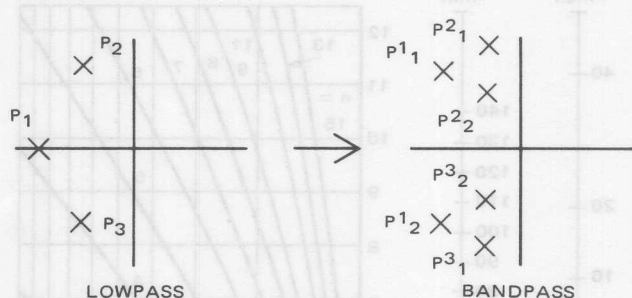
$$\omega_2^2 = (0)^2 + (-0.62646)^2$$
$$\omega_2 = 0.62646$$

$$\text{also, } \omega_2 \xi_2 = 0.62646$$
$$\xi_2 = 1$$

Step (5) — Filter Section Q and Center Frequency

The complex conjugate pole of the low pass prototype is transformed into a pair of complex conjugate bandpass poles, whereas the real pole of the low pass prototype is transformed into a complex conjugate pair of bandpass poles.

The bandpass filter will take on a form of three 2-pole bandpass filter sections in cascade. When bandpass sections are cascaded, each section center frequency and Q must be determined from the low pass damping factors (ξ) and natural frequencies (ω).

Given:
ω_1 = 1.069, ξ_1 = 0.293
F_1 = 970 Hz, F_2 = 1370 Hz
Then:
$$F_0 = \sqrt{F_1 F_2} = 1152.78 \text{ Hz (geometric center)} \quad (4)$$

$$Q_0 = \frac{F_0}{F_2 - F_1} = \frac{1152.78 \text{ Hz}}{400} \quad (5)$$

$$Q_0 = 2.8819$$

Section Q:

$$Q_1 =$$ (6)

$$\left[\frac{\left[\left(\frac{\omega_1}{Q_0}\right)^2 + 2\right] + \sqrt{\left\{\left[\left(\frac{\omega_1}{Q_0}\right)^2 + 2\right] + 2\right\}^2 - 4\left(\frac{2\xi_1\omega_1}{Q_0}\right)^2 + 2}}{2\left(\frac{2\xi_1\omega_1}{Q_0}\right)^2}\right]^{1/2}$$

Yielding:

$$Q_1 = 9.345$$

Section 2 is a reflected image about F_0 of section 1 for a 3 section cascaded filter (odd order). Recall that a third order low pass when transformed to a bandpass results in two pairs of complex poles (sections 1 and 2) from the low pass complex pole and one pair of complex poles (section 3) from the low pass real pole.

$$Q_1 = Q_2 = 9.345$$

For section 3:

$$Q_3 = \frac{Q_0}{\xi_2\omega_2} = \frac{2.882}{(1)(0.627)} = 4.596 \quad (7)$$

Center Frequencies:

$$F_1 = MF_0 \quad (8)$$

where:

$$M = \frac{\xi_1\omega_1 Q_1}{Q_0} + \sqrt{\left(\frac{\xi_1\omega_1 Q_1}{Q_0}\right)^2 - 1} \quad (9)$$

$$M = \frac{(0.293)(1.069)(9.345)}{2.882} + \sqrt{\left[\frac{(0.293)(1.069)(9.345)}{2.882}\right]^2 - 1}$$

$$M = 1.1932$$

$$F_1 = (1.1932)(1152.78) = 1375.52 \text{ Hz}$$

The image F_2 becomes:

$$F_2 = \frac{1}{M} F_0 = \frac{1152.78}{1.1932} = 966.1 \text{ Hz} \quad (10)$$

For section 3 the center frequency is:

$$F_3 = F_0 = 1152.78 \text{ Hz} \quad (11)$$

Step (6) — Center Frequency Loss and Filter Passband Gain

The gain produced by the active elements in the band-pass filter should overcome loss due to the stagger tuned filter sections. Each section of a cascade bandpass filter, except the section centered about ω_0, has a loss as represented by Equation 12. The overall filter center angular frequency ω_0 (Equation 13), section Q, and section center angular frequency ω_n (Equation 14) are required to determine each section's center frequency loss. Once the individual losses are determined, they are summed to arrive at the total cascaded filter loss A_{VO} ($j\omega_0$).

