POWERLINE INTERCOM

PROJECT REPOR

Submitted B

M.MANOHARA

R.AR

C.KUMA

M.MOHA

Guided By
Ms. VASUKI, M.E.

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Department Of Electronics and Communication Engineering

KUMARAGURU COLLEGE OF TECHNOLOGY

COIMBATORE - 641 006

DEPARTMENT OF ELECTRONICS AND COMMUNICATION ENGINEERING

KUMARAGURU COLLEGE OF TECHNOLOGY

COIMBATORE - 641 006

CERTIFICATE

This is to Certify that the Project report entitled

" POWERLINE INTERCOM "

has been submitted by

Mr. M.MANOHARAN, R.ARUL, C.KUMAR, M.MOHAN.

in partial fulfillment for the award of the Degree of Bachelor of Engineering in the Electronics and Communication Engineering Branch of the Bharathiar University, Coimbatore during the academic year 1999 - 2000.

Guide *Head of the Department*

Certified that the candidate with the University Registration No.................................. was examined in the project work viva-voce held on 16/03/2000

Internal Examiner *External Examiner*

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SYNOPSIS

The power line carrier communication is a method in which the information signals are sent through the power line. In this project the voice signal is fed to the AC socket and the same signal can be received from any other AC socket. The voice signal from the speakers telephone is amplified to the level of modulation requirements, frequency modulated and then transmitted through the power line. At the receiving end it is demodulated and fed to the receiver in the listeners telephone. The selection of called terminal is done through switching circuit from the caller's telephone. The coupling between telephone and the AC socket is made through a transformer and a coupling capacitor. The coupling technique used here is phase to ground technique. The same operation when repeated with the transmitter and receiver interchanged provides full duplex operation. At the receiving end using the same coupling technique as the transmitter, the signal is fed to the demodulator from the receiving AC socket. The reconstructed signal is amplified and then given to the called telephone.

CONTENTS

1. CHAPTER

INTRODUCTION

Nowadays communication becomes essential for everyone. Several methods commonly used for communication include optical, RF, microwave, hardwire link and infrared. Another alternative, a close relative of the hardwire link communication is power line carrier communication. The term power line carrier communication refers to sending the carrier signal of high frequency and low voltage along the power line which carries low frequency and high voltage signal of the various available communication systems telephone is the most widely used communication device.

In general the necessary control signals like ringing tone, busy tone, off hook tone and the speech signal are transmitted through the telephone cables. This project aims at sending these control signals and the voice signals through the power line. Such a technique will tremendously reduce the cost of the overall system and hence is highly attractive from the commercial point of view.

The AC power line makes a good communication link everywhere in the world; no new installations of hard wire link are required since most of the buildings are already wired. The link for communication between two points can be established any where, where there is an AC socket by simply plugging in a line card. Although this technique seems to be attractive, this is seldom used since the power line contains 230 volts, 50Hz signal but also some spikes whose amplitude is in Kilovolts. Both of which present fundamental design challenges for carrier communication through the power lines. Meeting the challenges necessarily leads to complex circuit designs.

Here, the voice signal at the transmitting side is converted into electrical signal. The microphone does this conversion, which is in the telephone. This electrical signal, which is of normally small amplitude, is amplified to the level of modulation requirements. Of the various techniques of modulation available, frequency modulation is used here because its power level is constant and relatively immune to noise. The frequency-modulated signal is again amplified in order to satisfy the power line constraints and fed to the power line through the AC socket by using coupling equipments. The coupling equipments are coupling capacitors, band pass filter and a transformer.

Similarly at the receiving end the frequency modulated signal is received via coupling equipments as at the transmitting side and the 50Hz power line signal is rejected. The signal is then fed to the demodulator, which produces the original signal, which was transmitted. The demodulated signal is fed to the called telephone set. To select a particular terminal the control signals from the telephone set, the DTMF receiver and the switching circuits are used. After selecting the particular terminal a link between the calling terminal and called terminal is established.

2. CHAPTER

POWER LINE

2.1. INTRODUCTION

A transmission line is set of conductors being run from one place to another. Such line, therefore have four distributed parameters:

a) Series resistance

b) Inductance

c) Shunt capacitance and

d) Shunt Conductance

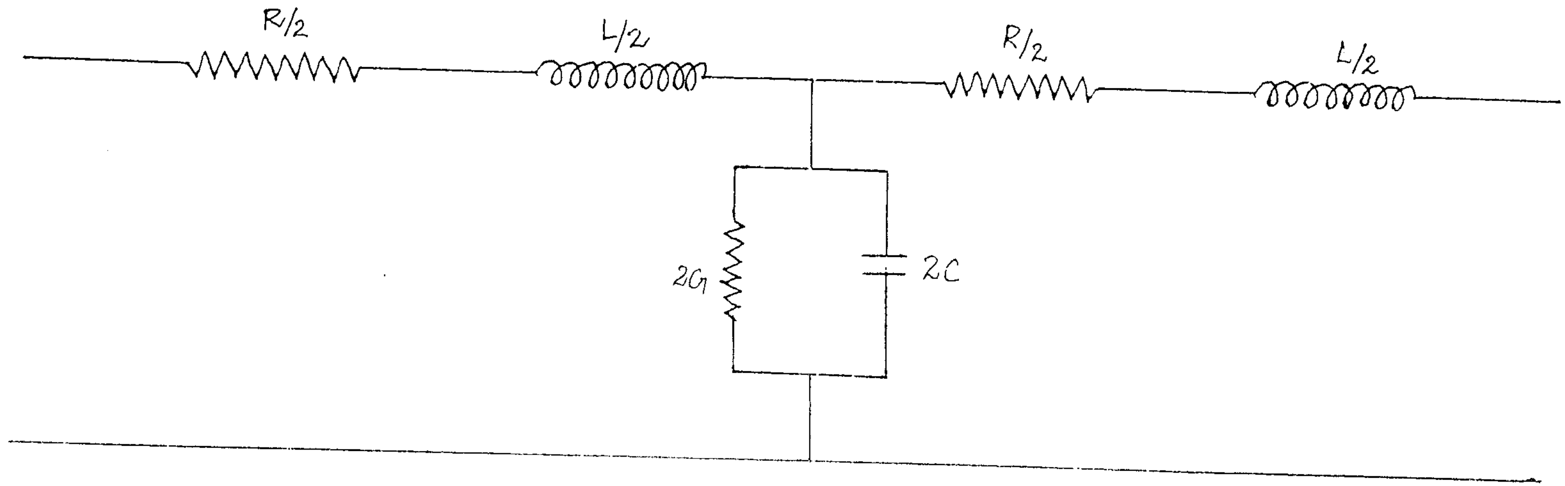
The voltage and currents vary harmonically along the line with respect to the distance of the point under consideration. The electrical power is transmitted over the power lines at approximately the speed of light.

The equivalent circuit of unit section of power line may be represented as shown in fig. 2.1

Actually there are no lumped constants but instead resistance, inductance, capacitance and conductance are distributed along the whole length of the line.

If a long line consisting of two parallel uniform conductors carries current there is a magnetic field around the conductors and a voltage drop along them. A magnetic field, which is proportional to current, indicates that the line has series inductance and voltage drop indicates

Fig. 2.1. EQUIVALENT CIRCUIT OF POWER LINE



Where,

R - Series Resistance,

L - Series Inductance.

G - Shunt Conductance,

C - Shunt Capacitance.

the presence of a series resistance. Voltage applied across the conductor produce an electric field between the conductors and also charges on them. This indicates that the line contains a shunt capacitance and since this capacitance is never loss less or perfect, it will have some shunt conductance.

2.2. NOISE

The noise on the line can broadly put into two categories, which are referred as broadband and impulse noise. Broadband noises come from many sources with universal motors being one notorious example. The level varies from a few hundreds of millivolts to some kilovolts. Impulse noise also varies widely in a level ranging from millivolts to tens of volts and sometimes much more. In general, noise is worst in factories, moderate in offices and least in private concerns.

2.3. FREQUENCY OF OPERATION

The CCITT recommendations provide for transmission of 300Hz to 3400Hz voice signals. The usual frequency for operation of power line carrier is between 10 kHz to 450 kHz; also the actual range varies from country to country and sometimes from administration to administration. For example, in Japan the range is from 10 to 450 kHz, in Germany it is 15 to 375 kHz. Whereas in the USA it is 30 to 200 kHz only.

2.4. ATTENUATION

Line attenuation is very difficult to estimate because it is extremely load dependent. A high power load can significantly reduce the impedance of the line at the point of connection and thus dominate attenuation for all points of communication that occur beyond the offending load unless that load is isolated with chokes. Another large component of the net attenuation can be the signal loss incurred in coupling across the multiple windings of a power distribution transformer. This alone can amount to 20 to 40 dB, depending on the carrier frequency and transformer construction. In extreme cases of attenuation, frequency-translating repeaters can be used to boost signal levels.

2.5. ADVANTAGE OF POWER LINE COMMUNICATION

When the distance involved is large it will not be economical to provide separate wires for communication purposes. In fact for such large distances, the power lines themselves provide a very good medium of transmission of information.

Its advantages are:

(a) No separate wires are needed for communication purpose hence the cost of constructing separate lines is saved.

(b) When compared with ordinary lines the power lines have appreciably higher mechanical strength.

(c) Power line has large cross sectional area resulting in very low resistance per unit length. Consequently the carrier signals suffer much less attenuation than when they travel on user telephone lines of equal length.

(d) Power lines are well insulated to provide only negligible leakage between conductors and ground even in adverse weather conditions.

3. CHAPTER

TELEPHONE

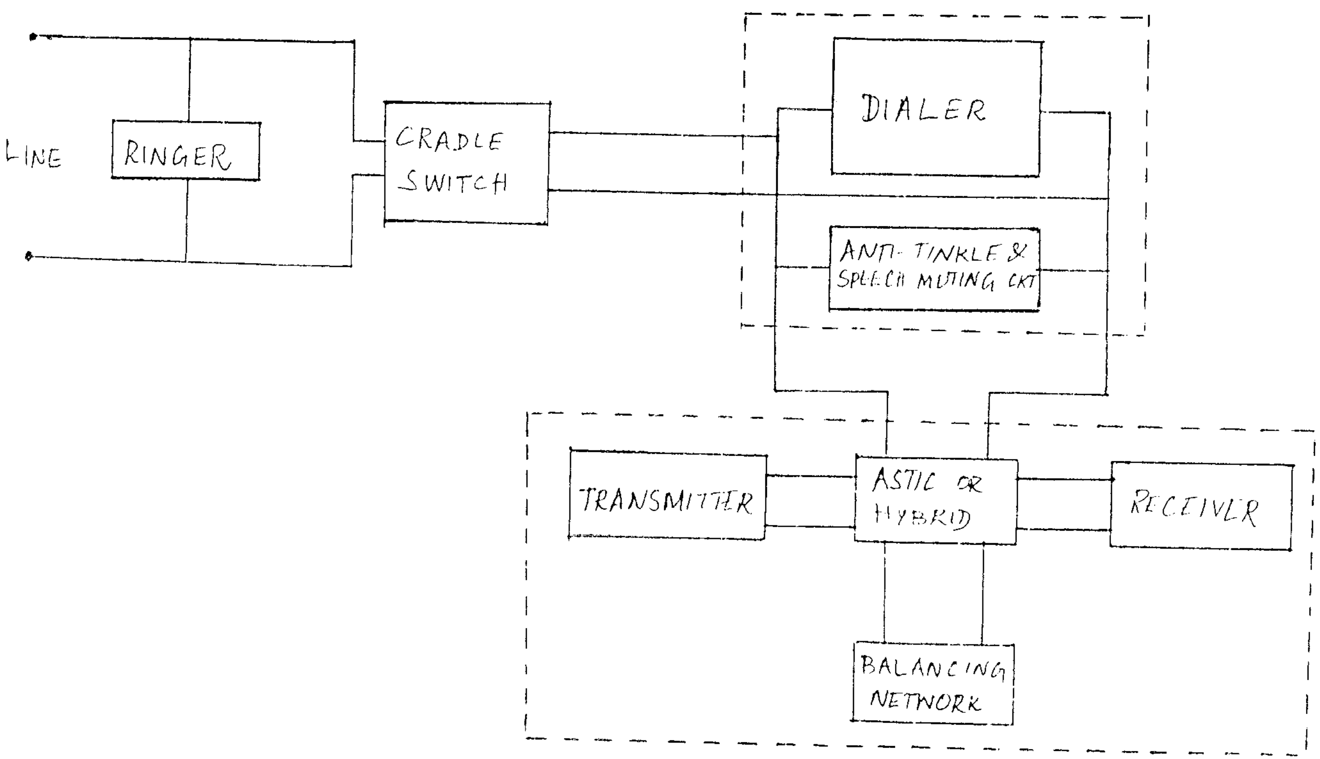
3.1. INTRODUCTION

A telephone set is a device with which a subscriber can send or receive a telephone call. It is the most widely used communication device. The basic building blocks and the operation of the telephone are dealt with in this chapter. Figure 3.1 gives the basic block diagram of a telephone.

3.2. RINGER

This circuit gives audible signals to attract the attention of the subscribers to an incoming telephone call. The ringer comprises of an electronic circuit that gives AF signals to a Piezo Transducer normally 75V AC at 17Hz to 133Hz. The AC ringing voltage from the line is applied to a bridge rectifier, which is integrated on a chip. Over voltage is protected by anti tapping and protection circuit.

Fig. 3.1. BLOCK DIAGRAM OF A TELEPHONE SET



3.3. CRADLE SWITCH

This switch is used to connect / disconnect the ringer and speech circuits to the line. When telephone handset is on the cradle the speech circuit is disconnected but ringer is connected to the line, so that the ringer can give an indication in case there is an incoming call. In the OFF-HOOK condition ringer is disconnected and the speech circuit is connected to the line.

3.4. DIALER

Electronic dialer circuit is employed for this purpose. Keypad is used for dialing the number. While dialing a number, speech circuit of the telephone remains disconnected. Similarly, the high voltage spikes produced during dialing may produce sound in the ringer. To prevent this a circuit known as anti-tinkle and speech muting circuit is employed. Dialing a number is done by DTMF generation. In this method each digit is assigned one frequency from both the lower group frequency and the higher group frequencies. A unique tone for each combination of low and high group of frequencies is transmitted over the transmission path of telephone line. For a keypad with 12 keys, 12 unique tone combinations are produced. For example if we press key 8, tones of 852 and 1336 Hz are generated.

3.4.1. ADVANTAGE OF DTMF

a) Dialing is very fast as compared to pulse dialing.

b) It is more compatible with electronically controlled exchanges.

c) After the call has been connected, it can be used for low speed data transmission.

3.5. SPEECH CIRCUIT

Speech circuit comprises a hybrid coil, which passes the incoming speech signals to the telephone receiver and passes the outgoing speech signals from the telephone transmitter to the line. Thus it converts a two-wire line circuit into a four-wire circuit required for telephone handset. The circuit includes the following:

a) Microphone amplifier

b) Receiver amplifier with gain control

c) Regulated power supply for the speech circuit.

The equivalent circuit of a speech circuit is shown in fig.3.5.

Z_{f1} and Z_{f2} together with balancing network Z_B perform the function of hybrid converting two-wire circuit into four wires. It also provides impedance matching under different operating condition.

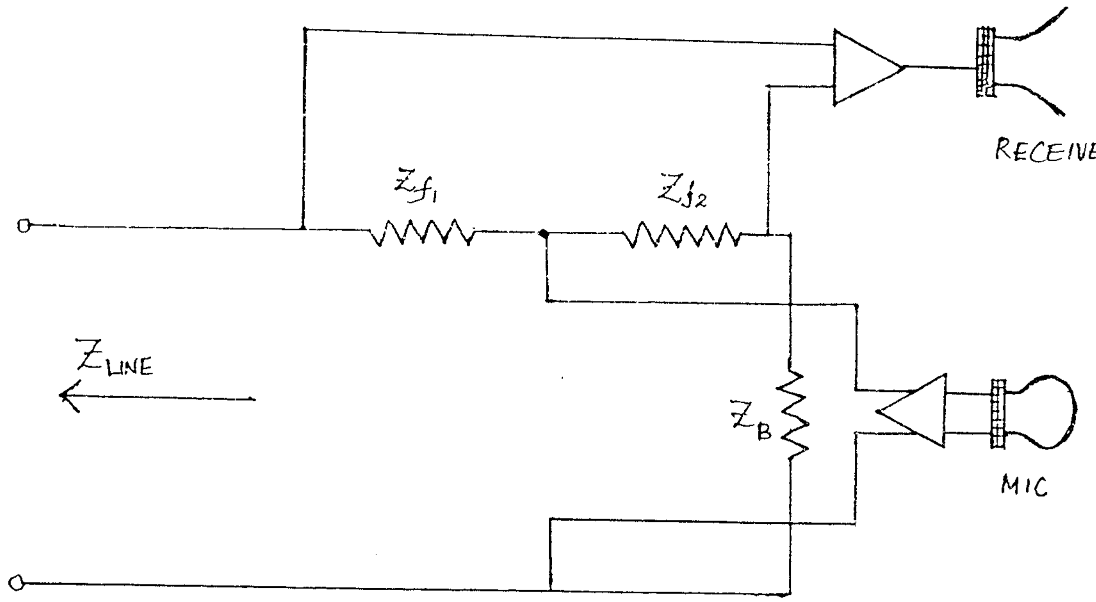


Fig. 3.5. EQUIVALENT CIRCUIT OF A SPEECH CIRCUIT

3.6. BALANCE CIRCUIT

This circuit is used to provide impedance matching between the line and the telephone handset under different conditions such as receiving / sending calls.

3.7. ANTI – SIDETONE INDUCTION COIL (ASTIC)

Signals produced by the microphone may be as much as 40dB higher than the incoming signals. These signals will feed into the receiver producing loud unwanted sound in the user's ear when he speaks. For induction of sidetone, antisidetone induction coil circuits are used.

3.8. PROTECTION CIRCUIT

The operation of a conventional telephone system is based upon the minimum operating current in the local loop. The carbon granule transmitter requires a current between 20mA to 23mA. The microphone has an equivalent resistance of about 400 ohms. When the telephone handset is on-hook the current must be reduced below the minimum loop current needed to operate.

These devices are prone to damage by high voltage spikes due to lightning, induction from ac power line, etc. Zener diodes are generally employed to provide protection against the high voltage and to provide protection against reverse polarity voltages a bridge rectifier is commonly included in all electronic telephones.

3.9. PARAMETERS OF TELEPHONE

3.9.1. DC VOLTAGE

DC potential across the telephone line when the handset is on-hook nc - 48 volts and -10 volts when the handset is off-hook.

3.9.2. LINE CURRENT

When the handset is off-hook the current drawn is about 50-60mA.

3.9.3. DC RESISTANCE

Typical value of the resistance introduced across the telephone line is 300 ohms.

3.9.4. DIAL TONE

It is a continuous tone of 400Hz modulated by a 33Hz signal. It indicates that lines are connected to a first free selector and the number can be dialed.

3.9.4. RING TONE

The ringer is activated by an electrical signal of 75 volts peak-to-peak amplitude and a frequency in the range 17-133Hz. It is made on for 0.4sec and off for 0.25 sec.

3.10. DIALING SIGNAL

3.10.1. DTMF DIALING

In dual tone multi frequency dialing each digit is represented by a combination of two sinusoidal signal of different frequency range i.e. for each button one low frequency and one high frequency signal are produced. The frequency for each button is shown in fig.3.10

3.10.2. PULSE DIALING

In this technique dialing is done through dial pulses, which consists of a train of breaks in the telephone loop current and number of breaks is equal to the numerical value of the digit dialed. The make time is 33.33ms and break time is 66.66ms.

3.11. SPECIAL FEATURES

3.11.1. PAUSE

This key is used to introduce an additional delay of 2 to 2.4 sec between two digits.

3.11.2. FLASH

This key produces a loop break of duration of 280 to 320 ms.

3.11.3. HOLD

This key holds the called party on the line.

3.11.4. BUSY TONE

When the called party is engaged, a busy tone is sent by the exchange to the calling party. It is an interrupted tone of 400Hz with equal on and off times.