This value is used in determining filter section gain such that the designed bandpass filter meets design gain goals. The receive filter block must amplify the minimum input line signal to a minimum required limiter input signal.

$$AVO_n (j\omega_0) \text{ dB loss} = 20 \log \frac{\frac{\omega_n\omega_0}{Q_n}}{\sqrt{(\omega_n^2 - \omega_0^2)^2 + \left(\frac{\omega_n\omega_0}{Q_n}\right)^2}} \quad (12)$$

$$\omega_0 = 2\pi\sqrt{F_1 F_2} \quad (13)$$

$$\omega_n = 2\pi F_n \quad (14)$$

The following will illustrate the use of Equation 12 to solve for the center frequency loss of the modem answer filter example.
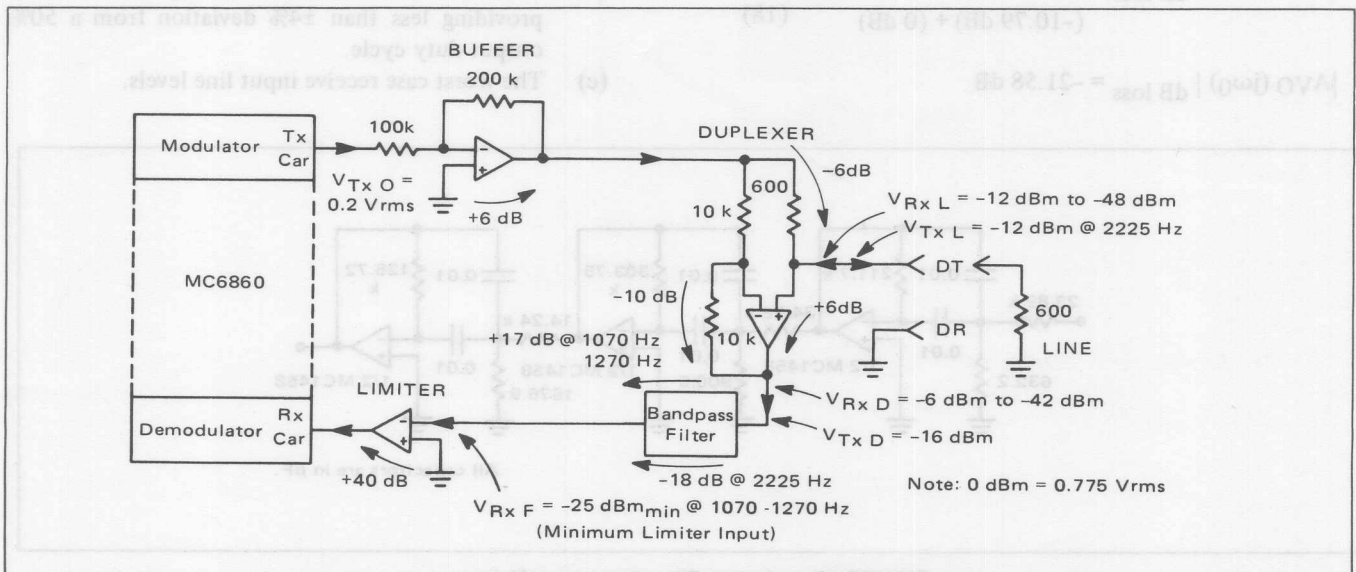

FIGURE 12 — System Level Constraints

Section 1

$$\omega_0 = 2\pi \sqrt{(970)(1370)} = 7.2431 \times 10^3 \text{ rad/s} \qquad (15)$$

$$\omega_1 = 2\pi (1375.52) = 8.6426 \times 10^3 \text{ rad/s} \qquad (16)$$

$$Q_1 = 9.345$$

$$|A_{VO1}(j\omega_0)|_{\text{dB loss}} = 20 \log \left[\frac{\dfrac{(8.6426 \times 10^3)(7.243 \times 10^3)}{9.345}}{\sqrt{\left[(8.642 \times 10^3)^2 - (7.243 \times 10^3)^2\right]^2 + \left[\dfrac{(8.642 \times 10^3)(7.243 \times 10^3)}{9.345}\right]^2}} \right] \qquad (17)$$

$$|A_{VO1}(j\omega_0)|_{\text{dB loss}} = 20 \log (0.2886)$$

$$|A_{VO1}(j\omega_0)|_{\text{dB loss}} = -10.794 \text{ dB}$$

Section 2

$$\omega_0 = 7.243 \times 10^3 \text{ rad/s}$$

$$\omega_2 = 2\pi (966.1) = 6.07 \times 10^3 \text{ rad/s}$$

$$Q_2 = 9.345$$

$$|A_{VO2}(j\omega_0)|_{\text{dB loss}} = 20 \log (0.2886)$$

$$|A_{VO2}(j\omega_0)|_{\text{dB loss}} = -10.794 \text{ dB}$$

Section 3

$$\omega_0 = 7.243 \times 10^3 \text{ rad/s}$$

$$\omega_3 = 7.243 \times 10^3 \text{ rad/s}$$

$$Q_3 = 4.596$$

$$|A_{VO3}(j\omega_0)|_{\text{dB loss}} = 20 \log (1)$$

$$|A_{VO3}(j\omega_0)|_{\text{dB loss}} = 0 \text{ dB, due to } \omega_n = \omega_0$$

The total filter center frequency loss is equal to the sum of all sectional losses.

$$|A_{VO}(j\omega_0)|_{\text{dB loss}} = (-10.79 \text{ dB}) + $$
$$(-10.79 \text{ dB}) + (0 \text{ dB}) \qquad (18)$$

$$|A_{VO}(j\omega_0)|_{\text{dB loss}} = -21.58 \text{ dB}$$

Figure 12 illustrates the design goals that are used to determine the receive filter passband gain for the answer only modem system. The answer filter provides 35 dB of attenuation to 2225 Hz relative to the filter passband. This results in –34 dBm of unwanted signal level being present at the limiter input. To maintain a probability of error $(P_e) \leqslant 1 \times 10^{-5}$, a signal-to-noise ratio at the limiter input must be greater than +12.12 dB. The theoretical probability of error (P_e) curve for non-coherent FSK is determined by:

$$P_e = 1/2 \, e^{ - \left[\dfrac{\left(\dfrac{V_s}{V_n}\right)^2 \left(\dfrac{BW_n}{BW_s}\right)}{2} \right] } \qquad (19)$$

where V_s = signal level
V_n = noise level
BW_n = noise bandwidth (400 Hz)
BW_s = signal bandwidth (300 Hz)

In calculating the voltage gain required by the receive active filter block, the following constraints should be considered:

(a) The signal to noise performance required by the modem system.

(b) The receive limiter minimum input level while providing less than ±4% deviation from a 50% output duty cycle.

(c) The worst case receive input line levels.
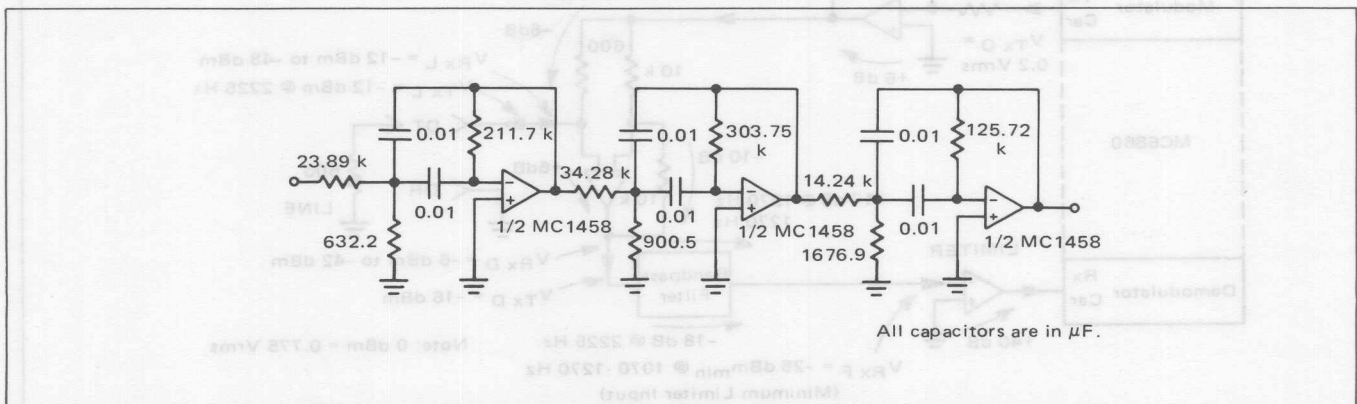