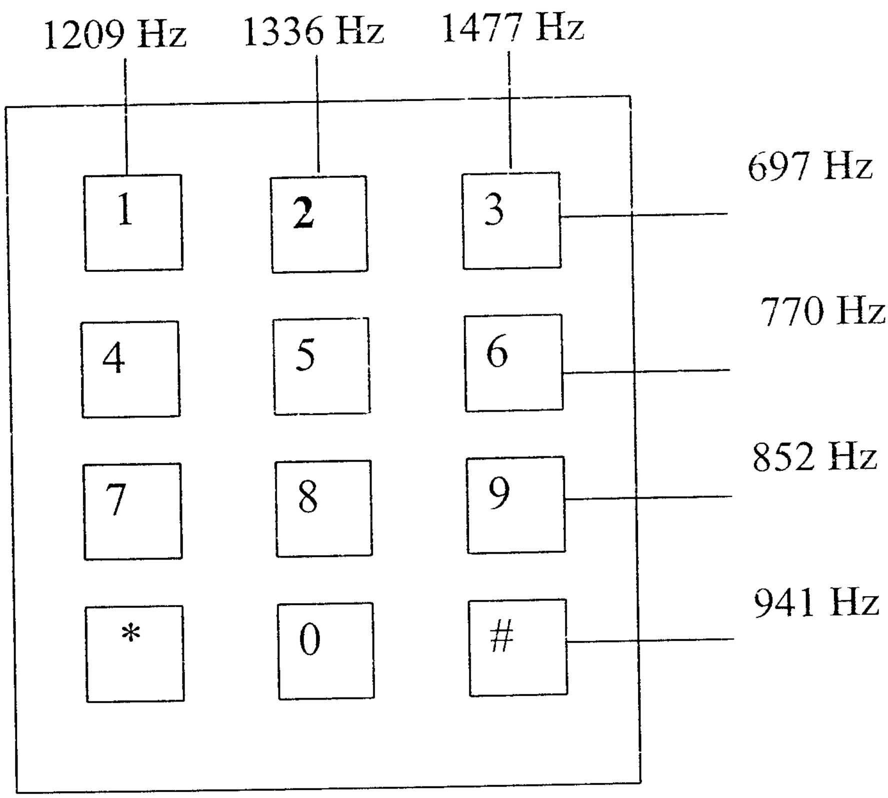


FIGURE 3.10 DTMF KEY PAD

4.CHAPTER

IC DESCRIPTION

4.1. INTRODUCTION

In modern communication system, the communication will be very difficult if IC's are not used. The IC which are used in this project are

a. IC XR 2206
b. IC XR 2211
c. IC 741
d. IC M – 8870

4.2. IC XR-2206

The modulation technique used here is frequency modulation. The IC used for this operation is XR 2206.The XR 2206 is a monolithic function generator integrated circuit capable of producing high quality sine, square, triangle, ramp waveforms of high stability and accuracy .The output waveform can be both amplitude and frequency modulated by an external voltage. Frequency of operation can be selected externally over a range of 1 Hz to more than 1MHz. The functional block diagram for XR 2206 is shown in the fig.4.1

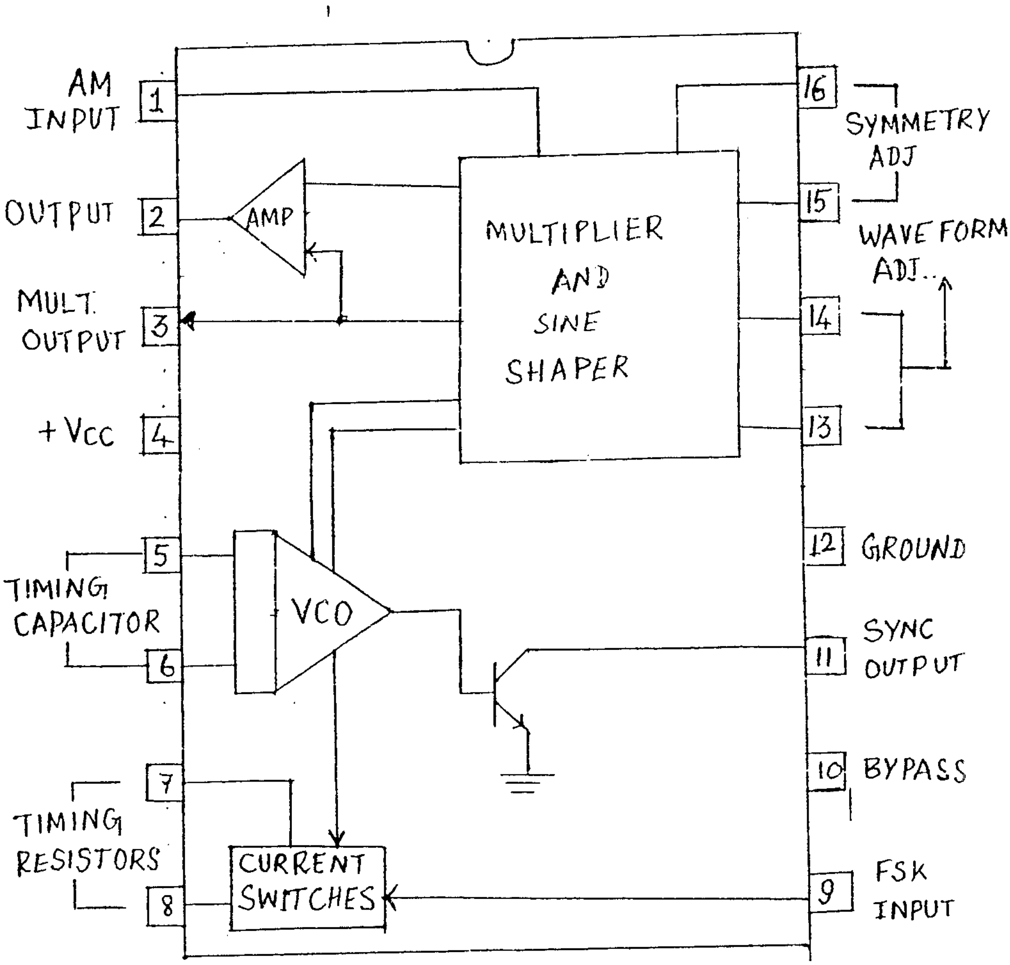


Fig. 4.1. FUNCTIONAL BLOCK DIAGRAM OF XR-

IC XR 2206 is comprised of four functional blocks

a) Voltage controlled oscillator

b) Analog multiplier and Sine-shaper

c) Unity gain buffer amplifier.

d) Set of current switches.

The VCO actually produces an output frequency proportional to an input current, which is produced by a resistor from the timing terminal to ground. The current switches route one of the timing pins current to the VCO controlled by an FM input pin to produce an output frequency. With two timing pins two discrete output frequencies can be independently produced for FM generation application.

Frequency modulation is a system in which amplitude of the modulated carrier is kept constant, while its frequency is varied by the modulating signal. The amount by which the carrier frequency is varied from the unmodulated value is called frequency deviation. This is made proportional to the instantaneous value of the modulating voltage. The amplitude of the frequency-modulated wave remains constant at all times. This is infact the greatest advantage of FM.

4.3. IC XR-2211

The demodulation is performed by using TONE DECODER IC (XR 2211). The IC XR 2211 is a monolithic phase locked loop system especially designed for data communication and operates over a wide frequency range of 1Hz to 300 KHz. It can accommodate analog signals between 2mV and 3V. The circuit consists of basic PLL for tracking an input signal frequency within the pass band, and a quadrature phase detector, which provides carrier detection. The functional block diagram of IC XR 2211 is shown in figure 4.2

The input signal is AC coupled to the input terminal. The internal impedance of the input pin is 20 kohms. The input signal is preamplified and fed to the loop phase detector and quadrature phase detector. Loop phase detector output provides a high impedance output for the detector. The PLL loop filter is formed by resistor R_f and capacitor Cf connected to output pin. With no input signal or with no phase error within the PLL, the DC level at output pin is very nearly equal to reference Voltage. The peak voltage swing available at the phase detector output is equal to reference Voltage.

External timing resistor R_o determines the VCO free - running frequency connected from VCO to ground. The VCO free running frequency is $f_o = 1 / R_oC_o$

Where Co is the timing capacitor.

Normally VCO terminal is at low impedance and is internally biased at a DC level equal to reference voltage. The maximum current drawn from timing resister must be limited to less than or equal to 3mA for proper operation of the circuit.

Reference voltage output pin must be bypassed to ground with a 0.1 microfarad capacitor. This pin is internally biased at the reference voltage reference voltage. The DC voltage at this pin forms an internal reference for the voltage level at low detector filter pin, FSK comparator input pin, loop and detector output pin and timing resister pin. The connection information is shown in Fig. 4.3

The VCO outputs are internally connected to the phase detector section of the circuit and form a PLL to track the instantaneous frequency of the input signal. If an FM input signal V_s of frequency FM is applied to the input of PL, the phase detector compares the phase and frequency of the incoming signal to that of the output V_o of the VCO .If the two signal differ in frequency and / or phase, an error voltage is generated. The phase detector is basically a multiplier and produces the sum (F_S+F_O) and difference (F_s-F_O) components at its output. The high frequency component (F_S+F_O) is removed by the loop filter and difference frequency component is applied as control voltage V_c to the VCO. The signal V_c shifts the VCO frequency in a direction to reduce the frequency difference between F_S and F_O. The VCO continues to change frequency till it is exactly the same as the input signal frequency.

Once locked, PLL tracks changes of the signal and demodulated output is taken from loop phase detector through a post detection filter R_f and C_f and an external buffer amplifier.

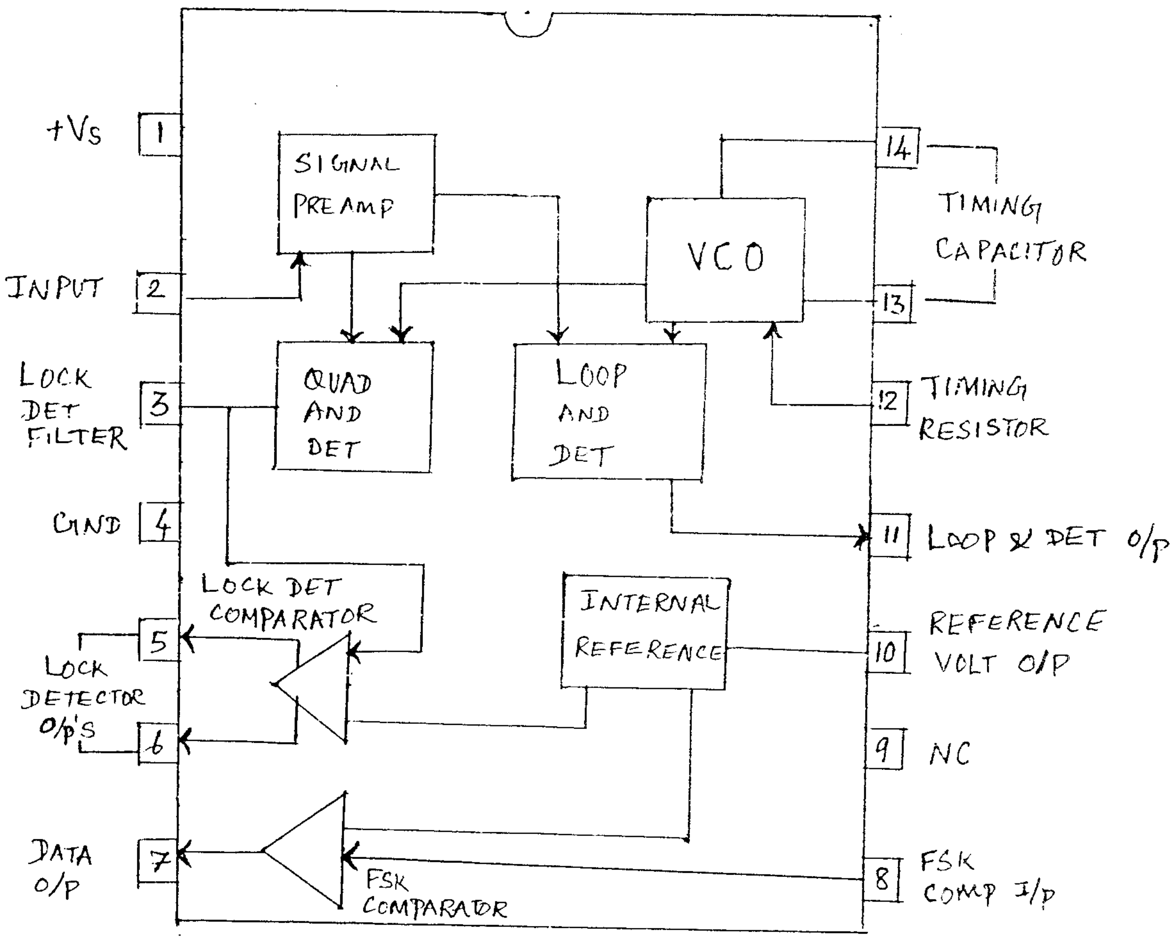


Fig. 4.2. CONNECTION INFORMATION OF IC XR-2211

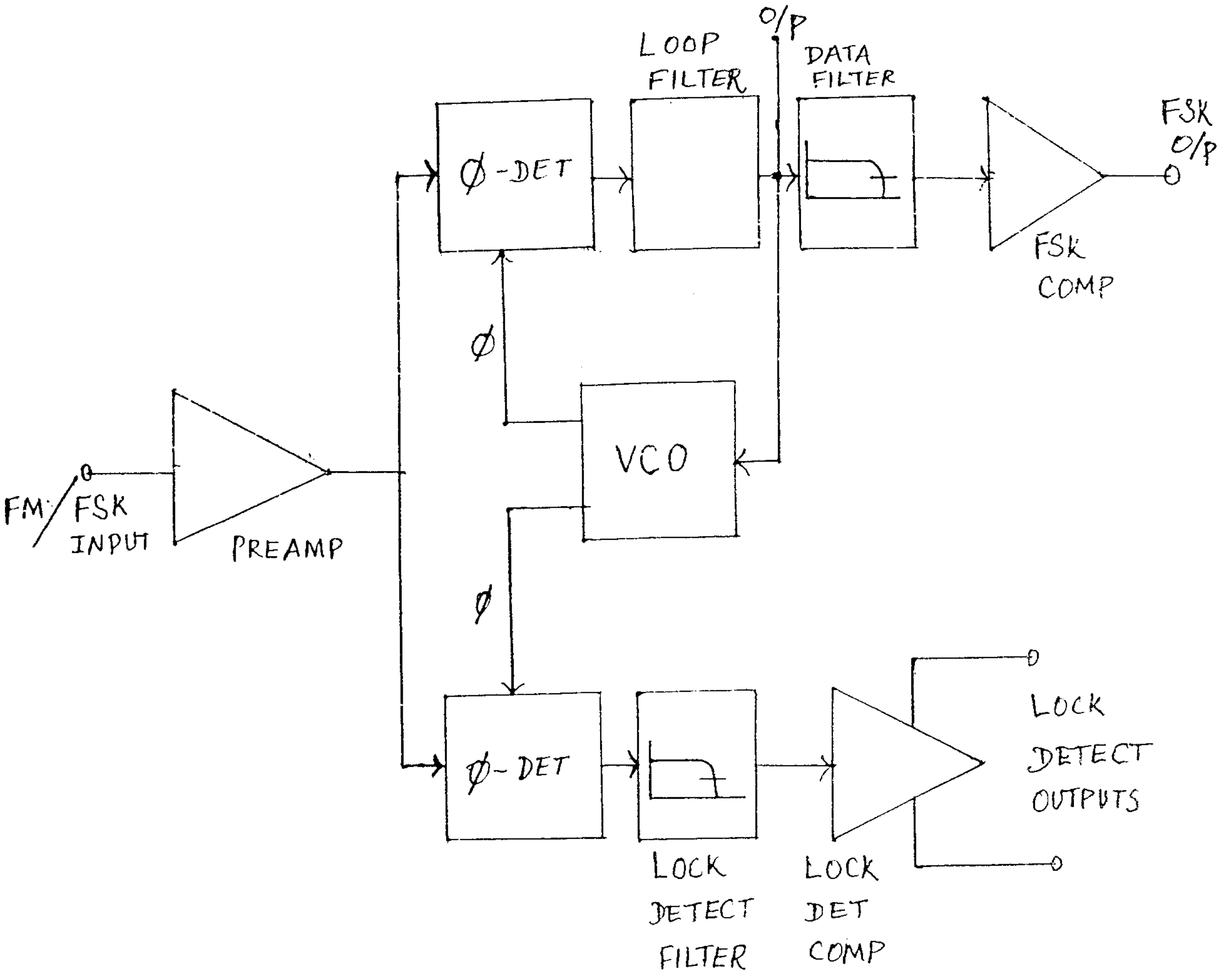


Fig. 4-3. FUNCTIONAL BLOCK DIAGRAM OF IC XR-2211

4.4. IC 741

Commercial IC operational amplifier usually consists of four cascade blocks as shown in figure 4.4.1. The first two stages are cascade difference amplifier used to provides high gain. The third stage is a buffer and the last stage is the output driver.

The buffer is usually an emitter follower whose input impedance is very high, so that it prevents loading of the high gain stage. The output stage is designed to provide low output impedance demanded by the ideal operational amplifier characteristics. The connection diagram of Op-Amp IC 741 is shown in Fig 4.4.

The output voltage should swing symmetrically with respect to ground. To allow such symmetrical swing the amplifier is provided with both positive and negative supply voltage. Power supply voltage of +15V and –15V are common.The internal circuit diagram of op_amp is shown in figure 4.5.

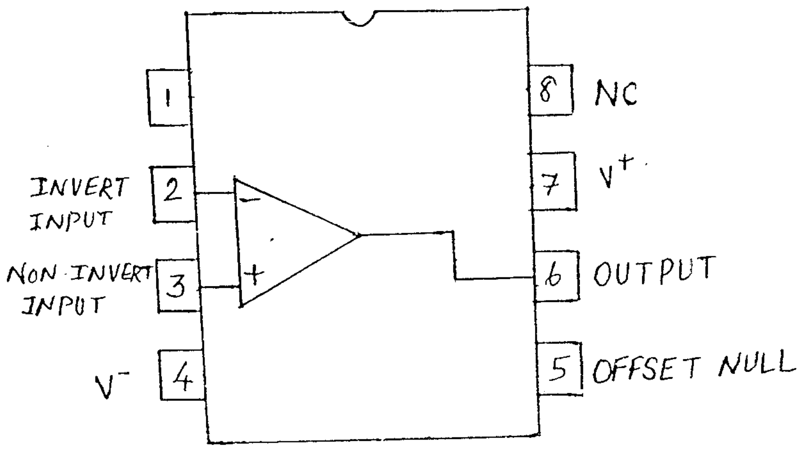


Fig.4.4. CONNECTION DIAGRAM OF IC 741

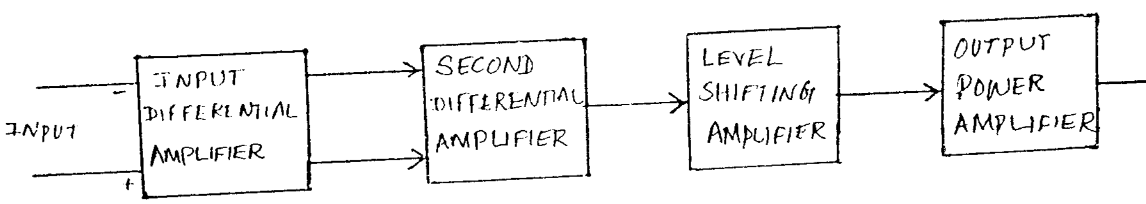


Fig.4.4.1. BLOCK DIAGRAM OF OP-AMP

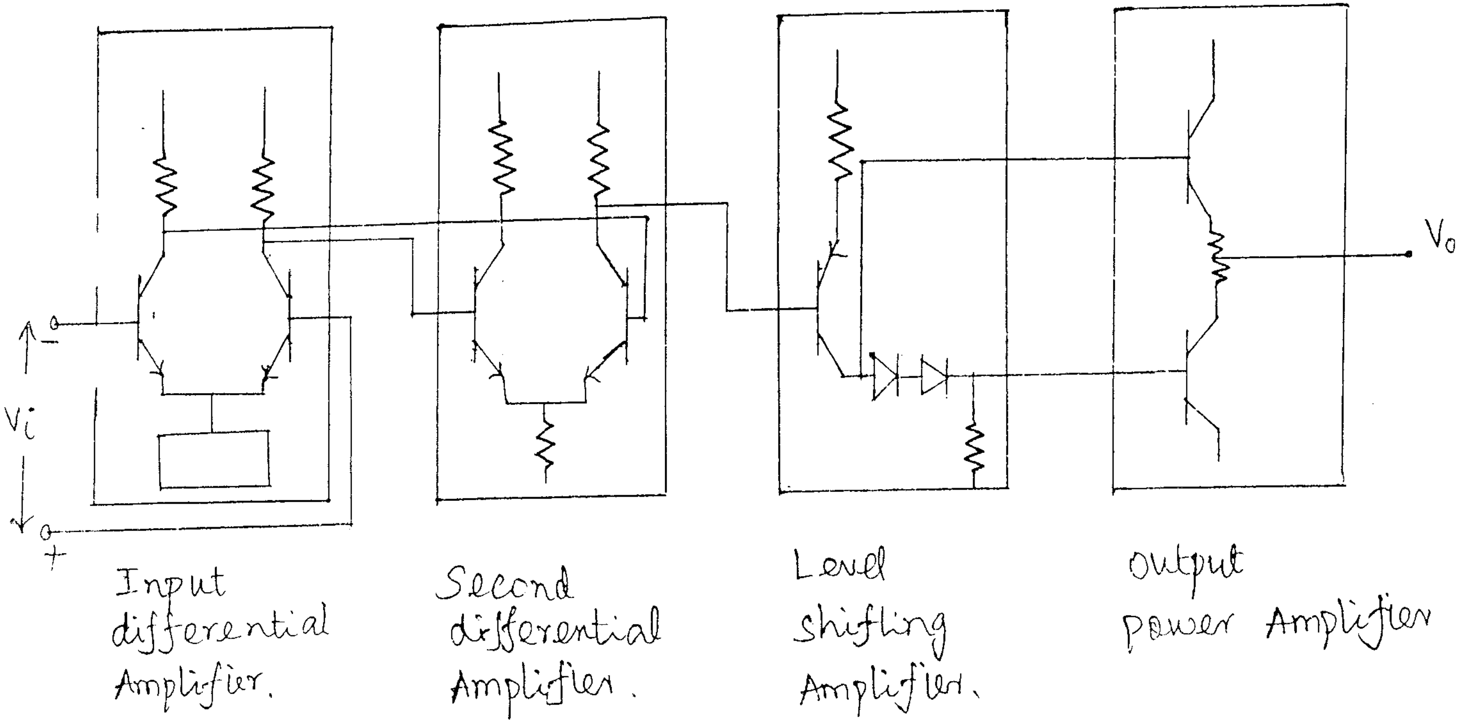


Fig. 4.5. INTERNAL CIRCUIT DIAGRAM OF IC 741.

4.4.1. INPUT DIFFERENTIAL AMPLIFIER

This stage determines the ultimate gain stability, common mode rejection, bias drift, input impedance, slewing rate, bandwidth and noise of the opamp. Subsequent stage have little effect on these parameters, i.e., if the first stage has a voltage gain of 10, the error in the second stage will only be 1/10 as noticeable as errors of equal size in the first stage. It is therefore modatory that the input differential amplifier be carefully designed and produced with repentable quality

4.4.2. SECOND DIFFERENTIAL AMPLIFIER

This stage is almost identical to the input differential amplifier except that a resistor takes the place of the current source; the CMRR requirement is not as great in the second stage. So a simple resistor current source is sufficient. Only one output is used on this second stage

4.4.3. LEVEL SHIFTING AMPLIFIER

The quiescent output voltage of operational amplifier should be zero if the input differential voltage is zero. A standard PNP common-emitter amplifier as shown in figure is often used. The load is made up of two diodes and a resistor and this provides a temperature compensation voltage divider for driving the output power amplifier and small resistor is placed in the emitter provides some negative output. This stage also transforms the high impedance of the second differential amplifier to low impedance capable of driving the output power amplifier.

4.4.4. OUTPUT POWER AMPLIFIER

This stage is usually an emitter follower of the complementary type; i.e the NPN transistor handles the positive output signal and the PNP transistor handles the negative output signal. An emitter follower provides high input impedance, high current gain, wide bandwidth and low output impedance. Since this stage must drive devices outside the opamp, it has substantial current driving capability. Current limiting beyond a fixed value is usually provided to protect the Op-Amp.