All capacitors are in µF.

FIGURE 13a – Answer Filter Component Values

(d) At the maximum input line levels, the designed filter gain should not saturate any active stage of the filter.
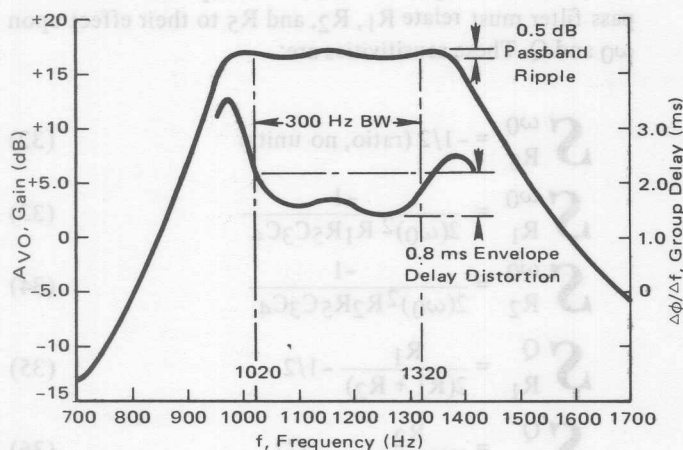

FIGURE 13b — Answer Filter Gain and Group Delay

The use of the MLM311 as a receive signal limiter provides 40 dB of signal gain while maintaining a limited output level having less than ±2% deviation from a 50% duty cycle with a –25 dBm applied input level ($V_{Rx\ F}$).

The telephone line receive level for the answer only example ranges between –12 dBm and –48 dBm. An active duplexer provides 6 dB of signal gain to these line levels resulting in filter input levels ($V_{Rx\ D}$) between –6 dBm and –42 dBm.

From the above information, the active filter must provide the following passband gain.

$$A_{VO} = |V_{Rx\ Dmin}| - |V_{Rx\ Fmax}| \quad (20)$$
$$A_{VO} = 42\ dB - 25\ dB = 17\ dB\ passband\ gain$$

The amount of operational amplifier gain used in the filter design is based on both the passband gain requirements and the filter center frequency loss.

$$A_{VOtotal} = |A_{VO}\ (passband)| + |A_{VO}\ (center\ frequency\ loss)| \quad (21)$$

$$A_{VOtotal} = 17\ dB + 21.58\ dB = +38.58\ dB$$

This requires that each of the three filter sections provide a gain of:

$$A_{VO} = \frac{+38.58\ dB}{3} = +12.86\ dB\ or\ 4.41\ volts/volt. \quad (22)$$

Step (7) – Filter Component Values

Now that each section gain, center frequency, and design Q is known, the actual filter component values can be calculated (reference Figure 9).