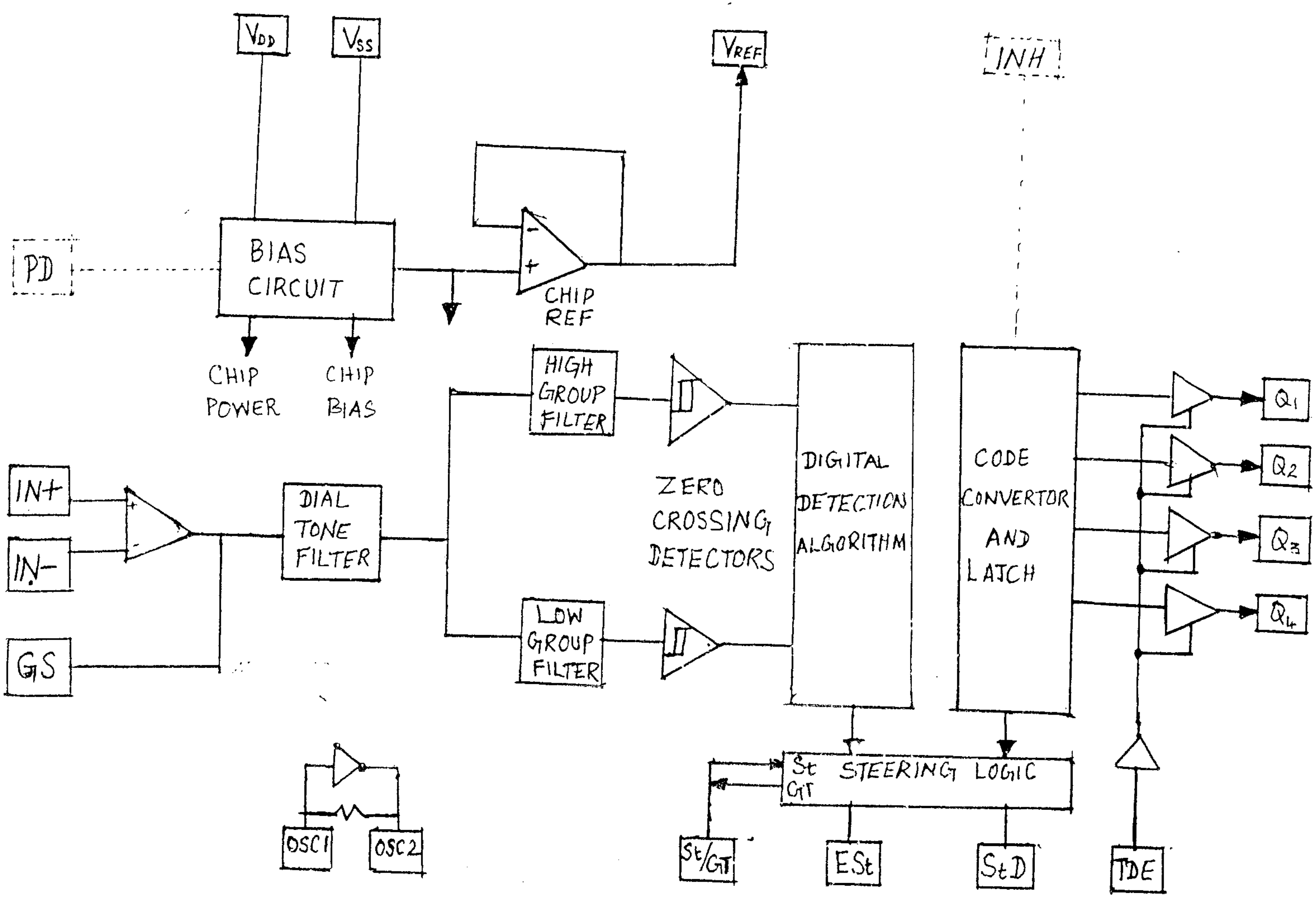
4.5. IC M-8870

4.5.1. INTRODUCTION

The DTMF RECEIVER uses M-8870 IC that integrates both bands split filter and decoder function into a single 18-pin DIP package. The M-8870 offers low power consumption (35 mW maximum) and precise data handling. Its filter section uses switched capacitor technology for both the high and low group filters and for dial tone rejection. Its decoder uses digital counting techniques to detect and decode all 16 DTMF tone pairs into a 4-bit [redacted] code.

M-8870 operating functions include a band split filter that separates the high and low tones of the received pair, and a digital decider that verifies both the frequency and duration of the received tones before passing the resulting 4-bit codes to the output bus.The Block Diagram of DTMF receiver is shown in Fig 4.6.

Fig. 4.6. BLOCK DIAGRAM OF A DTMF RECEIVER



4.5.2. FILTER

The low and high group tones are separated by applying the dual-tone signal to the inputs of the two 9^{th} order switched capacitor band pass filter with bandwidth that corresponds to the bands enclosing the low and high group tones. The filter also incorporates notches at 350 and 440 Hz providing excellent dial tone rejection. Each filter output is followed by a single switched capacitor section that smoothes the signal prior to limiting. High gain comparators provided with hystersis to prevent detection of unwanted low-level signals and noise perform signal limiting.

4.5.3. DECODER

The decoder uses a digital counting technique to determine the frequency of the limited tones and to verify that they correspond to standard DTMF frequencies. A complex averaging algorithm is used to protect against tone simulation by extraneous signal while tolerating small frequency variations. The algorithm ensures an optimum combination of immunity to talk off and tolerance to interfering signals and noise. When the detection recognizes the simultaneous presence of two valid tones (known as "signal condition"), it raises the early steering flag ($E_{st.}$). Any subsequent loss of signal conditions will cause E_{st}. to fall.

4.5.4. STEERING CIRCUIT

Before a decoded tone pair is registered, the receiver checks for a valid signal duration (referred as "character-recognition condition"). This check is performed by an external RC time constant driven by E_{st}. Logic high on E_{st}. causes V_c to raise as the capacitor discharges. Provided that signal condition is maintained (E_{st}. remains high) for the validation period (t_{GTF}), V_c reaches the threshold (V_{ts}) of the steering logic to register the tone pair, thus latching its corresponding 4-bit code into the output latch. At this point the GT output is activated and drives V_c to V_{dd}. GT continues to drive high as long as Est. remains high. Finally, after a short delay to allow the output latch to settle, the "delayed steering" output flag (std) goes high, signaling that the received tone pair has been registered. The contents of the output latch are made available on the 4-bit output by raising the three state control input (OE) to logic high. The steering circuit works in reverse to validate the interdigit pause between signals. Thus, as well as rejecting signals too short to be considered valid, the receiver will tolerate signal interruptions too short to be considered a valid pause.

4.5.5. GUARD TIME ADJUSTMENT

Where independent selection of receive and pause are not required. The simple steering circuit is applicable.

Component value is chosen according to the following formulae:

$$T_{rec} = T_{dp} + T_{gtp}$$

$$T_{gtp} = 0.67\ RC$$

The value of T_{rec} is the minimum signal duration to be recognizeszed by the reciver. The value of R for a Trec of 40ms would be 270-kilo ohm. Different steering arrangements are used to select independently the guard times for tone present (T_{gtrp}) and tone absent (T_{gta}). This may be necessary to meet system specifications that place both accept and reject limits on both tone durations and interdigit pause.

Guard time adjustment also allows the designer to tailor system parameters such as talk off and noise immunity. Increasing T_{rec} improves talk off performances, since it reduces the probability that tones simulated by speech will maintain signal condition long enough to be registered. On the other hand, a relatively short T_{rec} with a long T_{do} would be appropriate for extremely noisy environments where fast acquisition time and immunity to dropouts would be required.

4.5.6. INPUT CONFIGURATION

The input arrangement of the M-8870 provides a differential input op-amp as well as a bias source (V_{ref}) to bias the inputs at mid rail. Provision is made for connection of a feedback resistor to the op-amp output (GS) for gain adjustment.

4.5.7. DTMF CLOCK CIRCUIT

The internal clock circuit is completed with the addition of a standard 3.57945 MHz television color burst crystal.

5. CHAPTER

HARDWARE DESCRIPTION

5.1. INTRODUCTION

In many organizations, the exchange of information between the departments is very difficult if the intercom facility is not there. Normally two wires are run from the terminal to switching center. In this project those two wires are replaced by power lines, which are already present everywhere.

5.2. BLOCK DIAGRAM

The system block diagram is shown in figure 5.2.

5.2.1. TELEPHONE

This device is used to generate the necessary control signals which are important in order to select the particular terminal, and to send the speech signals. The internal operation of the telephone has been discussed in chapter 3.

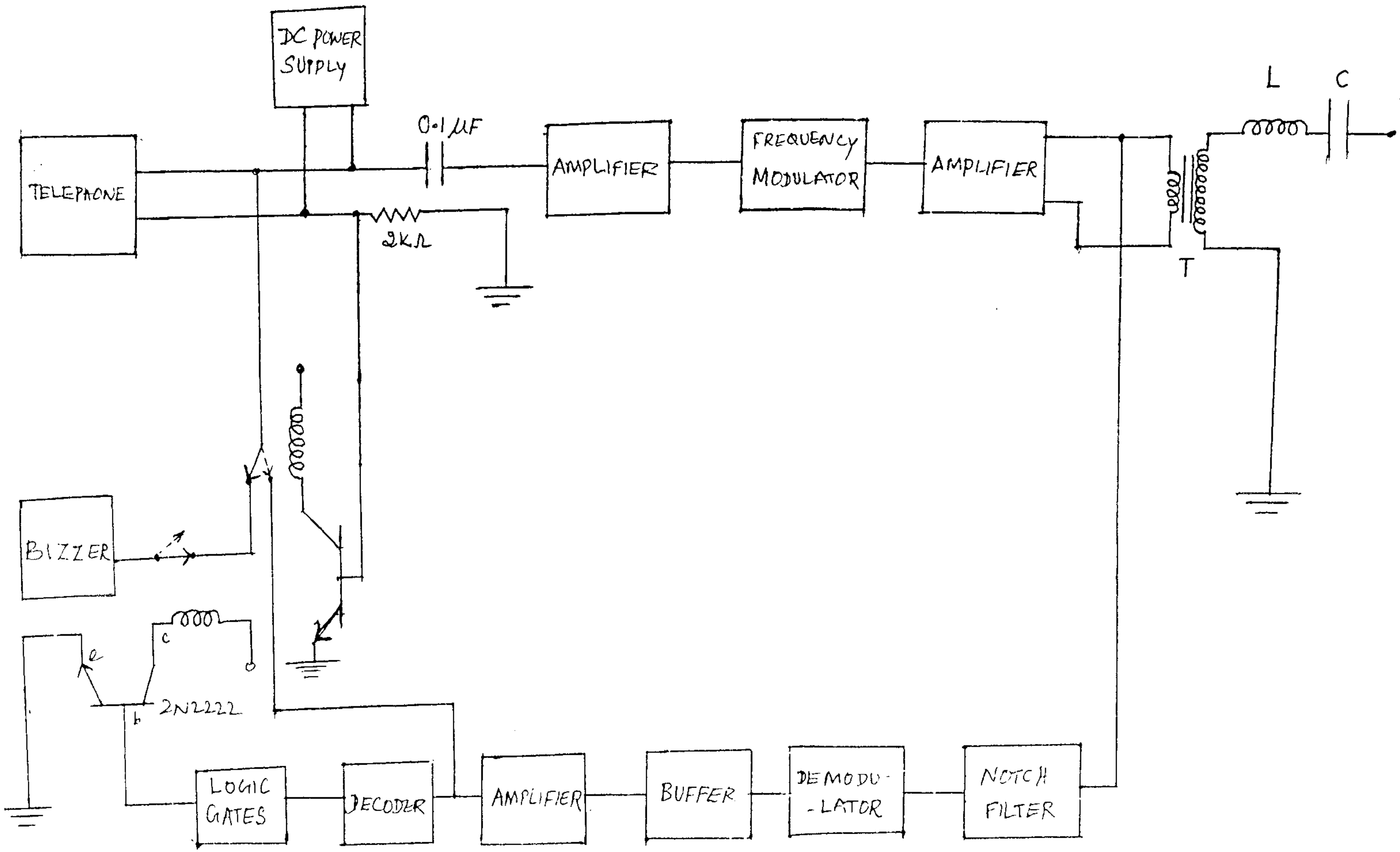


Fig.5.2. BLOCK DIAGRAM

5.2.2. AMPLIFIER

Amplifier is a device which is used to boost the voltage or power level of the input signal at the output. The opamp is used in the system to amplify the speech signal from the telephone as well as the control signals. The input impedance of the opamp is very high and the output impedance is very low.

Operational amplifier is used to amplify the signal. The non-inverting ac amplifier is used. The ac signal is superimposed with dc level. It should be essential to block the dc component. This is achieved by using an ac amplifier with a coupling capacitor. The ac noninverting amplifier circuit is shown in figure 5.2.2

The gain of the circuit is given by

$$Av = 1 + R_f / R_i$$

Where

R_f – feedback resistor
R_i – input resistor

The resistor R_3 is added to provide a dc returns to ground. However this reduces the overall input impedance of the amplifier, which now becomes approximately R_3

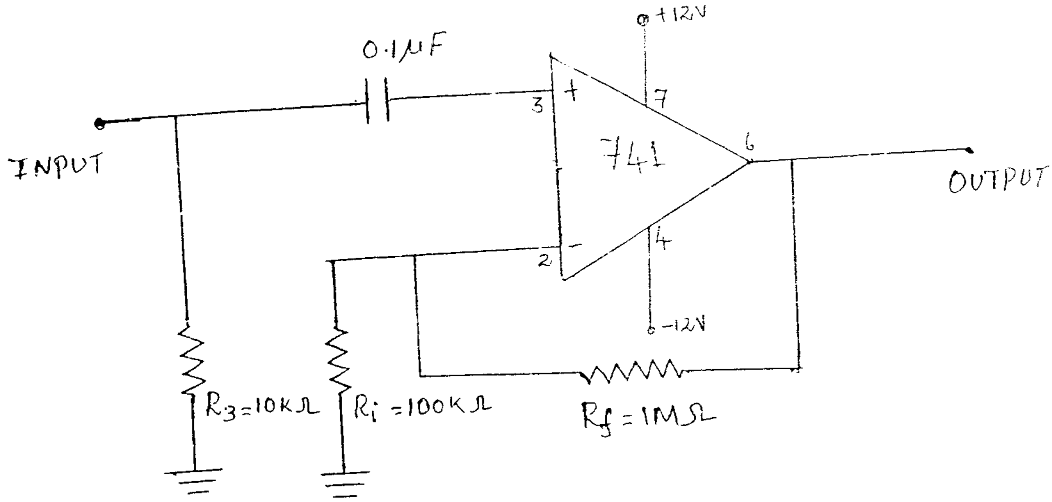


Fig.5.2.2. AMPLIFIER

5.2.3. DESIGN EQUATION

For non inverting amplifier

The voltage gain is given by $Av = (1+R_f / R_i)$

Where

R_f –feedback resistor
R_i –Input resistor

R_i should be kept fairly large to avoid loading effect

5.2.4. CALCULATION

Input Signal Range = 20 –50 mV

Gain fixed is $A_v = 10$

Assume

$R_f = 1M$ ohm
$R_I = 100$ K ohm

5.3. MODULATOR

The modulator deviates the frequency of the carrier signal according to the instantaneous voltage of the modulating signal; such a modulation technique is called as frequency modulation. The amplitude of the input signal should be in the range of 2 to 7 volts in order to get perfect modulated wave for the desired frequency.

5.3.1. FREQUENCY MODULATOR

XR 2206 IC does the modulation. The XR 2206 is a monolithic function generator integrated circuit capable of producing high quality sine, square, triangle waveform of high stability and accuracy. An external voltage can modulate the output waveform. Frequency of operation is selected externally through the RC components, which is 21 KHz. The circuit diagram is shown in figure 5.2.3.

The frequency of oscillation, f_0 is determined by the external timing Capacitor C, and by the timing resistor R. The frequency is given as $f_0 = 1 / RC$ and can be adjusted by varying either R or C. Temperature stability is optimum for resistor range up to 200 kohm, recommended values of C are from 1000pF to 0.1mF

Adjusting resistor minimizes the carrier signal harmonic distortion. Connecting resistor at sine shaper pin does the fine adjustment Maximum amplitude at the output is inversely proportional to the external resistor, R_3. FM sine wave amplitude is approximately 60mV

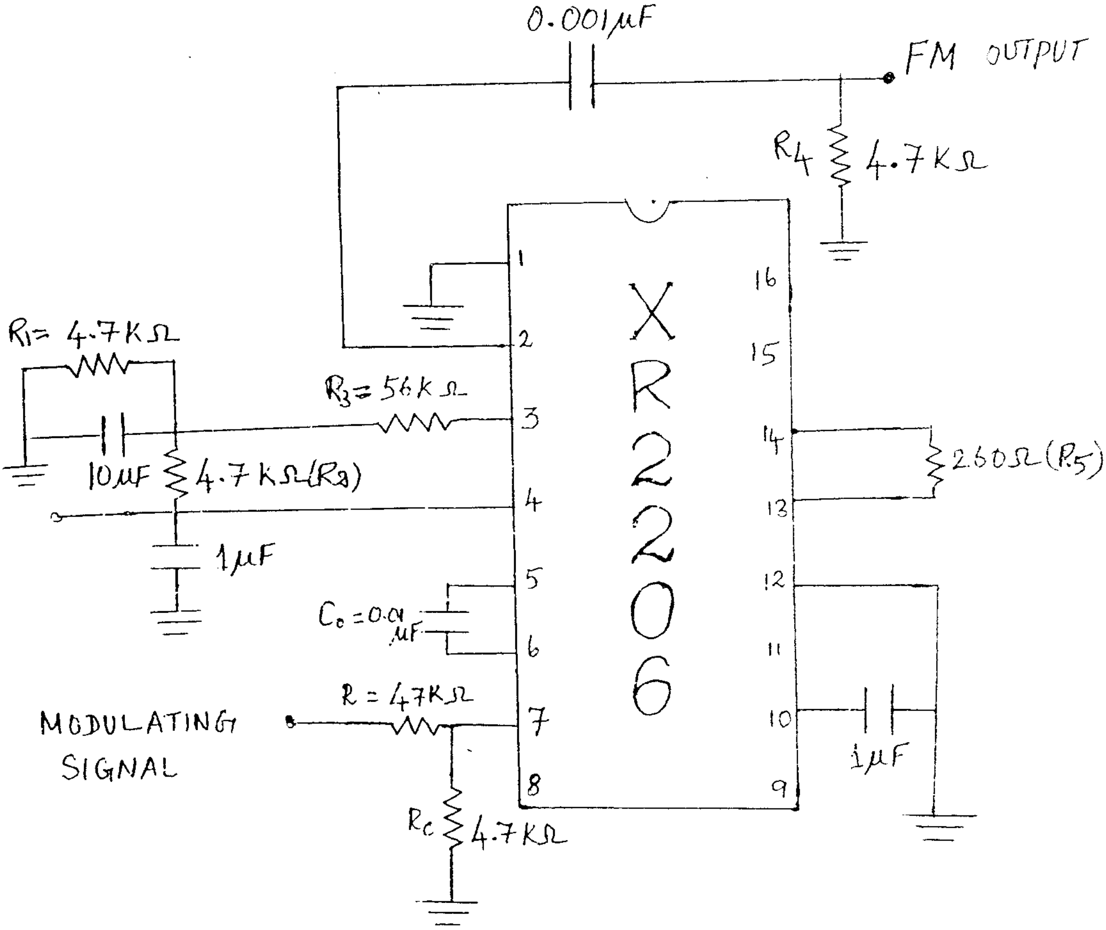


Fig.5.2.3. FREQUENCY MODULATOR

peak to peak per kohm R3.Thus for R_3=50 k would be producing approximately 3Vsinusoidal output amplitude.

5.3.2. DESIGN EQUATION

Frequency operation of circuit is $f_0 = 320\ It / C_o$

Where

I_t –current flowing through the timing terminal in mA

Timing terminal is internally biased at 3.125V and maximum allowable timing current is 3 mA.

Timing current $I_t = I_b + I_c$

Where

I_b - current flow through resistor timing resistor R

I_c - current flow through voltage divider resistor R_c

$$I_b = 3.125 / R$$

$$I_c = (3.125 - V_c) / R_c$$

Where

V_c - control voltage

The frequency of oscillation related to Vc is

$$f_o = 1 / RC_o (1 + R / R_c (1 - V_c / 3.12))$$

By differentiate above equation we get voltage to frequency conversion gain

i.e. $K = df_c / dV_c = -0.32/R_cC_o$ Hz

5.3.3. CALCULATION

The carrier frequency fixed for modulation is 21 KHz

To obtain 21 KHz, Vc should be zero.

$$f_o = 1 / R_p C_o$$

$$R_p = R \parallel R_c$$

Assume

$$C_o = 0.01 \text{ uF}$$

$$R_p = 1 / f_o C_o$$

$$= 4.8 \text{ kohm}$$

Assume

$$R_c = 4.7 \text{ kohm}$$

$$R_p = R \parallel R_c$$

Therefore

$$R = 47 \text{ K ohm}$$

$$I_b = 3.125 / R$$

$$= 0.66 \text{ mA}$$

$$I_c = 3.125 / R_c$$

$$= 0.066 \text{mA}$$

Therefore timing current

$$I_t = I_b + I_c$$

$$= 0.72 \text{ mA}$$

5.4. DEMODULATOR

The demodulator is used to reproduce the original signal which was frequency modulated. This can detect the signals even at less amplitude (from 30 mv) without any noise.

5.4.1. FREQUENCY DEMODULATOR

Using Tone decoder IC XR 2211 does the frequency demodulation. The XR 2211 is monolithic PLL system especially designed for data communication and operates over a frequency range of 1 Hz to 300 KHz. The circuit consists of a basic PLL, which is locked to the input FM signal, and the VCO tracks the instantaneous frequency of the input signal. The filtered error voltage, which controls the VCO and maintains lock with input signal, is demodulated output

FM detection circuit is shown in Figure 5.2.4. The carrier frequency is fixed by external timing resistor R_o and timing capacitor C_o. The input FM signal is applied to input pin through AC coupled capacitor. Normally internal impedance at input pin is 20 kohm. Then the input signal is internally pre amplified and the internal PLL lock to a FM signal and the VCO tracks the instantaneous frequency of input

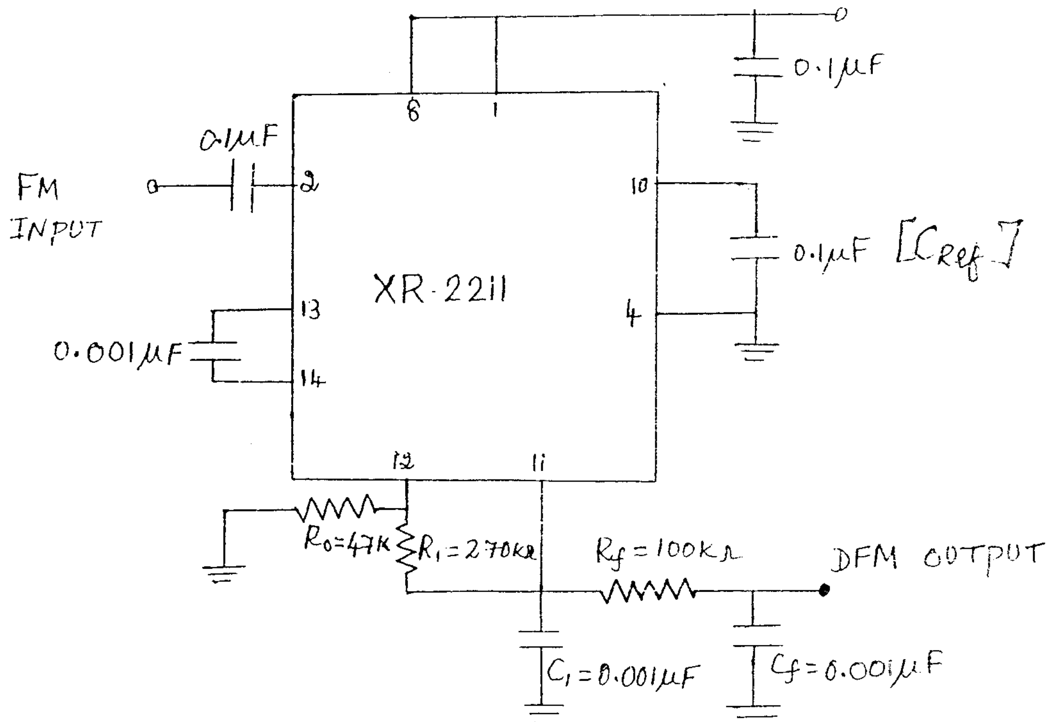


Fig.5.2.4. FREQUENCY DEMODULATOR

signal. R_1 and C_1 form the PLL loop filter. It filters the signal phase error with in the PLL. When no input signal is at input, FM output is equal to the reference voltage V_r.