Section 1:

$F_1 = 1375.52$ Hz
$\omega_1 = 8.6426 \times 10^3$ rad/s
$Q_1 = 9.345$
$A_{VO1} = 4.41$ (gain of section)

$C_3 = C_4 = 0.01\ \mu F$ (using equal value capacitors)

$$R_5\ (uncorrected) = \frac{2Q_1}{\omega_1 C} = \frac{2\ (9.35)}{2\pi(1375.5)\ (1 \times 10^{-8})}$$
$$= 216.4\ k\Omega \quad (23)$$

$$R_1\ (uncorrected) = \frac{R_5}{2\ A_{VO1}} = \frac{216.4\ k}{2(4.41)}$$
$$= 24.5\ k\Omega \quad (24)$$

$$R_2\ (uncorrected) = \frac{R_1 R_5}{4Q_1{}^2 R_1 - R_5}$$
$$= \frac{(24.5\ k)\ (216.4\ k)}{4(9.35)^2\ (24.5\ k) - 216.4\ k}$$
$$= 634.9\ \Omega \quad (25)$$

These three resistor values, if used to initially implement the first bandpass section, would not produce exact design goals. Filter response will shift due to non-ideal operational amplifier parameters such as dc gain (A_{VOL}), gain bandwidth product (GBW), and input impedance (z_{in}).

To offset any shift in filter response, new values for selection Q, gain and frequency should be calculated taking into account the operational amplifier parameters. These corrected values will be used to obtain new values for R_5, R_1, and R_2, resulting in a filter response very near design goals.

Corrected values for ω_n, Q_n, and A_{VOn} are calculated using the following MC1458 operational amplifier parameters.

$A_{VOL} = 1 \times 10^5$ volts/volt
GBW $= 1 \times 10^6$ Hz, 6.283×10^6 rad/s
z_{in} $= 1 \times 10^6$ ohms

$$\omega_{C1} = \frac{\omega_1}{1 - Q_1\left(\dfrac{\omega_1}{GBW}\right)} \quad (26)$$

$$\omega_{C1} = 8.755 \times 10^3\ rad/s,\ 1393.4\ Hz$$

$$Q_{C1} = \frac{Q_1}{1 - Q_1\left[\dfrac{2Q_1}{A_{VOL}} + \left(\dfrac{R_5}{z_{in}} - 1\right)\dfrac{\omega_1}{GBW}\right]} \quad (27)$$

Plugging in values we obtain:

$$Q_{C1} = 9.27$$

$$A_{VOC1} = \frac{A_{VO1}}{1 - Q_1\left[\dfrac{2Q_1}{A_{VOL}} + \left(\dfrac{R_5}{z_{in}}\right)\dfrac{\omega_1}{GBW}\right]} \quad (28)$$

$$A_{VOC1} = 4.43$$

Using these corrected values of section center frequency, Q, and section gain, solve for the corrected values of R_1, R_2, and R_5:

$$R_5 = \frac{2QC_1}{\omega C_1 C} \qquad (29)$$

$$R_5 = \frac{2(9.27)}{(8.755 \times 10^3)(1 \times 10^{-8})} = 211.7 \text{ k}\Omega$$

$$R_1 = \frac{R_5}{2A_{VO}C} \qquad (30)$$

$$R_1 = \frac{2.117 \times 10^5}{2(4.43)} = 23.89 \text{ k}\Omega$$

$$R_2 = \frac{R_1 R_5}{4QC_1^2 R_1 - R_5} \qquad (31)$$

$$R_2 = \frac{(2.389 \times 10^4)(2.117 \times 10^5)}{4(9.27)^2(2.389 \times 10^4) - 2.117 \times 10^5}$$

$$R_2 = 632.2 \ \Omega$$

Section 2:

$F_2 = 966.1$ Hz
$\omega_2 = 6.07 \times 10^3$ rad/s
$Q_2 = 9.345$
$A_{VO2} = 4.43$
$C_3 = C_4 = 1 \times 10^{-8}$ F

Solving as in Section 1 using Equations 23 through 31, we obtain:

$\omega_{C2} = 6.1255 \times 10^3$ rad/s, 974.9 Hz
$Q_{C2} = 9.30$
$A_{VOC2} = 4.43$
$R_5 = 303.75$ kΩ
$R_1 = 34.28$ kΩ
$R_2 = 900.5 \ \Omega$

Section 3:

$F_3 = 1152.73$ Hz
$\omega_3 = 7.243 \times 10^3$ rad/s
$Q_3 = 4.596$
$A_{VO3} = 4.41$
$C_3 = C_4 = 1 \times 10^{-8}$ F

Solving as in section 1 and 2, we obtain:

$\omega_{C3} = 7.281 \times 10^3$ rad/s, 1158.87 Hz
$Q_{C3} = 4.58$
$A_{VOC3} = 4.41$
$R_5 = 125.72$ kΩ
$R_1 = 14.24$ kΩ
$R_2 = 1676.9 \ \Omega$

The complete answer filter is shown in Figure 13a with the filter response and envelope delay curves shown in Figure 13b. If the filter is not optimum after construction, it may be fine tuned by the following method.