The demodulated output is taken from the loop phase detector through a post detection filter made up of R_f and C_f. It is a normally high impedance output at loop phase detector, so buffer is used for impedance matching. Internal voltage is set by connecting the capacitor C_{ref}

5.4.2. DESIGN EQUATION

1.VCO center frequency $f_o = 1/ R_oC_o$ Hz

2.Internal reference voltage V_r

$V_r= (Vs/2)-650Mv$

3.Loop low pass filter time constant T

$T = R_1C_1$

4.loop damping S

$S = (SQRT (C_o / C_1)) / 4$

5.Loop tracking bandwidth,

$D_F/f_o= R_0/R_1$

6.Filter time constant T_f

$T_f = R_F C_F$

7.Loop phase detector conversion gain

$K_f = -2V_r / 3.14$ (v/radians)

8.VCO conversion gain

$K_o = -1 / R_1C_oV_r$ (Hz/v)

9.Total loop gain

$K_t = 4 / C_oR_1$ (radians/s/v)

10.Peak phase detector current

$I_a = V_r / 25$ (mA)

5.4.3. CALCULATION

1. Carrier frequency f_o =21 KHz

$$f_o = 1 / R_oC_o$$

Assume

$$C_o = 0.001uF$$

$$R_o = 47 \text{ kohm.}$$

2. Loop tracking bandwidth D_f =3655 HZ

$$R_1 = R_oF_o / D_f$$

$$= 270 \text{ k ohm}$$

3. Normally loop damping S = ½ is optimum for most FM detector application

$$C_1 = C_o / 16S$$

$$C_1 = 0.0002uF$$

4. Filter time constant T_f

$$T_f = R_f C_f$$

Assume

Filter constant =0.1mS

R_f =100 Kohm

C_f =0.001uF

5. Internal reference voltage

V_r = Vs/2 -650mV

V_r = 5.35 V

6. FM detector gain

i.e. the output voltage change per unit of f_e deviation can be give

$V_{out} = R_1 V_r / 100 R_o$ (V / % deviation)

= 3.07

7. Peak phase detector current I_a

$I_a = V_r / 25$

= 215 mA

5.5. DTMF RECEIVER

This receiver receives the DTMF signal generated by the telephone and a corresponding 4-bit [redacted] code is generated at the output. This process is done by the filter and by the preprogrammed algorithm. Filter separates low and high frequencies and then the frequencies are compared with the standard values, then the output is obtained.

5.5.1. DTMF DECODER

This circuit is shown in fig.5.2.5, the DTMF receiver is used to convert the dual tone multi frequency signal to 4-bit output. The input is given between inverting and non-inverting inputs. The output 4 bits are obtained at output pins.

The crystal oscillator is used to produce clock frequency for the IC operation; frequency of the crystal used is 3.57945 MHz. The resistors R_1 and R_4 fix the voltage gain. The resistor R_5 is feedback resistor to the op-amp and used for gain adjustment. The resistor R and capacitor C connected at E_{st} and st / gt pin fix the minimum signal duration to be recognized by the receiver. A suitable value of R for a T_{rec} of 40 ms would be 300 kohm, this resistor selects the guard time for tone detection and it improves the talk off performance and noise immunity.

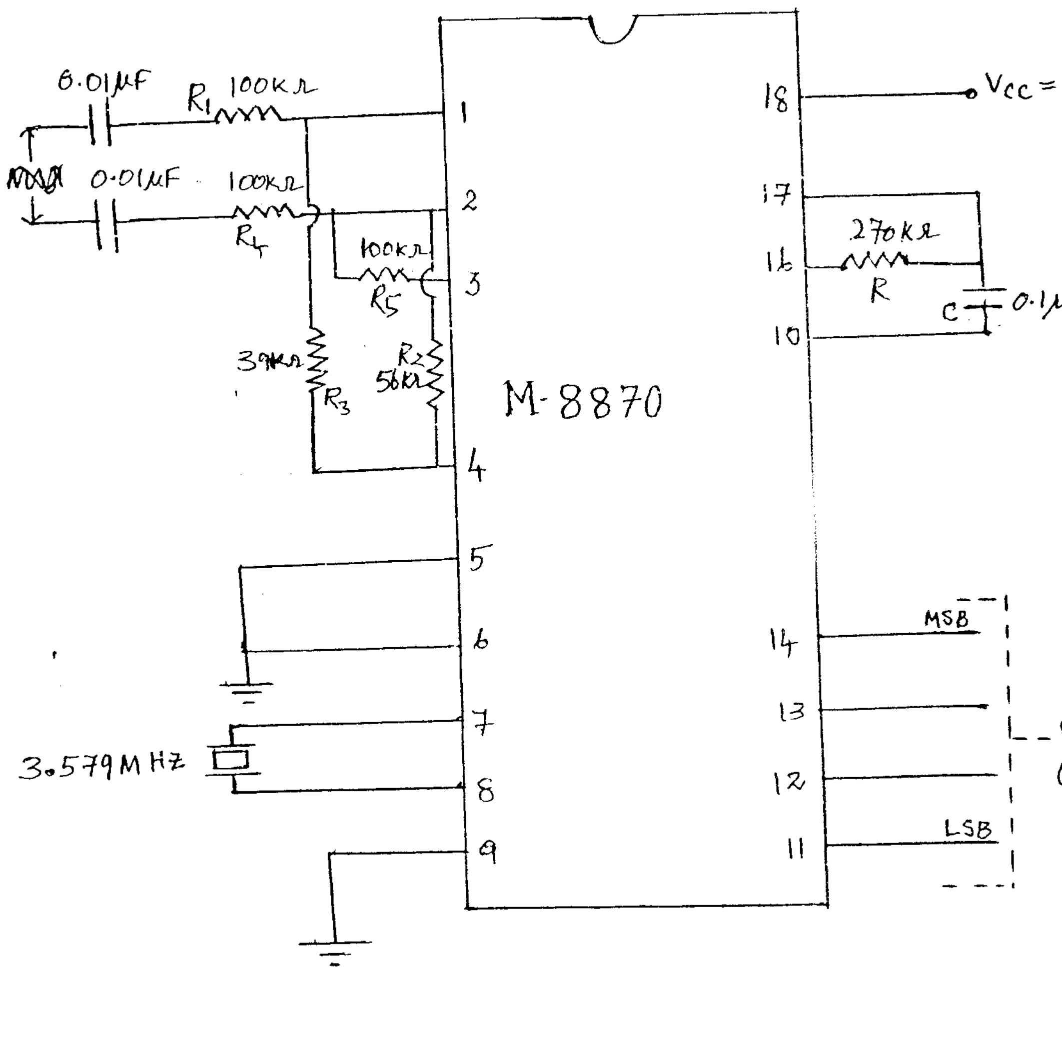


Fig. 5.2.5. DTMF RECEIVER

5.5.2. DESIGN EQUATION

GUARD TIME ADJUSTMENT

The minimum signal duration to be recognized by the receiver is

$$T_{rec} = T_{tdp} + T_{gtp}$$

Where

T_{dp} -tone present detection time

T_{gtp} – guard time for tone present

$$T_{gtp} = RC \ln [V_{dd} / (V_{dd} - Vt_{st})]$$

Where

V_{dd} – positive power supply for circuit

V_{tst} – threshold voltage of steering logic to register tone pair

For guard time for tone absent is given by

$$T_{gta} = RC \ln (V_{dd} / V_{tst})$$

DIFFERENTIAL INPUT CONFIGURATION

Voltage gain of differential input amplifier (A_t) $=R_4 / R_1$

Where

R_4 - resistor connected at non inverting terminal

R_1 - resistor connected at inverting terminal

$$R_3 = R_2 \,||\, R_5$$

Where

R_2 - resistor connected at non inverting input pin from reference voltage output pin

R_5 - gain select resistor

5.5.3. CALCULATION

GUARD TIME ADJUSTMENT

$$T_{gtp} = RC \;\ln V_{dd}/(V_{dd} - V_{tst})$$

Where

$$V_{dd} = 5V$$
$$V_{tst} = 2.442$$
$$T_{gtp} = 0.67RC$$

Tone present detection time T_{tdp} =20mS

Minimum tone duration accept T_{rec}=40 mS

$$T_{rec} = T_{gtp} + T_{dp}$$

$$0.67RC = 20 \text{ mS}$$

Assume

$$C = 0.1uF$$
$$R = 300 \text{ kohm}$$

DIFFERENTIAL INPUT CONFIGURATION

Assume

C_1=10 nF

C_2 =10nF

R_1 =100 kohm

R_2=56 kohm

R_4 =100 kohm

R_5=100 kohm

$R_3=R_2 \parallel R_5$

R_3=39 kohm

Voltage gain

$A_v = R_4/R_1$

$=1$

5.6. COUPLING EQUIPMENT

The equipments used to connect the system with the power line are coupling capacitor, transformer and filter.

5.6.1. TRANSFORMER

A transformer is a device with two or more stationary electrical circuits that are conductively disjoined but magnetically coupled by a common time varying magnetic field.

Transformers are basically passive devices for transforming voltage and current. One of the windings, generally termed as secondary winding, transforms energy through the principle of mutual induction and delivers power to the load. The voltage levels at the primary and secondary windings are usually different and any increase or decrease of the secondary voltage is accompanied by corresponding decrease or increase in current.

Simply transformer is a static piece of apparatus used for transforming power from one circuit to another without change in frequency. The physical basis of transformer is mutual inductance between two circuit linked common magnetic flux through a path of low reluctance.

If the number of turns in primary N_1 and in secondary N_2 and the voltage in primary is E_1 and in secondary is E_2 for an ideal transformer

$$E_2 / E_1 = N_2 / N_1 = k$$

Where

k is the voltage transformation ratio

If $N_2 > N_1$ i.e., $k > 1$, then the transformer is called as step up transformer.

If $N_2 < N_1$ i.e., $k < 1$, then the transformer is called as step down transform

Input Power = Output Power

$$V_1 * I_1 = V_2 * I_2$$

$$V_1 / V_2 = I_2 / I_1 = 1 / k = N_1 / N_2$$

Where I_1, I_2 are primary and secondary currents.

5.6.2. COUPLING CAPACITOR

The most important technical problem in a power line carrier is to devise methods and equipment to couple the low voltage and high frequency carrier to the high voltage and low frequency power line.

During the initial stages of carrier operation on power lines, an aerial wire of more than 300 feet length was used on one end and other end was connected to the power line. This aerial couples the signals to power line mainly due to capacitance between them.

SOME OF THE DRAWBACKS OF THIS AERIAL COUPLING ARE

1. It was inefficient and created disturbances in near by broadcast receivers
2. It was affected by radio transmitted signals
3. Means had to be provided against high voltages appearing on it due to accidental contact with high voltage lines.

The modern practice is to achieve the coupling by connecting a capacitor between the carrier terminal and the high voltage lines. The particular value of the capacitor is a compromise between two conflicting requirements. Too low values will present large reactance to carrier currents, while too high a value will be uneconomical to build and would pass large currents at 50 Hz.The typical range is between 2200 to 4400 pF.

5.6.3. FILTER

This filter is also called line-tuning unit and offers the necessary means of feeding the modulator output from the carrier set to the coupling capacitor. The typical coupling filter is as shown in fig. 5.2.6.3(a).

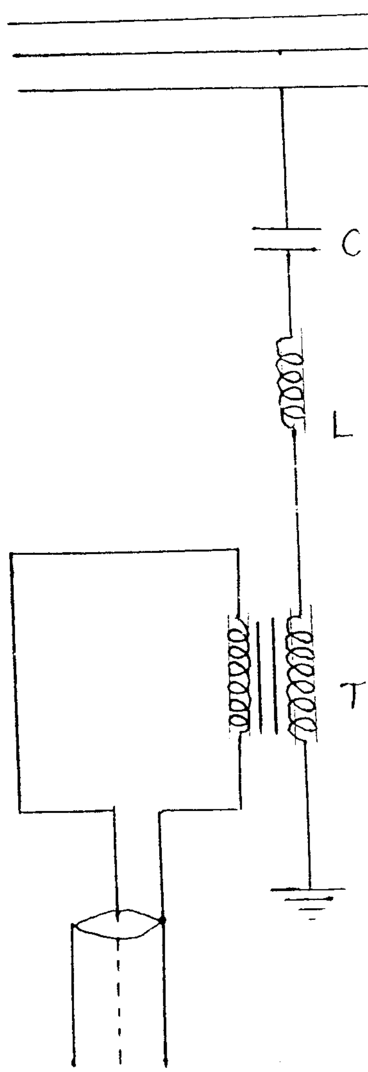


Fig. 5.2.6.3(a). COUPLING FILTER [Single Channel

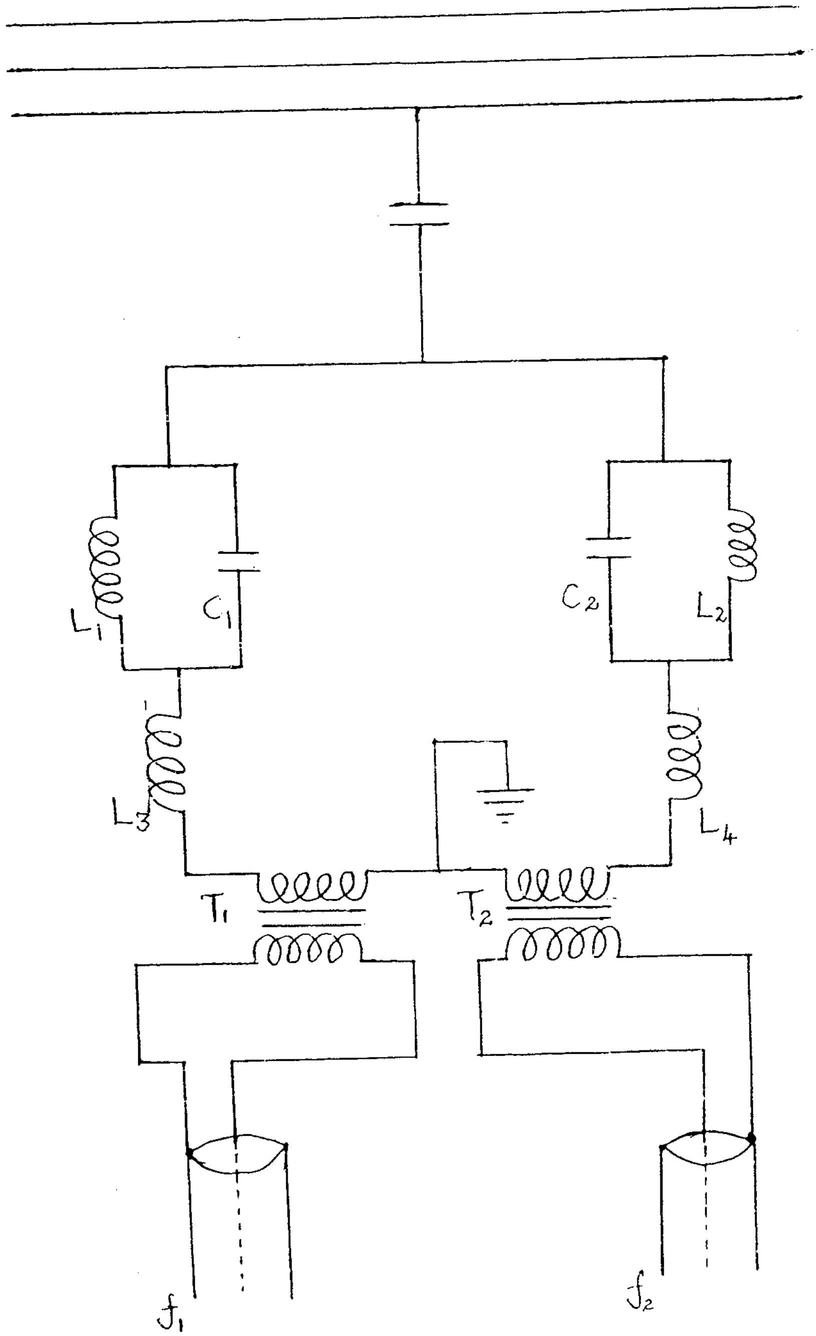


Fig.5.2.6.3(b). COUPLING FILTER [Double Chann

The figure shows its elements when only one carrier channel is fed to the power line. The variable inductor L in series with the coupling capacitor C forms resonant circuits so as to pass the carrier frequency along with the associated sidebands. The transformer T act as a separating as well as matching transformer between the impedances of the lead in the coaxial cable and the coupling filter itself. When the two carrier channels are fed to the power line then the design of filter becomes as shown in Fig.5.2.6.3 (b)

In figure 5.2.6.3 (b) the parallel combination of L_1 and C_1 is tuned to frequency f_2, so that it offers very high impedances to the f_2 channel. The parallel combinations L_1,C_1 in series with L_3 are now tuned to f_1 so as to offer negligible impedance to f_1 and its side bands. Similarly the combination L_2 and C_2 is tuned to f_1 and the L_2, C_2 and L_4. Series circuit is tuned to f_2. Then both the f_1 and f_2 channels find separate low impedance path to the power line and at the same time is prevented from being shunted by the parallel path. T_1 and T_2 act as separating and matching transformers for the two channels.

5.6.4. COUPLING TECHNIQUES

PHASE TO GROUND COUPLING

The connections made are shown in Fig 5.3.1. One of the terminals of the secondary of the transformer is connected to the phase line while the other terminal is grounded. This is the most widely used technique.

PHASE-TO-PHASE COUPLING

The connection are made as shown in fig 5.3.2. It provides metallic go and return paths to the carrier currents. In this case the uncoupled conductor does not have any appreciable influence on the transmission.

TWO PHASE COUPLING

This is shown in Fig 5.3.3. It provides reliability of operation in case coupled conductor develops some fault. It needs twice the coupling equipments. For this reason, this method is not used in practice.

INTERPHASE COUPLING

The connection are made as shown in fig 5.3.3. When two power lines run on the same poles, one terminal of the carrier set is coupled to one phase conductor of one of the lines and the other to similar phase conductor of the other line. Although this method gives a more reliable carrier communications its application is limited, as two sets of lines on the same towers may not always be available.

It is apparent from the above discussions that the only practicable methods of coupling are phase to ground and phase-to-phase methods, and both these methods are widely used.

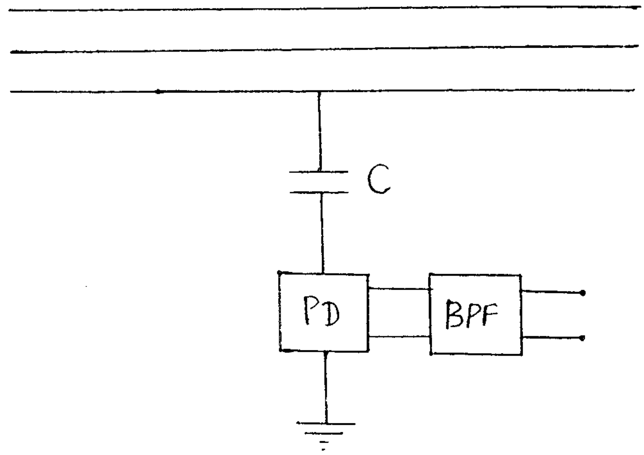


Fig.5.3.1. PHASE TO GROUND COUPLING

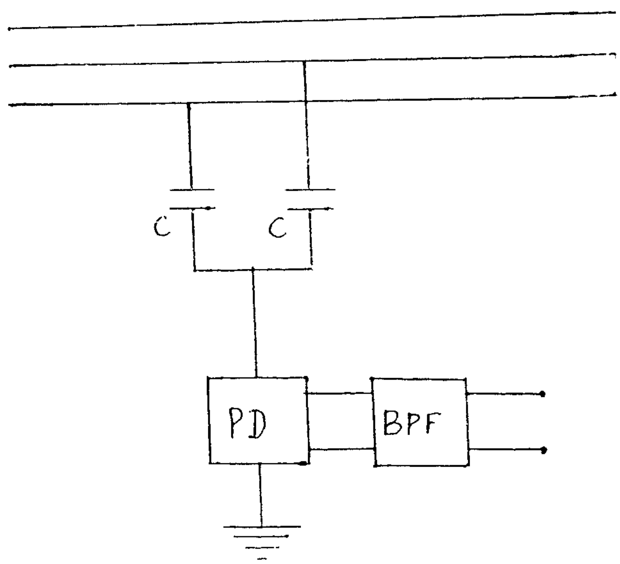


Fig.5.3.2. TWO PHASE COUPLING

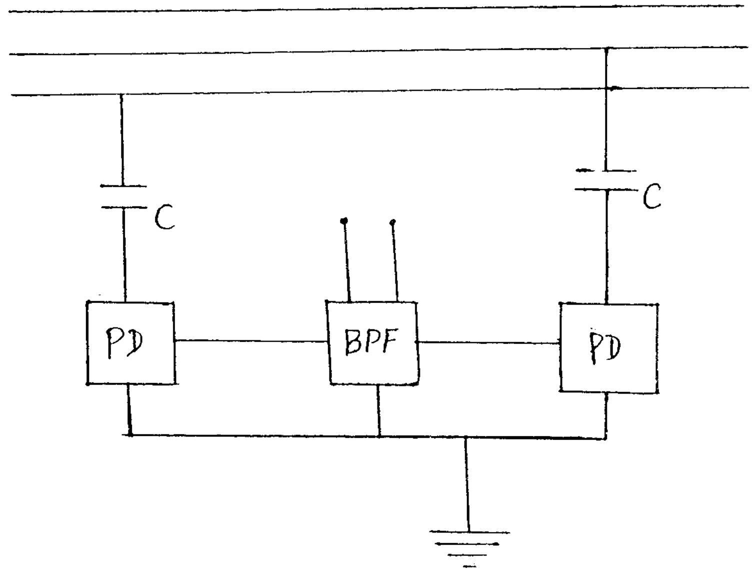


Fig. 5.3.3. PHASE TO PHASE COUPLING

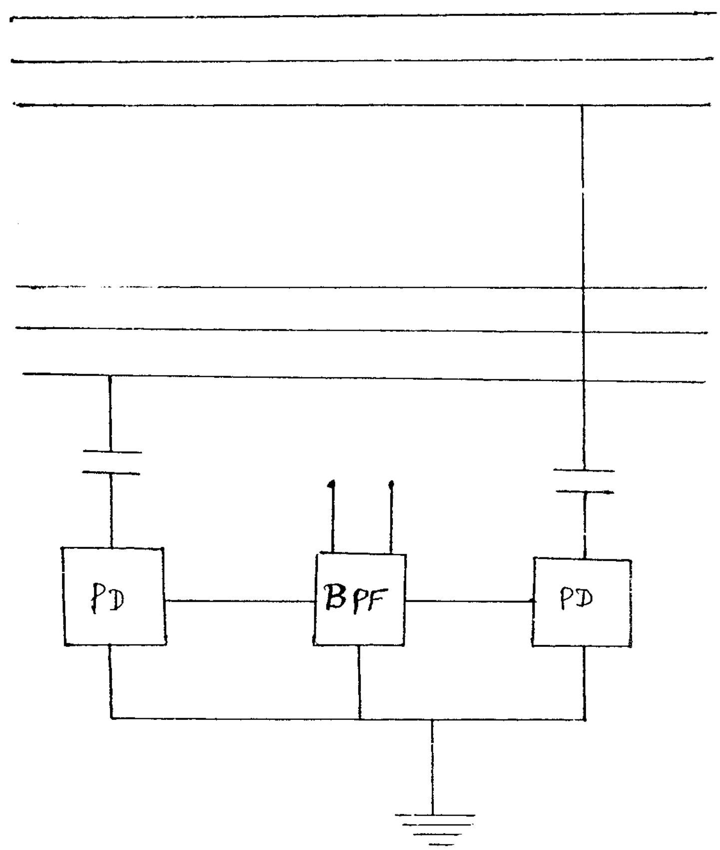


Fig. 5.3.4. INTER SYSTEM COUPLING

5.7. WORKING PRINCIPLE

If one wants to place a call to somebody through the intercom, they need to the indicate that to other end and this operation is accomplished by the control signal known as DTMF signal.