In tuning filters, one of the most useful parameters is the sensitivity of the filter to element variations. Sensitivity is defined as a measure of the dependence of a network upon the change of some parameter of the network. The sensitivities of importance to the multiple-feedback band-pass filter must relate R_1, R_2, and R_5 to their effect upon ω_0 and Q. These sensitivities are:

$$S_{R_5}^{\omega_0} = -1/2 \ (\text{ratio, no units}) \qquad (32)$$

$$S_{R_1}^{\omega_0} = \frac{-1}{2(\omega_0)^2 R_1 R_5 C_3 C_4} \qquad (33)$$

$$S_{R_2}^{\omega_0} = \frac{-1}{2(\omega_0)^2 R_2 R_5 C_3 C_4} \qquad (34)$$

$$S_{R_1}^{Q} = \frac{R_1}{2(R_1 + R_2)} - 1/2 \qquad (35)$$

$$S_{R_2}^{Q} = \frac{R_2}{2(R_1 + R_2)} - 1/2 \qquad (36)$$

$$S_{R_5}^{Q} = +1/2 \qquad (37)$$

In practice, $R_1 \gg R_2$ such that

$$S_{R_1}^{Q} \to 0$$

$$S_{R_2}^{Q} \to -1/2$$

These sensitivities imply that to change section Q, R_2 should be adjusted. If R_2 were increased, for example 20%, section Q will decrease 10%. Notice that the sensitivity of Q to changes in R_2 and R_5 is equal and opposite in magnitude. This implies that if R_2 and R_5 are changed by the same percentage, but in opposite directions, section Q will not change. Also, as R_5 is adjusted, it changes the section center frequency by a ratio of $-1/2$.

Filter Tuning Procedure
Section Center Frequency:
(a) Increase/decrease R_5 for a corresponding decrease/increase in section center frequency ω_0.
(b) Increase/decrease R_2 by the same percentage of increase/decrease applied to R_5 in step (a) to maintain constant section Q.

Section Q:
(a) Increase/decrease R_2 for a corresponding decrease/increase in section Q.

ORIGINATE FILTER DESIGN
Basically, the originate receiving filter design procedures are identical to the answer filter example. The one major difference is that the filter center frequency is shifted to accept 2025 — 2225 Hz signals. One might also note that the second harmonics of the local transmit signals in the originate mode (1070 — 1270 Hz) fall within and just

outside of the passband for the originate receive filter. For this reason, the originate only modem designer may want to provide a transmit bandpass filter to suppress harmonics produced by the local transmit carrier (see Figure 15).

The three section design parameters and component values for the 6-pole originate receive filter are:

Section 1:

$F_1 = 2425.81$ Hz
$Q_1 = 16.56$
$A_{VO1} = 4.48$
$C_3 = C_4 = 1 \times 10^{-8}$F
$R_1 = 24.26$ kΩ
$R_2 = 199.76$ Ω
$R_5 = 217.258$ kΩ

Section 2:

$F_2 = 1985.62$ Hz
$Q_2 = 16.67$
$A_{VO2} = 4.48$
$C_3 = C_4 = 1 \times 10^{-8}$F
$R_1 = 29.85$ k
$R_2 = 242.36$ Ω
$R_5 = 267.23$ kΩ

Section 3:

$F_3 = 2154.01$ Hz
$Q_3 = 8.32$
$A_{VO3} = 4.43$
$C_3 = C_4 = 1 \times 10^{-8}$F
$R_1 = 13.88$ kΩ
$R_2 = 458.85$ Ω
$R_5 = 122.913$ kΩ

The complete 6-pole receive originate filter is shown in Figure 14a, with the response and envelope delay curves shown in Figure 14b.