When the number is pressed, the corresponding DTMF signal for the button is generated by the telephone. This signal is amplified since the amplitude of the DTMF signal is in the range of millivolts. The modulator requires input signal level from 2 volts to 7 volts, the input signal is frequency modulated, again the signal is amplified in order to get efficient communication then the signal is coupled with the power line (230 volt, 50 Hz AC) signal by the coupling equipment. Coupling capacitor provides coupling operation. The transformer accomplishes the isolation and a filter is used to reject the 50 Hz signals, which has the chance of getting into the system and damaging it.

The DTMF signal is travel through the power line, then it is received at the receiving end by a coupling capacitor, transformer and a filter to reject the 50 Hz signal. The received signal is fed to the demodulator, which produces the original signal that was transmitted. The DTMF signal is passed to the decoder via a buffer. Because the output impedance of the demodulated IC is very high and therefore it is necessary to put the buffer for impedance matching between DTMF decoder and demodulator.

The DTMF receiver is an IC, which has the capability of identifying the frequency. When the DTMF signal is fed to the receiver it produces a 4-bit BCD code. According to the number, which was pressed at the transmitting or calling side, the logic gate is used to check the called terminal. If the terminal is not selected the output of this gate will be zero. When the terminal is selected this gate produces a one at the output. For example, if 7 were pressed, the 4-bit code obtained will be 0111 then the gates used are,

NOT gate to invert the MSB

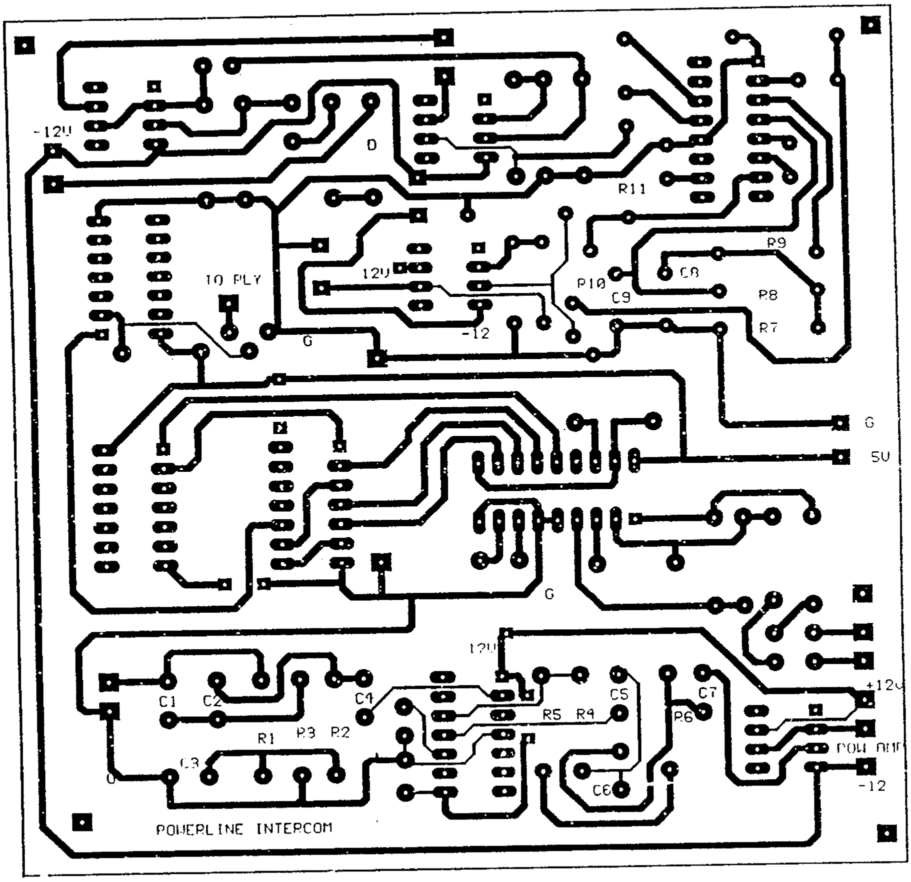
2-Input AND gate to multiply the input

If the output of decoder is 0111, then the gate will produce output of 5 volts, otherwise zero volts. This output voltage is used to switch on the buzzer at the called terminal. When the base terminal of the switching transistor is given 5 volts it will go into saturation and act as a short circuit. Hence the relay gets energized and the buzzer is turned on.

Normally the telephone will be connected in path 1 i.e. through the decoder - logic gate - relay 1. When the DTMF signal is received, transistor 1 will be on and relay 1 will be energized then the buzzer will ring. Instead of DTMF signal if the voice signal is fed to the decoder it will make the all the outputs high and hence the output of the logic gates would be zero. Then the transistor 1 will be off, hence relay 1 will be off therefore the buzzer will be off.

When the buzzer rings it will attract the attention of the called party indicating that somebody wants to contact him and he will lift the handset. As soon as the hand set is lifted, the resistor which is connected to the ground will have some voltage drop and this voltage drop is given to the base terminal of the transistor 2, then the relay 2 will be on and automatically the relay 1 will be off as well as the buzzer. The buzzer will be switched off at the moment of lifting the handset, and then the conversation takes place through the power line. The voice signal is amplified and frequency modulated and before being coupled with the ac power line it is once again amplified as in the case of DTMF signal transmission. Similarly at the receiving end the voice signal is demodulated then fed to the called party's telephone. After the conversation is over and the handset is replaced, immediately relay 2 will be off and the contact will move to the other side. At that time relay 1 will be switched off since the voice signal was being fed to the decoder. To connect the dc supply to the buzzer the two relays should be in the on position. If any one of the relays is switched over to the other side then the dc supply will be cut off automatically.

The same process is repeated for the transmission of the speech signal from the called party and hence the full duplex operation is provided.


-12V
D
R11
R9
C8
TO PLY
12V
P10
C9
R8
-12
G
R7
G
5V
G
12V
C1
C2
C4
R5
R4
C5
C7
+12V
R6
R1
R3
R2
POW AMP
C3
C6
-12
POWERLINE INTERCOM

PCB LAYOUT

6. CHAPTER

CONCLUSION

The intercom facility using the power line has been successfully completed. Normally two cables have to run from one place to other, and they have been replaced with the power lines. In this project there is no need for switching node, Since all the power lines are connected in mesh network. Thus the cost for the intercom is reduced.

SUGGESTIONS FOR FURTHER DEVELOPMENT

a).The number of terminals can be increased by having the carrier frequencies independently.

b).The repeaters may be used to provide long distance communication.

7. CHAPTER

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VOL I and VOL II

TELTONE INNOVATING SOLUTIONS

M-8870 DTMF Receiver

The Teltone® M-8870 is a full DTMF receiver that integrates both bandsplit filter and decoder functions into a single 18-pin DIP or SOIC package. Manufactured using state-of-the-art CMOS process technology, the M-8870 offers low power consumption (35 mW max) and precise data handling. Its filter section uses switched capacitor technology for both the high and low group filters and for dial tone rejection. Its decoder uses digital counting techniques to detect and decode all 16 DTMF tone pairs into a 4-bit code. External component count is minimized by provision of an on-chip differential input amplifier, clock generator, and latched tri-state interface bus. Minimal external components required include a low-cost 3.579545 MHz color burst crystal, a timing resistor, and a timing capacitor.

The new M-8870-02 provides a "power-down" option which, when enabled, drops consumption to less than .5 mW. The -02 version can also inhibit the decoding of fourth column digits.

Features

- Low power consumption
- Adjustable acquisition and release times
- Central office quality and performance
- Power-down and inhibit modes (-02 version)
- Single 5 volt power supply
- Dial tone suppression

Applications

- Telephone switch equipment
- Mobile radio
- Remote control
- Remote data entry

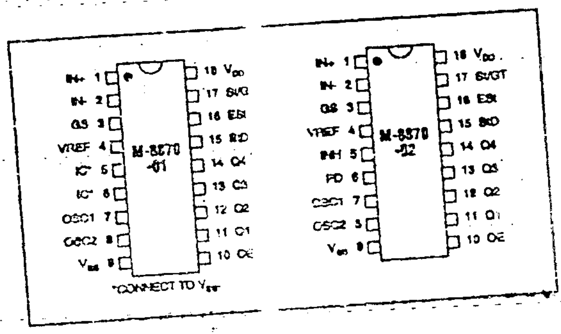


Figure 1 Pin Connections

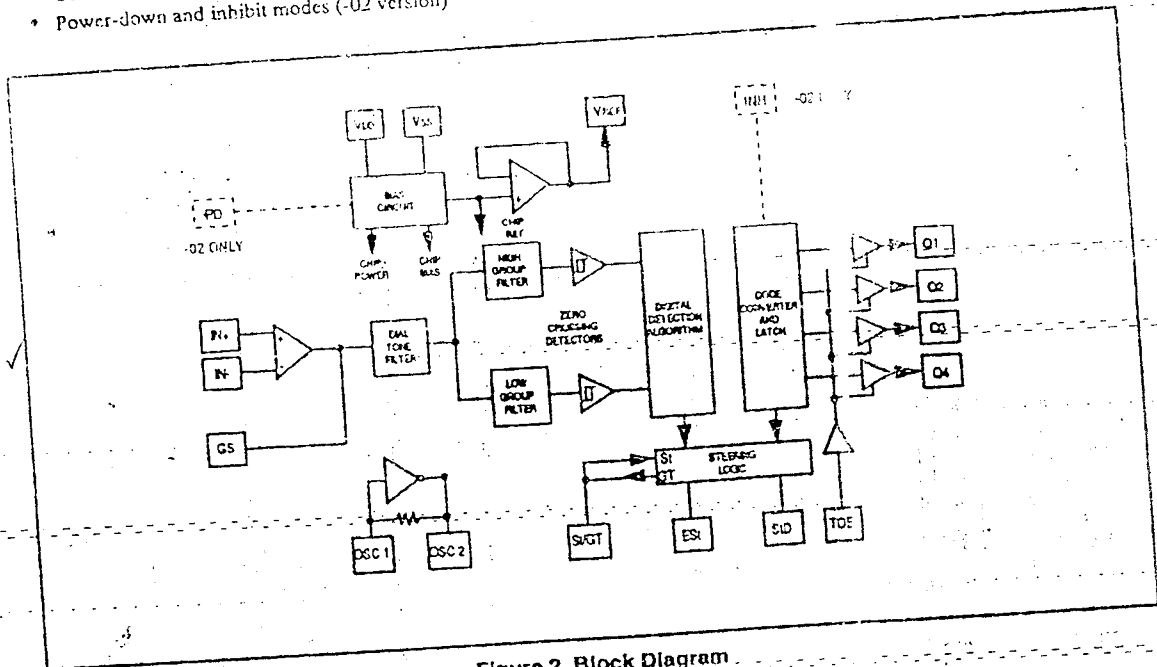


Figure 2 Block Diagram

FUNCTIONAL DESCRIPTION

M-8870 operating functions (see Figure 2) include a bandsplit filter that separates the high and low tones of the received pair, and a digital decoder that verifies both the frequency and duration of the received tones before passing the resulting 4-bit code to the output bus.

Filter

The low and high group tones are separated by applying the dual-tone signal to the inputs of two 9th order switched capacitor bandpass filters with bandwidths that correspond to the bands enclosing the low and high group tones. The filter also incorporates notches at 350 and 440 Hz, providing excellent dial tone rejection. Each filter output is followed by a single-order switched capacitor section that smooths the signals prior to limiting. Signal limiting is performed by high-gain comparators provided with hysteresis to prevent detection of unwanted low-level signals and noise. The comparator outputs provide full-rail logic swings at the frequencies of the incoming tones.

Decoder

The M-8870 decoder uses a digital counting technique to determine the frequencies of the limited tones and to verify that they correspond to standard DTMF frequencies. A complex averaging algorithm is used to protect against tone simulation by extraneous signals (such as voice) while tolerating small frequency va[illegible]ations. The algorithm ensures an optimum combination of [illegible]nmunity to talkoff and tolerance to interfering signals (thir[illegible] tones) and noise. When the detector recognizes the simult[illegible]eous presence of two valid tones (known as "signal conditi[illegible]n"), it raises the Early Steering flag (ESt). Any subsequent los[illegible] of signal condition will cause ESt to fall.

Steering Circuit

Before a decoded tone pa[illegible] is registered, the receiver checks for a valid signal duratio[illegible] referred to as "character-recognition-condition"). This ch[illegible]k is performed by an external RC time constant driven by E[illegible]t. A logic high on ESt causes V_C (see Figure 3) to rise as [illegible]e capacitor discharges. Provided that signal condition is m[illegible]ntained (ESt remains high) for the validation period (t_{GTF}), [illegible]C reaches the threshold (V_{TSt}) of the steering logic to reg[illegible]er the tone pair, thus latching its corresponding 4-bit code [illegible]e Table 5) into the output latch. At this point, the GT outpu[illegible] is activated and drives V_C to V_{DD}. GT continues to drive h[illegible]h as long as ESt remains high. Finally, after a short dela[illegible] to allow the output latch to settle, the "delayed steering" ou[illegible]put flag (StD) goes high, signaling that a received tone pair h[illegible]as been registered. The contents of the output latch are made [illegible]vailable on the 4-bit output bus by raising the three-state con[illegible]rol input (OE) to a logic high. The steering circuit works i[illegible] reverse to validate the interdigit pause between signals. T[illegible]s, as well as rejecting signals too short to be considered va[illegible]d, the receiver will tolerate signal interruptions (dropouts) [illegible]t to be considered a valid

Table 1 Pin Functions

Pin	Name	Description	
1	IN+	Non-inverting input	Connections to the front-end differential amplifier
2	IN-	Inverting input	
3	GS	Gain select. Gives access to output of front-end amplifier for connection of fe[illegible]dback resistor.	
4	V_{REF}	Reference voltage output (nominally $V_{DD}/2$). May be used to bias the inputs [illegible]t mid-rail.	
5	INH*	Inhibits detection of tones representing keys A, B, C, and D.	
6	PD*	Power down. Logic high powers down the device and inhibits the oscillator.	
7	OSC1	Clock input	3.579545 MHz crystal connected between these pins [illegible]mpletes internal oscillator.
8	OSC2	Clock output	
9	V_{SS}	Negative power supply (normally connected to 0 V).	
10	OE	Three-state output enable (input). Logic high enables the outputs Q1 - Q4. In[illegible]nal pullup.	
11-14	Q1, Q2, Q3, Q4	Three-state outputs. When enabled by OE, provides the code corresponding to the last valid tone pair received (see Table 6.)	
15	StD	Delayed steering output. Presents a logic high when a received tone pair has been registered and the output latch is updated. Returns to logic low when the voltage on St/GT falls below V_{TSt}.	
16	ESt	Early steering output. Presents a logic high immediately when the digital algorithm detects a recognizable tone pair (signal condition). Any momentary loss of signal condition will cause ESt to return to a logic low.	
17	St/GT	Steering input/guard time output (bidirectional). A voltage greater than VTSt detected at St causes the device to register the detected tone pair and update the output latch. A voltage less than VTSt frees the device to accept a new tone pair. The GT output acts to reset the external steering time constant, and its state is a function of ESt and the voltage on St. (See Table 6).	
18	V_{DD}	Positive power supply	

* -02 only. Connect to V_{SS} for -01 version.

pause. This capability, together with the ability to select the steering time constants externally, allows the designer to tailor performance to meet a wide variety of system requirements.

Guard Time Adjustment

Where independent selection of receive and pause are not required, the simple steering circuit of Figure 3 is applicable. Component values are chosen according to the formula:

$t_{REC} = t_{DP} + t_{GTP}$

$t_{GTP} \approx 0.67\ RC$

The value of t_{DP} is a parameter of the device and t_{REC} is the minimum signal duration to be recognized by the receiver. A value for C of 0.1 µF is recommended for most applications, leaving R to be selected by the designer. For example, a suitable value of R for a t_{REC} of 40 ms would be 300 K ohm. A typical circuit using this steering configuration is shown in Figure 4. The timing requirements for most telecommunication applications are satisfied with this circuit. Different steering arrangements may be used to select independently the guard times for tone-present (t_{GTP}) and tone-absent (t_{GTA}). This may be necessary to meet system specifications that place both accept and reject limits on both tone duration and interdigit pause.

Guard time adjustment also allows the designer to tailor system parameters such as talkoff and noise immunity. Increasing t_{REC} improves talkoff performance, since it reduces the probability that tones simulated by speech will maintain signal condition long enough to be registered. On the other hand, a relatively short t_{REC} with a long t_{DO} would be appropriate for extremely noisy environments where fast acquisition time and immunity to dropouts would be required. Design information for guard time adjustment is shown in Figure 5.

Input Configuration

The input arrangement of the M-8870 provides a differential input operational amplifier as well as a bias source (V_{REF}) to bias the inputs at mid-rail. Provision is made for connection of a feedback resistor to the op-amp output (GS) for gain adjustment.

In a single-ended configuration, the input pins are connected as shown in Figure 4 with the op-amp connected for unity gain and V_{REF} biasing the input at $1/2V_{DD}$. Figure 6 shows the differential configuration, which permits gain adjustment with the feedback resistor R5.

DTMF Clock Circuit

The internal clock circuit is completed with the addition of a standard 3.579545 MHz television color burst crystal. The crystal can be connected to a single M-8870 as shown in Figure 4, or to a series of M-8870s. As illustrated in Figure 7, a single crystal can be used to connect a series of M-8870s by coupling the oscillator output of each M-8870 through a 30 pF capacitor to the oscillator input of the next M-8870.

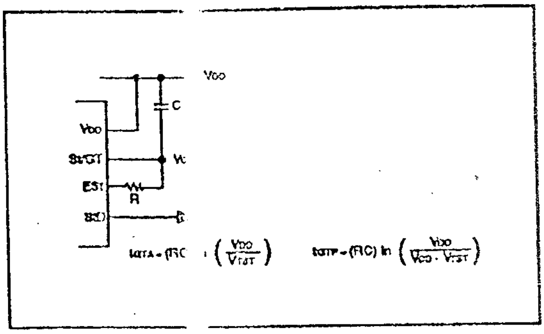


Figure 3 Basic Steering Circuit

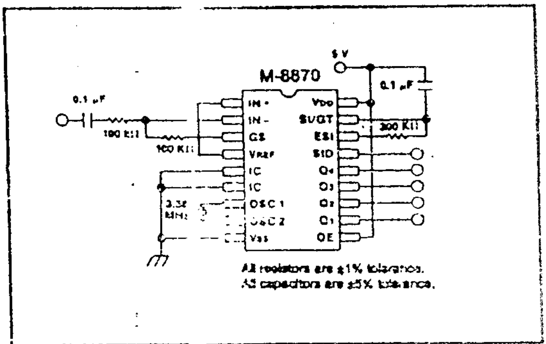


Figure 4 Single-Ended Input Configuration

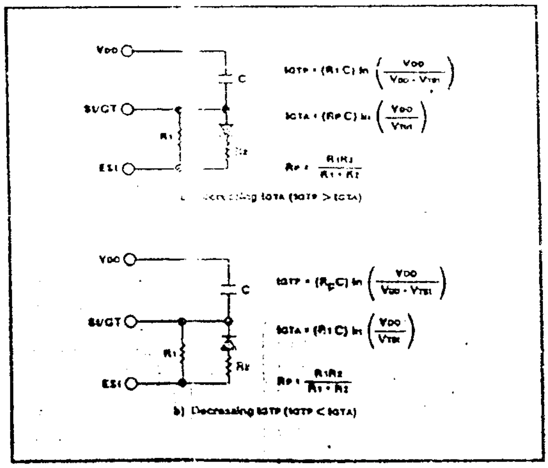


Figure 5 Guard Time Adjustment

Table 2 Absolute Maximum Ratings		
Parameter	Symbol	Value
Power supply voltage (V_{DD} - V_{SS})	V_{DD}	6.0 V max
Voltage on any pin	V_{dc}	V_{SS} -0.3, V_{DD} +0.3
Current on any pin	I_{DD}	10 mA max
Operating temperature	T_A	-40 °C to +85 °C
Storage temperature	T_S	-65 °C to +150 °C

Note: Exceeding these ratings may cause permanent damage. Functional operation under these conditions is not implied.