8-POLE, –50 dB RECEIVE AND 4-POLE, –25 dB TRANSMIT FILTER DESIGN

A complete full duplex modem system will most likely require operation with input signals down to –50 dBm at the line input. This requires a receive filter network having at least 8 poles to provide the necessary attenuation to adjacent duplex channel interference and a local transmit filter having 4 poles to provide 25 dB local transmit signal harmonic rejection. The construction of an 8-pole or 4-pole filter takes on the same cascaded form as the illustrated

6-pole design example. Therefore, only the component values for the 8-pole and 4-pole filters are tabulated in Figure 15 without the individual circuit diagrams.
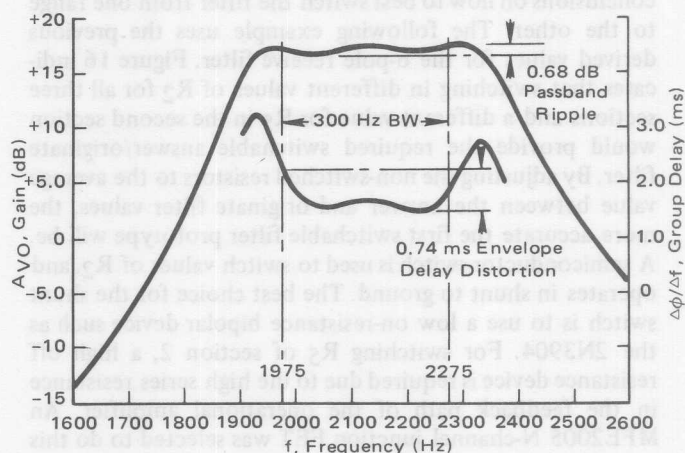

FIGURE 14b — Originate Filter Gain and Group Delay

RECEIVE ORIGINATE

Section	1	2	3	4
R_1 (Ω)	31.42 k	39.54 k	14.71 k	16.1 k
R_2 (Ω)	146.8	181.15	396.29	432.32
R_5 (Ω)	288.64 k	363.27 k	132.15 k	144.66 k

RECEIVE ANSWER

Section	1	2	3	4
R_1 (Ω)	31.08 k	46.34 k	14.51 k	17.1 k
R_2 (Ω)	468.48	690.57	1397.94	1643.88
R_5 (Ω)	283.33 k	422.31 k	131.38 k	154.8 k

TRANSMIT ORIGINATE

Section	1	2
R_1 (Ω)	15.73 k	20.56 k
R_2 (Ω)	1218.55	1586.55
R_5 (Ω)	130.47 k	170.47 k

TRANSMIT ANSWER

Section	1	2
R_1 (Ω)	16.17 k	18.78 k
R_2 (Ω)	366.95	423.79
R_5 (Ω)	133.25 k	154.81 k

Note: All Capacitors = 0.01 µF

FIGURE 15 — 8-Pole, –50 dB Receive and 4-Pole, –25 dB Transmit Filter Values

FIGURE 14a — Originate Filter Component Values

AUTOMATIC ANSWER/ORIGINATE MODEM SYSTEM

The filter design for a fully automatic answer/originate modem system must have switchable bandpass characteristics. By tabulating the previous component values for both the answer and originate filters, one can draw some conclusions on how to best switch the filter from one range to the other. The following example uses the previous derived values for the 6-pole receive filter. Figure 16 indicates that switching in different values of R_2 for all three sections and a different value for R_5 in the second section would provide the required switchable answer/originate filter. By adjusting the non-switched resistors to the average value between the answer and originate filter values, the more accurate the first switchable filter prototype will be. A semiconductor switch is used to switch values of R_2, and operates in shunt to ground. The best choice for the shunt switch is to use a low on-resistance bipolar device such as the 2N3904. For switching R_5 of section 2, a high off resistance device is required due to the high series resistance in the feedback path of the operational amplifier. An MFE2005 N-channel junction FET was selected to do this job. Figure 17a illustrates the fully automatic answer/originate switchable filter system. Also shown are the transmit buffer, duplexer, threshold detector, limiter, and mode control level translator sections. The level translator, which provides the correct on/off voltage levels to the bipolar FET switches, receives its answer/originate command from the MC6860 modem mode control output pin.