Table 3 DC Characteristics						
Parameter	Symbol	Min	Typ	Max	Units	Test conditions
Operating supply voltage	V_{DD}	4.75		5.25	V	
Operating supply current	I_{DD}		3.0	7.0	mA	
Standby supply current (see Note 3)	I_{DDQ}			100	µA	PD=V_{DD}
Power consumption	P_O		15	35	mW	f = 3.579 MHz, V_{DD} = 5.0 V
Low level input voltage	V_{IL}			1.5	V	
High level input voltage	V_{IH}	3.5			V	
Input leakage current	I_{IH}/I_{IL}		0.1		µA	V_{IN} = V_{SS} or V_{DD} (see Note 2)
Pullup (source) current on OE	I_{SO}		6.5	15.0	µA	OE = 0 V
Input impedance, signal inputs 1, 2	R_{IN}	8	10		M Ω	@ 1 kHz
Steering threshold voltage	V_{TSt}	2.2		2.5	V	
Low level output voltage	V_{OL}			0.03	V	No load
High level output voltage	V_{OH}	4.97			V	No load
Output low (sink) current	I_{OL}	1.0	2.5		mA	V_{OUT} = 0.4 V
Output high (source) current	I_{OH}	0.4	0.8		mA	V_{OUT} = 4.6 V
Output voltage V_{REF}	V_{REF}	2.4		2.7	V	No load
Output resistance V_{REF}	R_{OR}		10		kΩ	

Notes: 1. All voltages referenced to Vss unless otherwise noted. V_{DD} = 5.0V, V_{SS} = 0 V, T_A = 25 °C

2. Input pins defined as IN+, IN-, and OE.

3. -02 only.

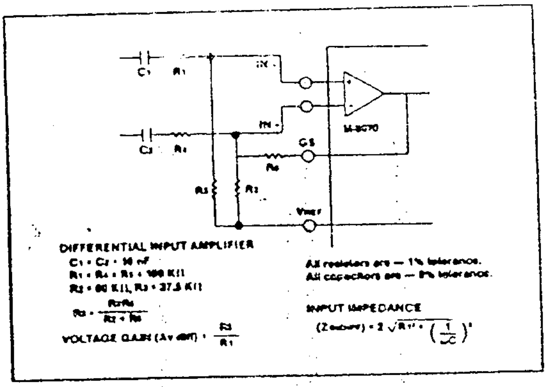


Figure 6 Differential Input Configuration

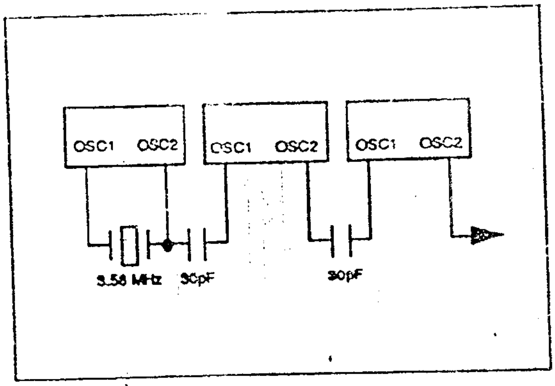


Figure 7 Common Crystal Connection

Parameter	Symbol	Min	Typ	Max	Units	Test conditions
Input leakage current	I_{IN}			±100	nA	$V_{SS} < V_{IN} < V_{DD}$
Input resistance	R_{IN}	10			MΩ	
Input offset voltage	V_{OS}			±25	mV	
Power supply rejection	PSRR	50			dB	1 kHz
Common mode rejection	CMRR	55			dB	$-3.0\ V < V_{IN} < 3.0\ V$
DC open loop voltage gain	A_{VOL}	60			dB	
Open loop unity gain bandwidth	f_C	1.2	1.5		MHz	
Output voltage swing	V_O	3.5			V_{P-P}	$R_L \geq 100\ k\Omega$ to V_{SS}
Tolerable capacitive load (GS)	C_L			100	pF	
Tolerable resistive load (GS)	R_L			50	kΩ	
Common mode range	V_{CM}	2.5			V_{P-P}	No load

Table 4 Operating Characteristics—Gain Setting Amplifier

All voltages referenced to V_{SS} unless otherwise noted. $V_{DD} = 5.0\ V$, $V_{SS} = 0\ V$, $T_A = 25\ °C$.

Table 5 Tone Decoding

F_{LOW}	F_{HIGH}	KEY (ref.)	OE	Q4	Q3	Q2	Q1
697	1209	1	H	0	0	0	1
697	1336	2	H	0	0	1	0
697	1477	3	H	0	0	1	1
770	1209	4	H	0	1	0	0
770	1336	5	H	0	1	0	1
770	1477	6	H	0	1	1	0
852	1209	7	H	0	1	1	1
852	1336	8	H	1	0	0	0
852	1477	9	H	1	0	0	1
941	1336	0	H	1	0	1	0
941	1209	*	H	1	0	1	1
941	1477	#	H	1	1	0	0
697	1633	A	H	1	1	0	1
770	1633	B	H	1	1	1	0
852	1633	C	H	1	1	1	1
941	1633	D	H	0	0	0	0
ANY	ANY	ANY	L	Z	Z	Z	Z

H = High
L = Low
Z = High Impedance

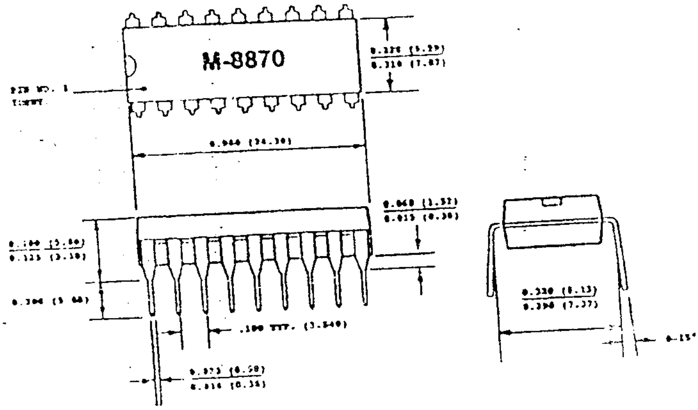
Table 6 AC Characteristics

Parameter	Symbol	Min	Typ	Max	Units	Notes
Valid input signal levels (each tone of composite signal)		-29		+1	dBm	1,2,3,4,5,8
		27.5		869	mVRMS	
Positive twist accept				10	dB	2,3,4,8
Negative twist accept				10	dB	
Frequency deviation accept limit				1.5% ±2 Hz	Nom.	2,3,5,8,10
Frequency deviation reject limit		±3.5%			Nom.	2,3,5
Third tone tolerance		-25	-16		dB	2,3,4,5,8,9,13,14
Noise tolerance			-12		dB	2,3,4,5,6,8,9
Dial tone tolerance		+18	+22		dB	2,3,4,5,7,8,9
Tone present detection time	t_{DP}	5	8	14	ms	See Figure 3
Tone absent detection time	t_{DA}	0.5	3	8.5	ms	
Minimum tone duration accept	t_{REC}			40	ms	User adjustable (times shown are obtained with circuit in Figure 5)
Maximum tone duration reject	t_{REC}	20			ms	
Minimum interdigit pause accept	t_{ID}			40	ms	
Maximum interdigit pause reject	t_{DO}	20			ms	
Propagation delay (St to Q)	t_{PQ}		6	11	µs	OE = V_{DD}
Propagation delay (St to StD)	t_{PStD}		9	16	µs	
Output data setup (Q to StD)	t_{QStD}		4.0		µs	
Propagation delay (OE to Q), enable	t_{PTE}		50	60	ns	R_L = 10kΩ, C_L = 50 pF
Propagation delay (OE to Q), disable	t_{PTD}		300		ns	
Crystal clock frequency	f_{CLK}	3.5759	3.5795	3.5831	MHz	
Clock output (OSC2), capacitive load	C_{LO}			30	pF	

All voltages referenced to VSS unless otherwise noted. VDD = 5.0 V, VSS = 0 V, TA = 25 °C, fCLK = 3.5795 MHz using circuit in Figure 5.

Notes:
1. dBm = decibels above or below a reference power of 1 mW into a 600 ohm load.
2. Digit sequence consists of all 16 DTMF tones.
3. Tone duration = 40 ms. Tone pause = 40 ms.
4. Nominal DTMF frequencies are used.
5. Both tones in the composite signal have an equal amplitude.
6. Bandwidth limited (0 to 3 kHz) Gaussian noise.
7. The precise dial tone frequencies are (350 and 440 Hz) ± 2%.
8. For an error rate of better than 1 in 10,000.
9. Referenced to lowest level frequency component in DTMF signal.
10. Minimum signal acceptance level is measured with specified maximum frequency deviation.
11. Input pins defined as IN+, IN-, and OE.
12. External voltage source used to bias VREF.
13. This parameter also applies to a third tone injected onto the power supply.
14. Referenced to Figure 5. Input DTMF tone level at -28 dBm.

Plastic and CERDIP Package



Surface Mount Package

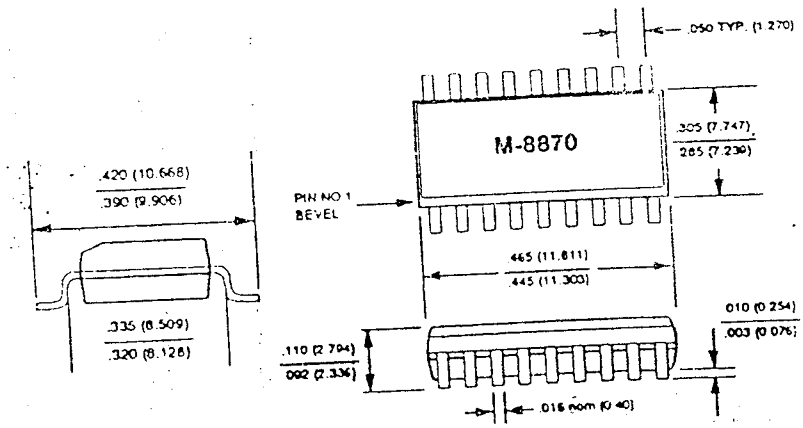
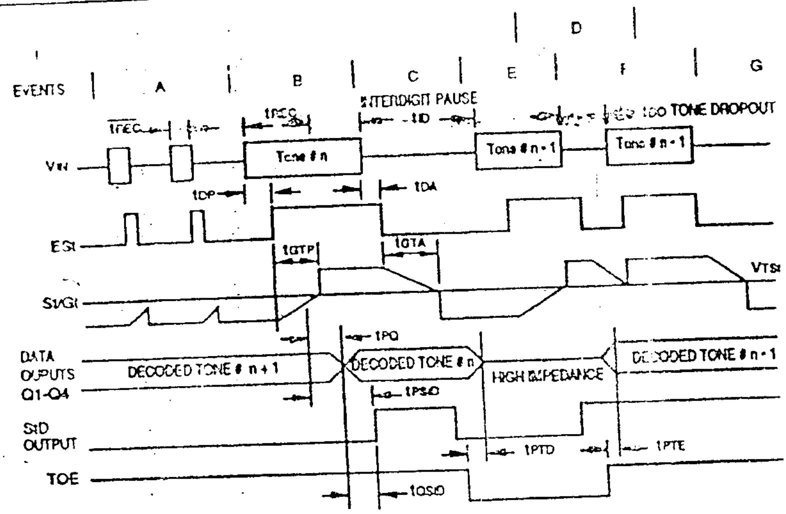


Figure 8 Package Dimensions



Explanation of Events

(A) Tone bursts detected, tone duration invalid, outputs not updated.
(B) Tone #n detected, tone duration valid, tone decoded and latched in outputs.
(C) End of tone #n detected, tone absent duration valid, outputs remain latched until next valid tone.
(D) Outputs switched to high impedance state.
(E) Tone #n + 1 detected, tone duration valid, tone decoded and latched in outputs (currently high impedance).
(F) Acceptable dropout of tone #n + 1, tone absent duration invalid, outputs remain latched.
(G) End of tone #n + 1 detected, tone absent duration valid, outputs remain latched until next valid tone.

Explanation of Symbols

VIN	DTMF composite input signal.
ESt	Early steering output. Indicates detection of valid tone frequencies.
St/GT	Steering input/guard time output. Drives external RC timing circuit.
Q1 - Q4	4-bit decoded tone output.
StD	Delayed steering output. Indicates that valid frequencies have been present/absent for the required guard time, thus constituting a valid signal.
OE	Output enable (input). A low level shifts Q1 - Q4 to its high impedance state.
$\overline{t_{REC}}$	Maximum DTMF signal duration not detected as valid.
t_{REC}	Minimum DTMF signal duration required for valid recognition.
t_{ID}	Minimum time between valid DTMF signals.
t_{DO}	Maximum allowable dropout during valid DTMF signal.
t_{DP}	Time to detect the presence of valid DTMF signals.
t_{DA}	Time to detect the absence of valid DTMF signals.
t_{GTP}	Guard time, tone present.
t_{GTA}	Guard time, tone absent.

Figure 9 Timing Diagram

Ordering Information

M-8870-01 DTMF Receiver	Plastic Package	M-8870-02 DTMF Receiver	Plastic Package
M-8870-01C DTMF Receiver	CERDIP Package	M-8870-02SM DTMF Receiver	Surface Mount Package
M-8870-01SM DTMF Receiver	Surface Mount Package		

EXAR

XR-2206

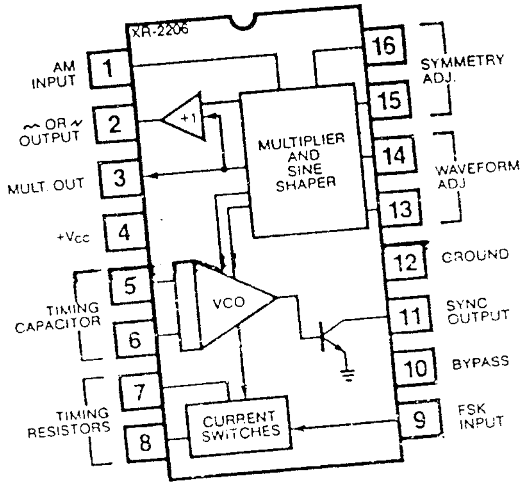
Monolithic Function Generator

GENERAL DESCRIPTION

The XR-2206 is a monolithic function generator integrated circuit capable of producing high quality sine, square, triangle, ramp, and pulse waveforms of high-stability and accuracy. The output waveforms can be both amplitude and frequency modulated by an external voltage. Frequency of operation can be selected externally over a range of 0.01 Hz to more than 1 MHz.

The circuit is ideally suited for communications, instrumentation, and function generator applications requiring sinusoidal tone, AM, FM, or FSK generation. It has a typical drift specification of 20 ppm/°C. The oscillator frequency can be linearly swept over a 2000:1 frequency range, with an external control voltage, having a very small affect on distortion.

FUNCTIONAL BLOCK DIAGRAM



FEATURES

Low-Sine Wave Distortion	0.5%, Typical
Excellent Temperature Stability	20 ppm/°C, Typical
Wide Sweep Range	2000:1, Typical
Low-Supply Sensitivity	0.01%V, Typical
Linear Amplitude Modulation	
TTL Compatible FSK Controls	
Wide Supply Range	10V to 26V
Adjustable Duty Cycle	1% to 99%

ORDERING INFORMATION

Part Number	Package	Operating Temperature
XR-2206M	Ceramic	−55°C to +125°C
XR-2206N	Ceramic	0°C to +70°C
XR-2206P	Plastic	0°C to +70°C
XR-2206CN	Ceramic	0°C to +70°C
XR-2206CP	Plastic	0°C to +70°C

APPLICATIONS

- Waveform Generation
- Sweep Generation
- AM/FM Generation
- V/F Conversion
- FSK Generation
- Phase-Locked Loops (VCO)

SYSTEM DESCRIPTION

The XR-2206 is comprised of four functional blocks; a voltage-controlled oscillator (VCO), an analog multiplier and sine shaper; a unity gain buffer amplifier; and a set of current switches.

The VCO actually produces an output frequency proportional to an input current, which is produced by a resistor from the timing terminals to ground. The current switches route one of the timing pins current to the VCO controlled by an FSK input pin, to produce an output frequency. With two timing pins, two discrete output frequencies can be independently produced for FSK Generation Applications.

ABSOLUTE MAXIMUM RATINGS

Power Supply	26V
Power Dissipation	750 mW
Derate Above 25°C	5 mW/°C
Total Timing Current	6 mA
Storage Temperature	−65°C to +150°C

XR-2206

ELECTRICAL CHARACTERISTICS

Test Conditions: Test Circuit of Figure 1, $V^+ = 12V$, $T_A = 25°$, $C = 0.01\ \mu F$, $R_1 = 100\ k\Omega$, $R_2 = 10\ k\Omega$, $R_3 = 25\ k\Omega$ unless otherwise specified. S_1 open for triangle, closed for sine wave.

PARAMETERS	XR-2206M MIN	XR-2206M TYP	XR-2206M MAX	XR-2206C MIN	XR-2206C TYP	XR-2206C MAX	UNITS	CONDITIONS
GENERAL CHARACTERISTICS								
Single Supply Voltage	10		26	10		26	V	
Split-Supply Voltage	±5		±13	±5		±13	V	
Supply Current		12	17		14	20	mA	$R_1 \geq 10\ k\Omega$
OSCILLATOR SECTION								
Max. Operating Frequency	0.5	1		0.5	1		MHz	$C = 1000\ pF$, $R_1 = 1\ k\Omega$
Lowest Practical Frequency		0.01			0.01		Hz	$C = 50\ \mu F$, $R_1 = 2\ M\Omega$
Frequency Accuracy		±1	±4		±2		% of f_o	$f_o = 1/R_1C$
Temperature Stability		±10	±50		±20		ppm/°C	$0°C \leq T_A \leq 70°C$, $R_1 = R_2 = 20\ k\Omega$
Supply Sensitivity		0.01	0.1		0.01		%/V	$V_{LOW} = 10V$, $V_{HIGH} = 20V$, $R_1 = R_2 = 20\ k\Omega$
Sweep Range	1000:1	2000:1			2000:1		$f_H = f_L$	f_H @ $R_1 = 1\ k\Omega$ f_L @ $R_1 = 2\ M\Omega$
Sweep Linearity								
10:1 Sweep		2			2		%	$f_L = 1\ kHz$, $f_H = 10\ kHz$
1000:1 Sweep		8			8		%	$f_L = 100\ kHz$, $f_H = 100\ kHz$
FM Distortion		0.1			0.1		%	±10% Deviation
Recommended Timing Components								
Timing Capacitor: C	0.001		100	0.001		100	µF	See Figure 4.
Timing Resistors: R_1 & R_2	1		2000	1		2000	kΩ	
Triangle Sine Wave Output								See Note 1, Figure 2.
Triangle Amplitude		160			160		mV/kΩ	Figure 1, S_1 Open
Sine Wave Amplitude	40	60	80		60		mV/kΩ	Figure 1, S_1 Closed
Max. Output Swing		6			6		V p-p	
Output Impedance		600			600		Ω	
Triangle Linearity		1			1		%	
Amplitude Stability		0.5			0.5		dB	For 1000:1 Sweep
Sine Wave Amplitude Stability		4800			4600		ppm/°C	See Note 2.
Sine Wave Distortion								
Without Adjustment		2.5			2.5		%	$R_1 = 30\ k\Omega$
With Adjustment		0.4	1.0		0.5	1.5	%	See Figures 6 and 7
Amplitude Modulation								
Input Impedance	50	100		50	100		kΩ	
Modulation Range		100			100		%	
Carrier Suppression		55			55		dB	
Linearity		2			2		%	For 95% modulation
Square-Wave Output								
Amplitude		12			12		V p-p	Measured at Pin 11.
Rise Time		250			250		nsec	$C_L = 10\ pF$
Fall Time		50			50		nsec	$C_L = 10\ pF$
Saturation Voltage		0.2	0.4		0.2	0.6	V	$I_L = 2\ mA$
Leakage Current		0.1	20		0.1	100	µA	$V_{11} = 26V$
FSK Keying Level (Pin 9)	0.8	1.4	2.4	0.8	1.4	2.4	V	See section on circuit controls
Reference Bypass Voltage	2.9	3.1	3.3	2.5	3	3.5	V	Measured at Pin 10.

Note 1: Output amplitude is directly proportional to the resistance, R_3, on Pin 3. See Figure 2.

Note 2: For maximum amplitude stability, R_3 should be a positive temperature coefficient resistor.

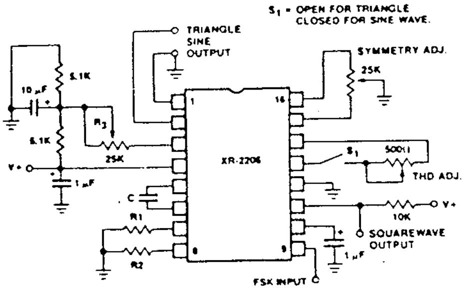


Figure 1. Basic Test Circuit.

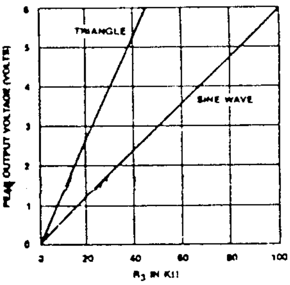


Figure 2. Output Amplitude as a Function of the Resistor, R_3, at Pin 3.

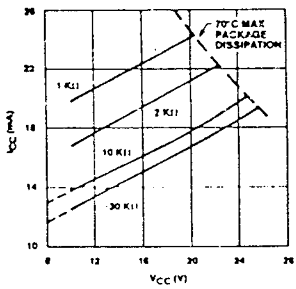


Figure 3. Supply Current versus Supply Voltage, Timing, R.

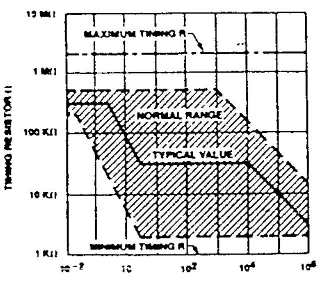


Figure 4. R versus Oscillation Frequency.

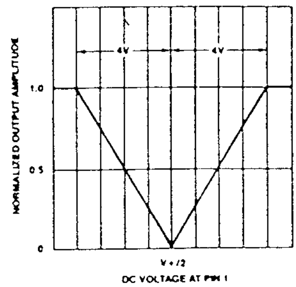


Figure 5. Normalized Output Amplitude versus DC Bias at AM Input (Pin 1).

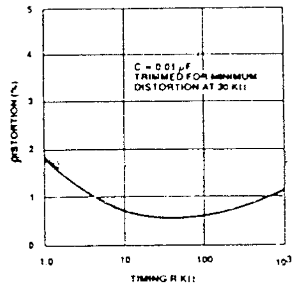


Figure 6. Trimmed Distortion versus Timing Resistor.

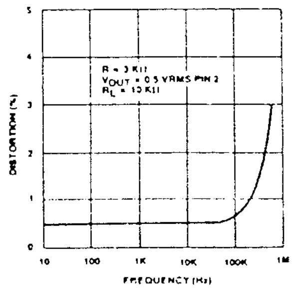


Figure 7. Sine Wave Distortion versus Operating Frequency with Timing Capacitors Varied.

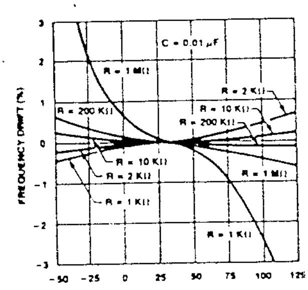


Figure 8. Frequency Drift versus Temperature.

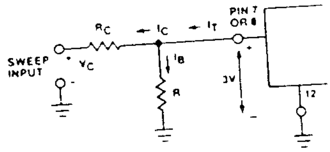


Figure 9. Circuit Connection for Frequency Sweep.

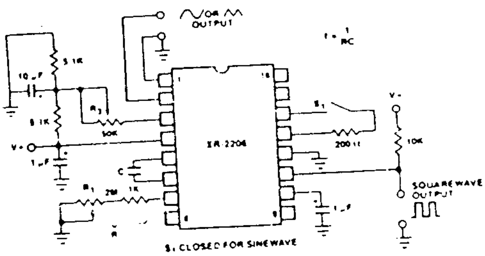


Figure 10. Circuit for Sine Wave Generation without External Adjustment. (See Figure 2 for Choice of R_3).

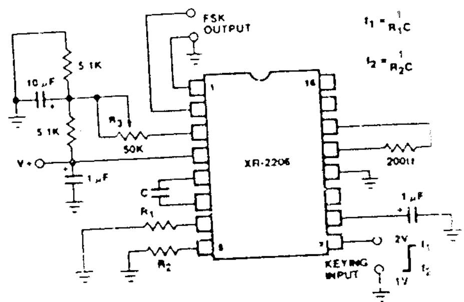


Figure 12. Sinusoidal FSK Generator.

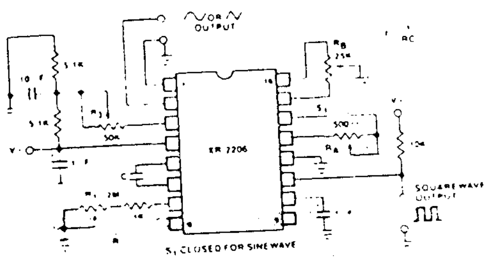


Figure 11. Circuit for Sine Wave Generation with Minimum Harmonic Distortion. (R_3 Determines Output Swing—See Figure 2.)

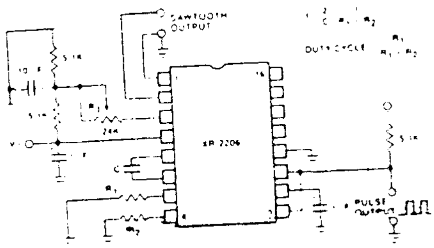


Figure 13. Circuit for Pulse and Ramp Generation.

Frequency-Shift Keying:

The XR-2206 can be operated with two separate timing resistors, R_1 and R_2, connected to the timing Pin 7 and 8, respectively, as shown in Figure 12. Depending on the polarity of the logic signal at Pin 9, either one or the other of these timing resistors is activated. If Pin 9 is open-circuited or connected to a bias voltage $\geq 2V$, only R_1 is activated. Similarly, if the voltage level at Pin 9 is $\leq 1V$, only R_2 is activated. Thus, the output frequency can be keyed between two levels, f_1 and f_2, as:

$$f_1 = 1/R_1C \text{ and } f_2 = 1/R_2C$$

For split-supply operation, the keying voltage at Pin 9 is referenced to V−.

Output DC Level Control:

The dc level at the output (Pin 2) is approximately the same as the dc bias at Pin 3. In Figures 10, 11 and 12, Pin 3 is biased midway between V+ and ground, to give an output dc level of $\approx V+/2$.

APPLICATIONS INFORMATION

Sine Wave Generation

Without External Adjustment:

Figure 10 shows the circuit connection for generating a sinusoidal output from the XR-2206. The potentiometer, R_1 at Pin 7, provides the desired frequency tuning. The maximum output swing is greater than V+/2, and the typical distortion (THD) is <2.5%. If lower sine wave distortion is desired, additional adjustments can be provided as described in the following section.

The circuit of Figure 10 can be converted to split-supply operation, simply by replacing all ground connections with V−. For split-supply operation, R_3 can be directly connected to ground.

With External Adjustment:

The harmonic content of sinusoidal output can be reduced to $\approx 0.5\%$ by additional adjustments as shown in Figure 11. The potentiometer, R_A, adjusts the sine-shaping resistor, and R_B provides the fine adjustment for the waveform symmetry. The adjustment procedure is as follows:

1. Set R_B at midpoint, and adjust R_A for minimum distortion.

2. With R_A set as above, adjust R_B to further reduce distortion.

Triangle Wave Generation

The circuits of Figures 10 and 11 can be converted to triangle wave generation, by simply open-circuiting Pin 13 and 14 (i.e., S_1 open). Amplitude of the triangle is approximately twice the sine wave output.

FSK Generation

Figure 12 shows the circuit connection for sinusoidal FSK signal operation. Mark and space frequencies can be independently adjusted, by the choice of timing resistors, R_1 and R_2; the output is phase-continuous during transitions. The keying signal is applied to Pin 9. The circuit can be converted to split-supply operation by simply replacing ground with V−.

Pulse and Ramp Generation

Figure 13 shows the circuit for pulse and ramp waveform generation. In this mode of operation, the FSK keying terminal (Pin 9) is shorted to the square-wave output (Pin 11), and the circuit automatically frequency-shift keys itself between two separate frequencies during the positive-going and negative-going output waveforms. The pulse width and duty cycle can be adjusted from 1% to 99%, by the choice of R_1 and R_2. The values of R_1 and R_2 should be in the range of 1 kΩ to 2 MΩ.

PRINCIPLES OF OPERATION

Description of Controls

Frequency of Operation:

The frequency of oscillation, f_0, is determined by the external timing capacitor, C, across Pin 5 and 6, and by the timing resistor, R, connected to either Pin 7 or 8. The frequency is given as:

$$f_0 = \frac{1}{RC} \text{ Hz}$$

and can be adjusted by varying either R or C. The recommended values of R, for a given frequency range, as shown in Figure 4. Temperature stability is optimum for 4 kΩ < R < 200 kΩ. Recommended values of C are from 1000 pF to 100 μF.

Frequency Sweep and Modulation:

Frequency of oscillation is proportional to the total timing current, I_T, drawn from Pin 7 or 8:

$$f = \frac{320\ I_T\ (mA)}{C\ (\mu F)} \text{ Hz}$$

Timing terminals (Pin 7 or 8) are low-impedance points, and are internally biased at +3V, with respect to Pin 12. Frequency varies linearly with I_T, over a wide range of current values, from 1 μA to 3 mA. The frequency can be controlled by applying a control voltage, V_C, to the activated timing pin as shown in Figure 9. The frequency of oscillation is related to V_C as:

$$f = \frac{1}{RC}\ 1 + \frac{R}{R_C}(1 - \frac{V_C}{3}) \text{ Hz}$$

XR-2206

where V_C is in volts. The voltage-to-frequency conversion gain, K, is given as:

$$K = \partial f/\partial V_C = -\frac{0.32}{R_C C}\ \text{Hz/V}$$

CAUTION: For safety operation of the circuit, I_T should be limited to ≤ 3 mA.

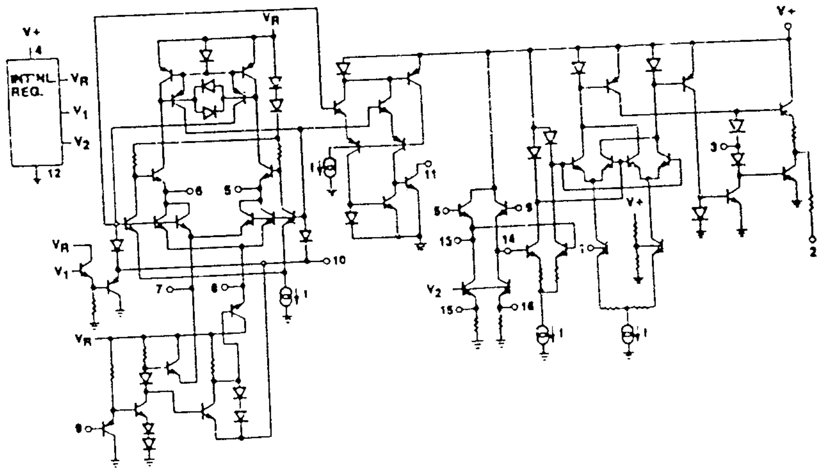
Output Amplitude:

Maximum output amplitude is inversely proportional to the external resistor, R_3, connected to Pin 3 (see Figure 2). For sine wave output, amplitude is approximately 60 mV peak per kΩ of R_3; for triangle, the peak amplitude is approximately 160 mV peak per kΩ of R_3. Thus, for example, R_3 = 50 kΩ would produce approximately ±3V sinusoidal output amplitude.

Amplitude Modulation:

Output amplitude can be modulated by applying a dc bias and a modulating signal to Pin 1. The internal impedance at Pin 1 is approximately 100 kΩ. Output amplitude varies linearly with the applied voltage at Pin 1, for values of dc bias at this pin, within ±4 volts of $V^+/2$ as shown in Figure 5. As this bias level approaches $V^+/2$, the phase of the output signal is reversed, and the amplitude goes through zero. This property is suitable for phase-shift keying and suppressed-carrier AM generation. Total dynamic range of amplitude modulation is approximately 55 dB.

CAUTION: AM control must be used in conjunction with a well-regulated supply, since the output amplitude now becomes a function of V^+.



EQUIVALENT SCHEMATIC DIAGRAM

Raytheon

FSK Demodulator/ Tone Decoder

XR-2211

Features

- Wide frequency range — 0.01Hz to 300kHz
- Wide supply voltage range — 4.5V to 20V
- DTL/TTL/ECL logic compatibility
- FSK demodulation with carrier-detector
- Wide dynamic range — 2mV to $3V_{RMS}$
- Adjustable tracking range — ±1% to ±80%
- Excellent temperature stability — 20ppm/°C typical

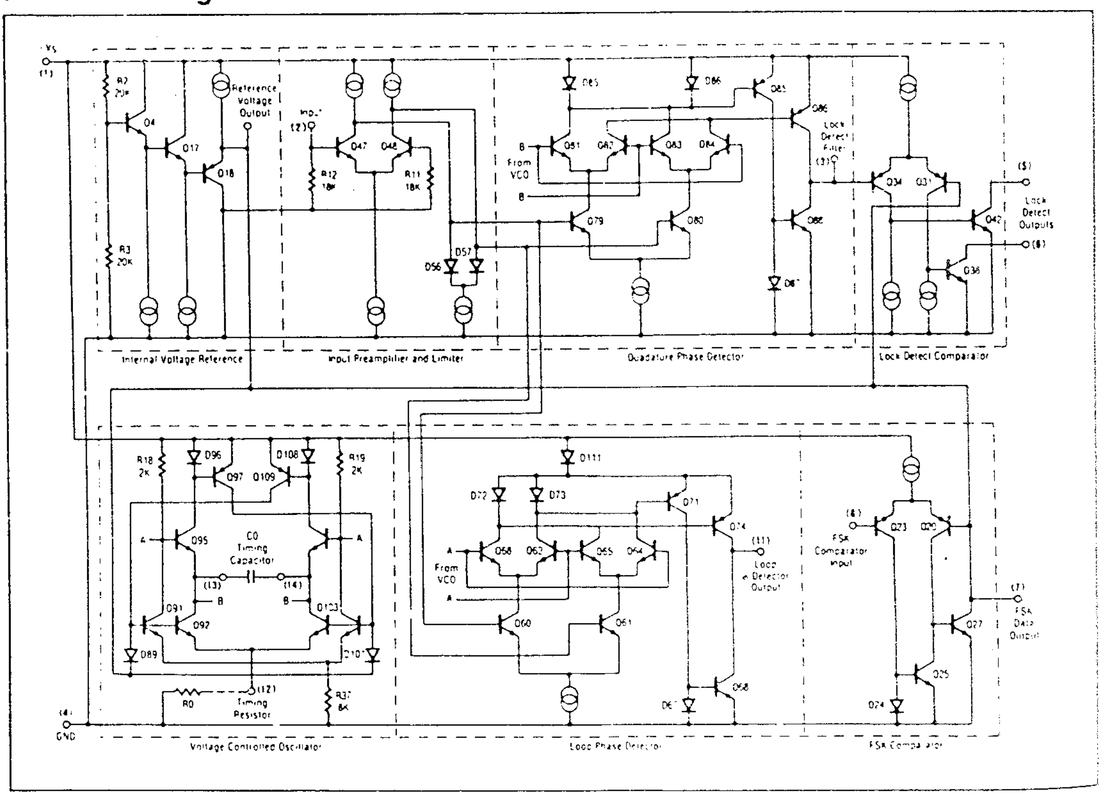
Applications

- FSK demodulation
- Data synchronization
- Tone decoding
- FM detection
- Carrier detection

Description

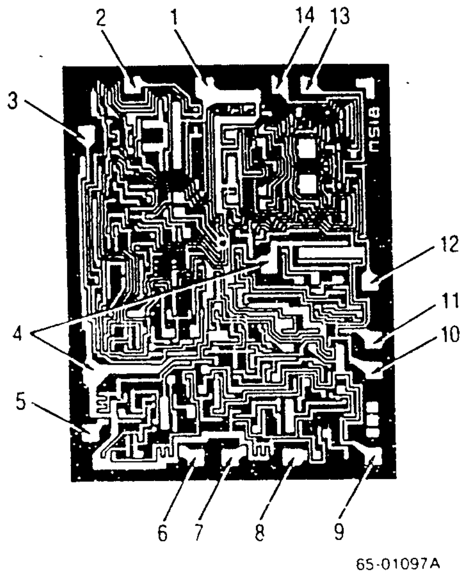
The XR-2211 is a monolithic phase-locked loop (PLL) system especially designed for data communications. It is particularly well suited for FSK modem applications, and operates over a wide frequency range of 0.01Hz to 300kHz. It can accommodate analog signals between 2mV and 3V, and can interface with conventional DTL, TTL and ECL logic families. The circuit consists of a basic PLL for tracking an input signal frequency within the passband, a quadrature phase detector which provides carrier detection, and an FSK voltage comparator which provides FSK demodulation. External components are used to independently set carrier frequency, bandwidth, and output delay.

Schematic Diagram



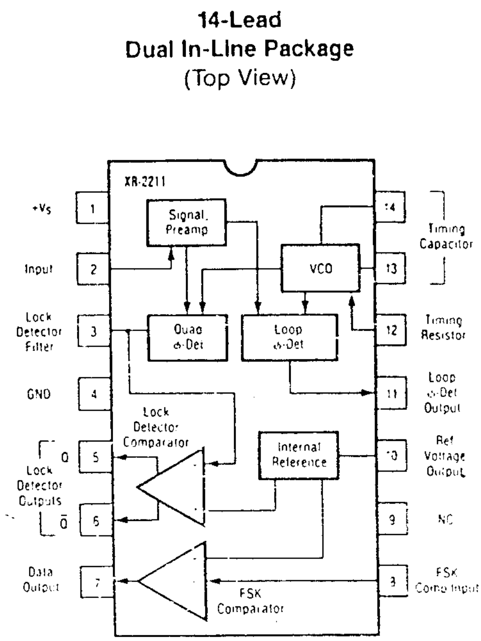
Courtesy of Raytheon Company.

Mask Pattern

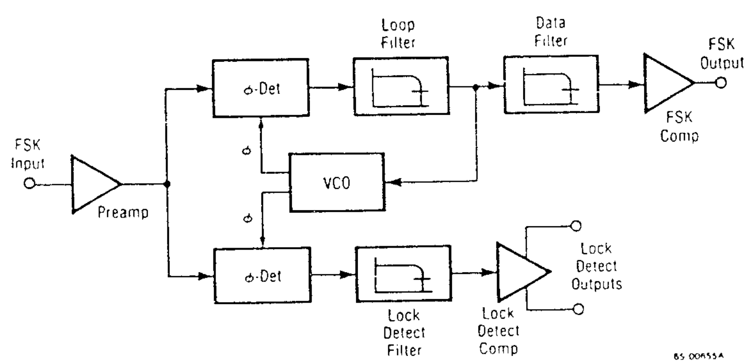


Die Size: 103 x 80 mils
Min. Pad Dimension: 4 x 4 mils

Connection Information



Functional Block Diagram



Courtesy of Ratheon Company.

Absolute Maximum Ratings

Supply Voltage .... +20V
Input Signal Level .... $3V_{RMS}$
Storage Temperature
Range .... -65°C to +150°C

Operating Temperature Range
XR-2211CN/CP .... 0°C to +75°C
XR-2211N/P .... -40°C to +85°C
XR-2211M .... -55°C to +125°C
Lead Soldering Temperature
(60 Sec) .... +300°C

Thermal Characteristics

	14-Lead Plastic DIP	14-Lead Ceramic DIP
Max. Junction Temp.	125°C	175°C
Max. P_D T_A < 50°C	468mW	1042mW
Therm. Res. θ_{JC}	—	50°C/W
Therm. Res. θ_{JA}	160°C/W	120°C/W
For T_A > 50°C Derate at	6.25mW per °C	6.33mW per °C

Ordering Information

Part Number	Package	Operating Temperature Range
XR-2211CN	Ceramic	0°C to +75°C
XR-2211CP	Plastic	0°C to +75°C
XR-2211N	Ceramic	-40°C to +85°C
XR-2211P	Plastic	-40°C to +85°C
XR-2211M	Ceramic	-55°C to +125°C
XR-2211M/883B*	Ceramic	-55°C to +125°C

*MIL-STD-883, Level B Processing

Electrical Characteristics (Test Conditions $+V_S$ = +12V, T_A = +25°C, R0 = 30kΩ, C0 = 0.033μF. See Figure 1 for component designations.)

Parameters	Test Conditions	XR-2211/M			XR-2211C			Units
		Min	Typ	Max	Min	Typ	Max	
General								
Supply Voltage		4.5		20	4.5		20	V
Supply Current	R0 ≥ 10kΩ		4.0	9.0		5.0	11	mA
Oscillator								
Frequency Accuracy	Deviation from f_0 = 1/R0C0		±1.0	±3.0		±1.0		%
Frequency Stability Temperature Coefficient	R1 = ∞		±20	±50		±20		ppm/°C
Power Supply Rejection	$+V_S$ = 12 ±1V		0.05	0.5		0.05		%/V
	$+V_S$ = 5 ±0.5V		0.2			0.2		%/V
Upper Frequency Limit	R0 = 8.2kΩ, C0 = 400pF	100	300			300		kHz
Lowest Practical Operating Frequency	R0 = 2MΩ C0 = 50μF			0.01		0.01		Hz
Timing Resistor, R0 Operating Range		5.0		2000	5.0		2000	kΩ
Recommended Range		15		100	15		100	kΩ

Courtesy of Ratheon Company.

Description of Circuit Controls

Signal Input (Pin 2)
The input signal is AC coupled to this terminal. The internal impedance at pin 2 is 20kΩ. Recommended input signal level is in the range of $10mV_{RMS}$ to $3V_{RMS}$.

Quadrature Phase Detector Output (Pin 3)
This is the high-impedance output of the quadrature phase detector, and is internally connected to the input of lock-detect voltage comparator. In tone detection applications, pin 3 is connected to ground through a parallel combination of R_D and C_D (see Figure 1) to eliminate chatter at the lock-detect outputs. If this tone-detect section is not used, pin 3 can be left open circuited.

Lock-Detect Output, Q (Pin 5)
The output at pin 5 is at a "high" state when the PLL is out of lock and goes to a "low" or conducting state when the PLL is locked. It is an open collector type output and requires a pull-up resistor, R_L, to $+V_S$ for proper operation. In the "low" state it can sink up to 5mA of load current.

Lock-Detect Complement, Q (Pin 6)
The output at pin 6 is the logic complement of the lock-detect output at pin 5. This output is also an open collector type stage which can sink 5mA of load current in the low or "on" state.

FSK Data Output (Pin 7)
This output is an open collector logic stage which requires a pull-up resistor, R_L, to $+V_S$ for proper operation. It can sink 5mA of load current. When decoding FSK signals the FSK data output will switch to a "high" or off state for low input frequency, and will switch to a "low" or on state for high input frequency. If no input signal is present, the logic state at pin 7 is indeterminate.

FSK Comparator Input (Pin 8)
This is the high-impedance input to the FSK voltage comparator. Normally, an FSK post-detection or data filter is connected between this terminal and the PLL phase-detector output (pin 11). This data filter is formed by R_F and C_F of Figure 1. The threshold voltage of the comparator is set by the internal reference voltage, V_R, available at pin 10.

Reference Voltage, V_R (Pin 10)
This pin is internally biased at the reference voltage level, V_R; V_R = V+/2 - 650mV. The DC voltage level at this pin forms an internal reference

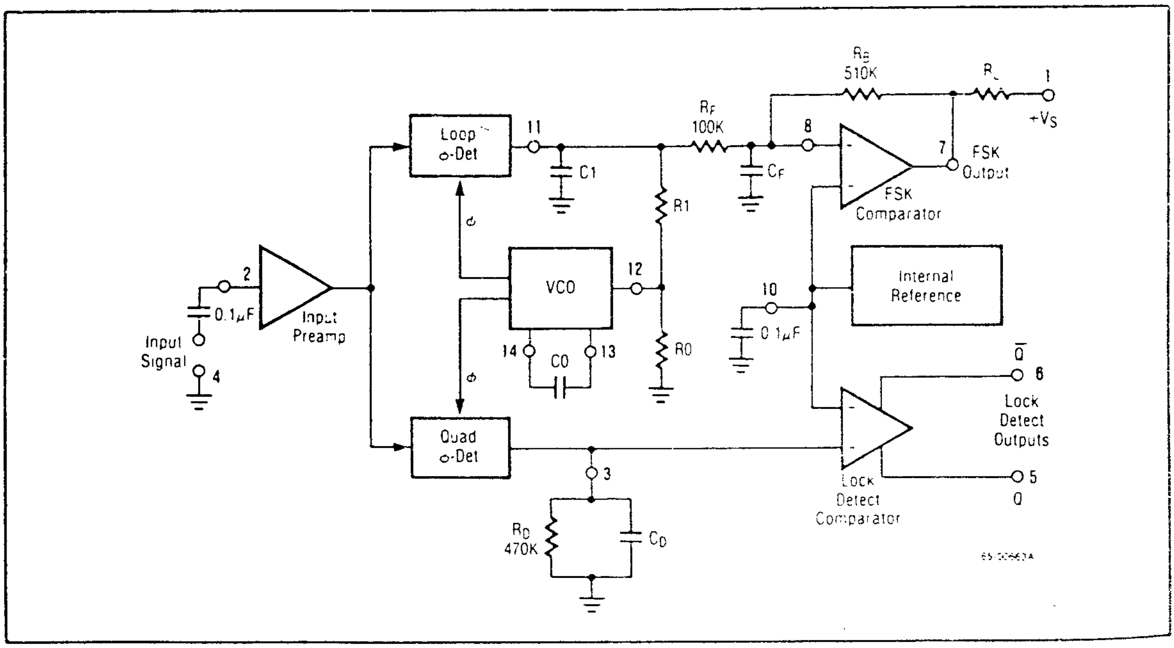


Figure 1. Generalized Circuit Connection for FSK and Tone Detection

Courtesy of Ratheon Company.

for the voltage levels at pin 3, 8, 11, and 12. Pin 10 must be bypassed to ground with a 0.1μF capacitor.