The measured response and envelope delay for the switchable 6-pole receive filter design is shown in Figure 17b.

Figure 18 illustrates the complete modem system with the RS-232 interface to the CBS data coupler, and the direct interface to a CBT data coupler. Automatic disconnect option inputs are handled by PC board mounted switches. The complete automatic modem, less the power supply, may be easily constructed on a single 4 x 5 printed circuit board.
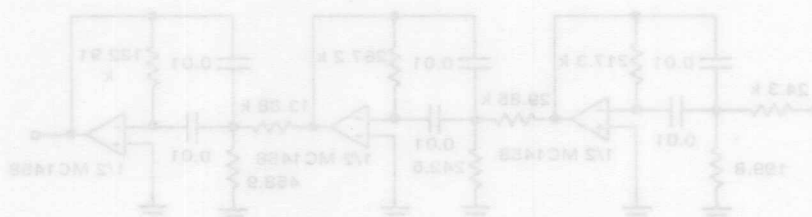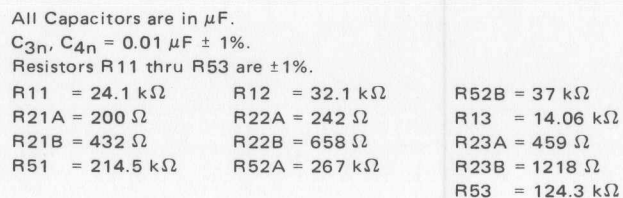
CONCLUSION

A low-speed modem design has been presented using the MC6860 LSI MOS digital Modem integrated circuit. Included has been a system design example using filter design tables and equations to develop a complete modem system. Also included have been component values for filter designs which may be used to develop full duplex modem systems.

The availability of this LSI modem circuit along with the presented filter designs should provide a very useful building block for the OEM modem and terminal designers by providing him precise digital modulation, demodulation, and supervisory control. The modem designer will find that a design approach using the MC6860 modem will also provide an impressive system size reduction as well as a better price-performance choice for his present and future low speed modem designs.

Resistor	Answer 1070-1270 Hz	Originate 2025-2225 Hz	Average or $\triangle$ Value	Answer Switched	Originate Switched
R_{11}	23.89 k	24.26 k	24.08 k	24.1 k	24.1 k
R_{21}	632.2	199.76	$\triangle$ 432.4	632	200
R_{51}	211.7 k	217.26 k	214.48 k	214.5 k	214.5 k
R_{12}	34.28 k	29.85 k	32.07 k	32.1 k	32.1 k
R_{22}	900.5	242.36	$\triangle$ 658.2	900	242
R_{52}	303.75 k	267.23 k	$\triangle$ 36.5 k	304 k	267 k
R_{13}	14.24 k	13.88 k	14.06 k	14.06 k	14.06 k
R_{23}	1676.9	458.85	1218.05	1677	459
R_{53}	125.72 k	122.91 k	124.32 k	124.3 k	124.3 k

FIGURE 16 — Switchable Modem Filter Values

FIGURE 17a — Switchable Filter/Duplexer

All Capacitors are in μF.
C_{3n}, $C_{4n} = 0.01 \mu F \pm 1\%$.
Resistors R11 thru R53 are $\pm 1\%$.

R11 = 24.1 kΩ	R12 = 32.1 kΩ	R52B = 37 kΩ
R21A = 200 Ω	R22A = 242 Ω	R13 = 14.06 kΩ
R21B = 432 Ω	R22B = 658 Ω	R23A = 459 Ω
R51 = 214.5 kΩ	R52A = 267 kΩ	R23B = 1218 Ω
		R53 = 124.3 kΩ

FIGURE 17b — Switchable Filter Response

FIGURE 18 — Modem System