Loop Phase Detector Output (Pin 11)
This terminal provides a high impedance output for the loop phase-detector. The PLL loop filter is formed by R1 and C1 connected to pin 11 (see Figure 1). With no input signal, or with no phase error within the PLL, the DC level at pin 11 is very nearly equal to V_R. The peak voltage swing available at the phase detector output is equal to $\pm V_R$.

VCO Control Input (Pin 12)
VCO free-running frequency is determined by external timing resistor, R0, connected from this terminal to ground. The VCO free-running frequency, f_0, is given by:

$$f_0(\text{Hz}) = \frac{1}{R0C0}$$

where C0 is the timing capacitor across pins 13 and 14. For optimum temperature stability R0 must be in the range of 10kΩ to 100kΩ (see Typical Electrical Characteristics).

This terminal is a low impedance point, and is internally biased at a DC level equal to V_R. The maximum timing current drawn from pin 12 must be limited to ≤3mA for proper operation of the circuit.

VCO Timing Capacitor (Pins 13 and 14)
VCO frequency is inversely proportional to the external timing capacitor, C0, connected across these terminals. C0 must be non-polarized, and in the range of 200pF to 10μF.

VCO Frequency Adjustment
VCO can be fine tuned by connecting a potentiometer, R_X, in series with R0 at pin 12 (see Figure 2).

VCO Free-Running Frequency, f_0
The XR-2211 does not have a separate VCO output terminal. Instead, the VCO outputs are internally connected to the phase-detector sections of the circuit. However, for setup or adjustment purposes, the VCO free-running frequency can be measured at pin 3 (with C_D disconnected) with no input and with pin 2 shorted to pin 10.

Design Equations

See Figure 1 for Definitions of Components.

1. VCO Center Frequency, f_0:

$$f_0(\text{Hz}) = \frac{1}{R0C0}$$

2. Internal Reference Voltage, V_R (measured at pin 10):

$$V_R = \left(\frac{+V_S}{2}\right) - 650\text{mV}$$

3. Loop Lowpass Filter Time Constant, τ:

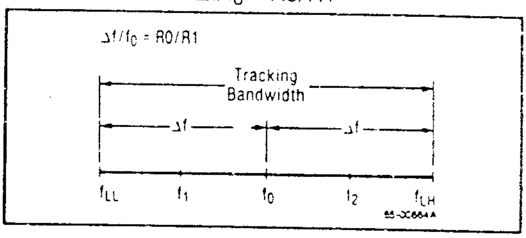
$$\tau = R1C1$$

4. Loop Damping, ζ:

$$\zeta = \left(\sqrt{\frac{C0}{C1}}\right)\left(\frac{1}{4}\right)$$

5. Loop Tracking Bandwidth, $\pm\Delta f/f_0$:

$$\Delta f/f_0 = R0/R1$$



6. FSK Data Filter Time Constant, τ_F:

$$\tau_F = R_F C_F$$

7. Loop Phase Detector Conversion Gain, K_ϕ: (K_ϕ is the differential DC voltage across pins 10 and 11, per unit of phase error at phase-detector input):

$$K\phi \text{ (in volts per radian)} = \frac{(-2)(V_R)}{\pi}$$

8. VCO Conversion Gain, K0, is the amount of change in VCO frequency per unit of DC voltage change at pin 11:

$$K0 \text{ (in Hertz per volt)} = \frac{-1}{C0R1V_R}$$

9. Total Loop Gain, K_T:

K_T (in radians per second per volt) = $2\pi K\phi K0$ = $4/C0R1$

10. Peak Phase-Detector Current, I_A:

$$I_A \text{ (mA)} = \frac{V_R}{25}$$

Courtesy of Ratheon Company.

FSK Decoding With Carrier Detect

The lock-detect section of the XR-2211 can be used as a carrier detect option for FSK decoding. The recommended circuit connection for this application is shown in Figure 3. The open-collector lock-detect output, pin 6, is shorted to the data output (pin 7). Thus, the data output will be disabled at "low" state, until there is a carrier within the detection band of the PLL, and the pin 6 output goes "high" to enable the data output.

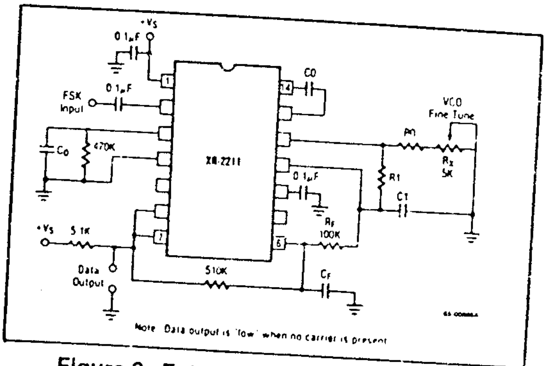


Figure 3. External Connectors for FSK Demodulation With Carrier Detect Capability

The minimum value of the lock-detect filter capacitance C_D is inversely proportional to the capture range, $\pm\Delta f_c$. This is the range of incoming frequencies over which the loop can acquire lock and is always less than the tracking range. It is further limited by C1. For most applications, $\Delta f_c < \Delta f/2$. For $R_D = 470k\Omega$, the approximate minimum value of C_D can be determined by:

$$C_D(\mu F) \geq 16/\text{capture range in Hz}$$

With values of C_D that are too small, chatter can be observed on the lock-detect output as an incoming signal frequency approaches the capture bandwidth. Excessively large values of C_D will slow the response time of the lock-detect output.

Tone Detection

Figure 4 shows the generalized circuit connection for tone detection. The logic outputs, Q and $\overline{Q}$ at pins 5 and 6 are normally at "high" and "low" logic states, respectively. When a tone is present within the detection band of the PLL, the logic state at these outputs becomes reversed for the duration of the input tone. Each logic output can sink 5mA of load current.

Both logic outputs at pins 5 and 6 are open-collector type stages, and require external pull-up resistors R_{L1} and R_{L2} as shown in Figure 4.

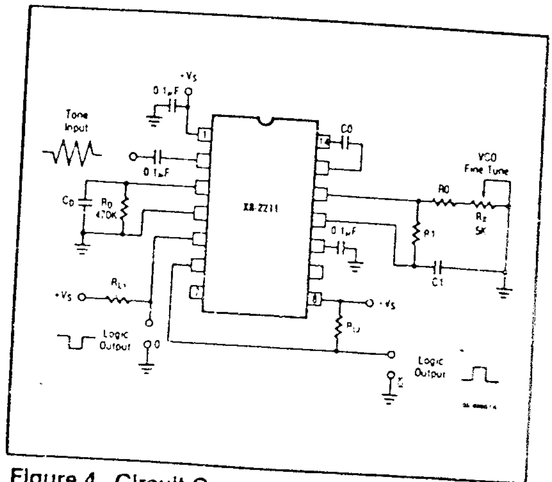


Figure 4. Circuit Connection for Tone Detection

With reference to Figures 1 and 4, the function of the external circuit components can be explained as follows: R0 and C0 set VCO center frequency, R1 sets the detection bandwidth, C1 sets the lowpass-loop filter time constant and the loop damping factor, and R_{L1} and R_{L2} are the respective pull-up resistors for the Q and $\overline{Q}$ logic outputs.

Design Instructions

The circuit of Figure 4 can be optimized for any tone-detection application by the choice of five key circuit components: R0, R1, C0, C1, and C_D. For a given input tone frequency, f_S, these parameters are calculated as follows:

1. Choose R0 to be in the range of 15kΩ to 100kΩ. This choice is arbitrary.
2. Calculate C0 to set center frequency, f_0 equal to f_S: $C0 = 1/R0f_S$.
3. Calculate R1 to set bandwidth $\pm\Delta f$ (see Design Equation No. 5): $R1 = R0(f_0/\Delta f)$

Note: The total detection bandwidth covers the frequency range of $f_0 \pm \Delta f$.

Courtesy of Ratheon Company.

4. Calculate value of C1 for a given loop damping factor:

$$C1 = C0/16\zeta^2$$

Normally $\zeta \sim 1/2$ is optimum for most tone-detector applications, giving C1 = 0.25 C0.

Increasing C1 improves the out-of-band signal rejection, but increases the PLL capture time.

5. Calculate value of filter capacitor C_D. To avoid chatter at the logic output, with R_D = 470kΩ, C_D must be:

$$C_D(\mu F) \geq (16/\text{capture range in Hz})$$

Increasing C_D slows the logic output response time.

Design Examples

Tone detector with a detection band of 1kHz ±20Hz:

Step 1: Choose R0 = 20kΩ (18kΩ in series with 5kΩ potentiometer).

Step 2: Choose C0 for f_0 = 1kHz: C0 = 0.05μF.

Step 3: Calculate R1: R1 = (R0) (1000/20) = 1MΩ.

Step 4: Calculate C1: for ζ = 1/2, C1 = 0.25μF, C0 = 0.013μF.

Step 5: Calculate C_D: C_D = 16/38 = 0.42μF.

Step 6: Fine tune the center frequency with the 5kΩ potentiometer, R_X.

Linear FM Detection

The XR-2211 can be used as a linear FM detector for a wide range of analog communications and telemetry applications. The recommended circuit connection for the application is shown in Figure 5. The demodulated output is taken from the loop phase detector output (pin 11), through a post detection filter made up of R_F and C_F, and an external buffer amplifier. This buffer amplifier is necessary because of the high impedance output at pin 11. Normally, a non-inverting unity gain op amp can be used as a buffer amplifier, as shown in Figure 5.

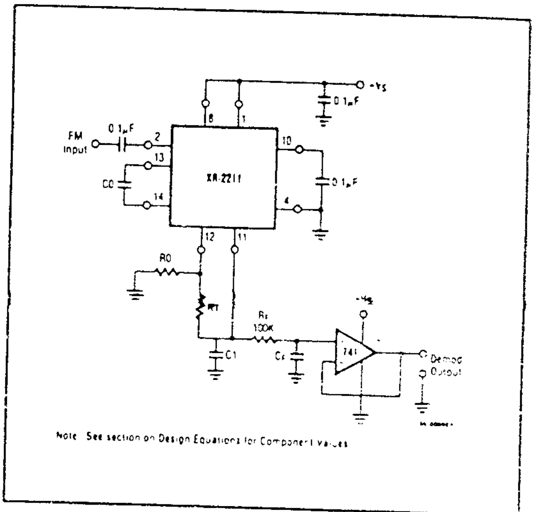


Figure 5. Linear FM Detector Using XR-2211 and an External Op Amp

The FM detector gain, i.e., the output voltage change per unit of FM deviation, can be given as:

$$V_{OUT} = R1\ V_R/100\ R0 \text{ Volts/\% deviation}$$

where V_R is the internal reference voltage. For the choice of external components R1, R0, C_D, C1 and C_F, see the section on Design Equations.

Courtesy of Ratheon Company.

ignal-to-noise ratio of nine db in favor of single-sideband ransmission. Thus, single-sideband transmission offers he equivalent of increasing the original carrier power eight times, and requires only half the bandwidth required by the a-m system.

The frequency-modulation (f-m) system is also used in power-line carrier work. In this system the amplitude or ntensity of the transmitted signal is constant and the frequency varies above and below a reference frequency in accordance with the intelligence being transmitted.

The deviation ratio, defined as the ratio of the maximum departure of the frequency from the reference value to the maximum frequency contained in the modulating signal, is a measure of the gain in signal-to-noise ratio of an f-m system over an a-m system of the same power. The f-m system provides marked increases in signal-to-noise ratio as the deviation ratio is increased. However, the minimum bandwidth required by frequency modulation is the same as that for a-m transmission of the same intelligence, and if a deviation ratio large enough to give a worthwhile increase in signal-to-noise ratio is used, the a-m bandwidth must be exceeded.

The frequency-shift system is a special form of frequency modulation that is used for telegraphic functions such as telemetering. In this system two closely-spaced frequencies are used. A continuous carrier wave of constant amplitude is shifted back and forth between the two frequencies, one frequency denoting a "mark" and one a "space" in the transmission of the impulses. By using highly stable crystal oscillators for the transmitted frequencies, and correspondingly stable and highly selective circuits in the receivers, it is possible to place the mark and space frequencies within 0.06 per cent of each other in the carrier spectrum. Even with this spacing, the equivalent f-m deviation ratio with the slow-speed keying required by practical impulse-telemetering systems is extremely high, with the result that a properly-designed frequency-shift system can provide substantial gains in signal-to-noise ratio with a small transmitted bandwidth.

III. PROPAGATION OF CARRIER ON TRANSMISSION LINES

19. Propagation Between Two Phase Conductors

Practically all textbooks on transmission give the classical solution for steady-state voltage and current at any point along a two-wire line[12,13]. This solution is approximately valid for carrier propagation between two phase conductors of a transposed three-phase power line, because transpositions tend to nullify the effect of the presence of the third conductor. The solution is based on the assumption that the line is composed of an infinite number of resistors and inductors in series, with an infinite number of capacitors and resistors shunting the line at equally-spaced points. This solution can be written as follows:

$$E_s = \left(\frac{E_r + I_r Z_c}{2}\right)\epsilon^{(\alpha+j\beta)l} + \left(\frac{E_r - I_r Z_c}{2}\right)\epsilon^{-(\alpha+j\beta)l} \quad (1a)$$

$$I_s = \left(\frac{I_r + \dfrac{E_r}{Z_c}}{2}\right)\epsilon^{(\alpha+j\beta)l} + \left(\frac{I_r - \dfrac{E_r}{Z_c}}{2}\right)\epsilon^{-(\alpha+j\beta)l}, \quad (1b)$$

in which E_s and I_s are the sending end voltage and current, respectively.

E_r and I_r are the receiving end voltage and current, respectively.

Z_c is the characteristic impedance as defined in the next paragraph.

$\alpha + j\beta$ is the propagation constant, to be defined later.

l is the distance from the *receiving* end, in the units of length used to define $\alpha + j\beta$.

20. Characteristic Impedance

Equations (1a) and (1b) show that when a voltage is applied to the sending end of the line, the voltage at any point on the line actually consists of two voltages, one a voltage traveling from the sending end of the line toward the receiving end, the other traveling from the receiving end back to the sending end. The former will be designated as E^+, the latter as E^-. Each of these voltages is accompanied by a corresponding current, I^+ and I^-, respectively. The ratio of either voltage to its corresponding current at any point in the line is a constant Z_c which is independent of the line length but is a function of the series resistance, the series inductance, the shunt conductance, and the shunt capacitance of the line per unit of length. This constant is the characteristic impedance of the line and can be expressed as

$$\frac{E^+}{I^+} = -\frac{E^-}{I^-} = Z_c = \sqrt{\frac{R + j\omega L}{G + j\omega C}} \quad (2)$$

where R = resistance in ohms per unit length.

L = inductance in henrys per unit length.

G = shunt conductance in mhos per unit length.

C = shunt capacitance in farads per unit length.

and $\omega = 2\pi f$ where f is the frequency in cycles per second.

In actual practice at high frequencies, such as those used in carrier transmission, the quantities $j\omega L$ and $j\omega C$ are so large by comparison with R and G that the latter can be neglected and the characteristic impedance expressed simply as

$$Z_c = \sqrt{\frac{L}{C}} \quad (3)$$

or by applying conventional formulas for L and C as

$$Z_c = 276 \log_{10}\frac{2D}{d} \quad (4)$$

where D is the distance between conductors and d is their diameter in the same units. Ordinary high-voltage transmission lines show characteristic impedances of 600 to 900 ohms between any pair of phase wires. Table 3 of Chap. 9 gives line-to-neutral surge impedances of a number of typical lines. The single-phase surge impedances are twice the values shown in this table.

21. Propagation Constant

Further study of the solution for a two-wire line shows that the phase and magnitudes of the voltage and current traveling toward the receiving end change as they progress along the line. The forward voltage and current at any point can be expressed as

$$E^+ = E_1^+ \epsilon^{(\alpha+j\beta)\Delta l} \quad (5a)$$

$$I^+ = I_1^+ \epsilon^{(\alpha+j\beta)\Delta l} \quad (5b)$$

where E_1^+ and I_1^+ are the values at some intermediate point on the line at a distance Δl toward the receiving end. Likewise, the voltage and current traveling in the opposite direction, E^- and I^-, change as they progress along the line, for

$$E^- = E_1^- \epsilon^{-(\alpha+j\beta)\Delta l} \quad (6a)$$

$$I^- = I_1^- \epsilon^{-(\alpha+j\beta)\Delta l} \quad (6b)$$

The quantity $\alpha+j\beta$ is the propagation constant of the line, which can be expressed as

$$\alpha+j\beta = \sqrt{ZY} = \sqrt{(R+j\omega L)(G+j\omega C)}. \quad (7)$$

The real part of $\alpha+j\beta$ is an exponent that expresses the reduction of the amplitudes of the forward and reverse voltages and currents as they appear at various points along their respective directions of travel. The imaginary part expresses the phase shift of the voltages and currents that results from the finite time required for the waves to travel from one point to another on the line.

22. Standing Waves

The forward and reverse voltages (and currents) aid and oppose each other at various points along the line, depending upon their respective phase positions. The total voltage and the total current therefore exhibit maxima and minima at equally spaced points separated by a distance that is a function of the frequency, giving rise to the phenomenon of standing waves. The magnitudes of the maxima and minima are a function of the amount of energy reflected from the receiving end of the line.

Standing waves increase the losses in a line as compared with the losses obtained without reflection or standing waves. They also result in increased radiation of energy from the line and other usually undesirable effects.

23. Attenuation

The attenuation of a proposed channel is of prime importance in carrier application, because it determines the fraction of the transmitted energy available at the receiving end to overcome noise and interfering voltages.

If, as in practical open-wire lines at carrier frequencies, the shunt conductance G is negligible and R is small compared to $j\omega L$, the real part of the propagation constant (Eq. 7) can be expressed as

$$\alpha = \frac{R}{2Z_c} \text{ nepers per unit of length} \quad (8a)$$

or

$$\alpha = \frac{4.34R}{Z_c} \text{ decibels per unit of length.} \quad (8b)$$

The resistance R is the resistance of the conductors per unit of length at the frequency in question. Calculation of R is difficult for the usual transmission line using stranded conductors, because common skin effect formulas apply accurately only to round conductors. Formulas for stranded conductors have been developed, and these give good results for unweathered surfaces and straight parallel strands, but are subject to errors depending on the condition of the conductor surface and the twisting of the strands in an actual line.

Most of the literature on power-line-carrier transmission reports measured values of attenuation in excess of figures calculated from theoretical considerations. The differences in these cases appear too great to be accounted for by expected errors in the determination of skin effect. For this reason, it is the usual practice in power-line-carrier application to use attenuation figures based on measurements on actual lines, rather than calculated figures. A table of approximate attenuation figures is given in a later section of this chapter.

24. Line Input Impedance

The reverse voltage and current expressed by the second terms of Eqs. (1a) and (1b) result from reflection of the forward voltage and current at the receiving end of the line. Equation (1) shows that if the line is terminated at the receiving end in an impedance equal to its characteristic impedance, Z_c, so that $\frac{E_r}{I_r} = Z_c$, there is no reverse voltage or current; i.e., no reflection at the receiver terminal. Under these conditions the input impedance Z_i at the sending end of the line is the surge impedance Z_c, and the ratio of total voltage to total current everywhere along the line is equal to Z_c. Also, if the line is sufficiently long, the second terms of Eqs. (1a) and (1b) are at the sending end negligible in magnitude by comparison with the first terms, even though the line is not terminated in Z_c. In this case also the input impedance is Z_c. This latter condition is often approached in practical carrier applications on isolated untapped lines. Carrier transmitters and receivers do not ordinarily provide a termination equal to the surge impedance of a line, but most carrier channels are sufficiently long that the input impedance of an isolated line is for practical purposes the characteristic impedance.

A special case that must be considered is that of short tap or spur lines that bridge a line over which carrier energy is to be transmitted. The input impedance of such a line may be extremely low under certain conditions and may constitute practically a short circuit across the carrier channel.

Consider, for example, the case of a low-loss line, open-circuited at the receiving end, one quarter wavelength long at a certain frequency. Voltage of this frequency applied to the input terminals, upon arriving at the receiving end, is reflected toward the source unchanged in magnitude and polarity, and has travelled a total of one-half wavelength by the time it reaches the source. It is exactly 180 degrees out of phase with the voltage being impressed at that instant and practically neutralizes it. It is impossible, therefore, to establish an appreciable voltage across the input terminals of a low-loss quarter wavelength line, because the reflected voltage always opposes any voltage that may be impressed. Such a line therefore appears as practically a short circuit at the particular frequency at which it is a quarter wavelength long.

The same phase relations apply for a line that is any odd multiple of a quarter wavelength long. The greater the number of quarter wavelengths, however, the greater the total attenuation of the voltage along the path from the terest is the case of line-to-ground coupling on long or properly terminated lines, in which reverse voltage and currents can be neglected. The equations are simplified in this case, and with carrier voltage E_S applied between phase one and ground, as shown in Fig. 13, the voltages to

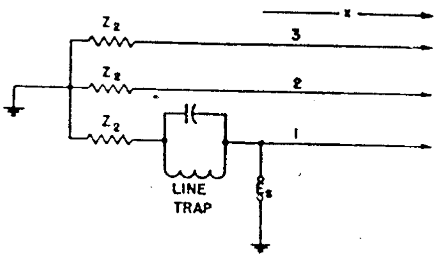


Fig. 13—Configuration assumed in discussion of propagation on ground return circuits. The impedances Z_2 may be lumped impedances or may be a continuation of the transmission line.

ground on phases 1, 2, and 3 at a distance x from the transmitting point are:

$$E_1 = \frac{E_s}{1+2Z'}\epsilon^{-k_0x} + \frac{2Z'}{1+2Z'}E_s\epsilon^{-kx}. \quad (9a)$$

$$E_2 = \frac{E_s}{1+2Z'}\epsilon^{-k_0x} - \frac{Z'}{1+2Z'}E_s\epsilon^{-kx}. \quad (9b)$$

$$E_3 = E_2. \quad (9c)$$

and the corresponding currents are:

$$I_1 = \frac{1}{Z_0}\frac{E_s}{1+2Z'}\epsilon^{-k_0x} + \frac{2Z'}{Z(1+2Z')}E_s\epsilon^{-kx} \quad (10a)$$

$$I_2 = \frac{1}{Z_0}\frac{E_s}{1+2Z'}\epsilon^{-k_0x} - \frac{1}{Z}\frac{Z'}{1+2Z'}E_s\epsilon^{-kx} \quad (10b)$$

$$I_3 = I_2 \quad (10c)$$

In which

k_0 = zero-sequence propagation constant (propagation constant for voltage applied to all three phases in parallel, with ground return).

Z_0 = zero-sequence characteristic impedance (characteristic impedance of all three phases in parallel, with ground return).

k = Positive- or negative-sequence propagation constant (propagation constant for a three-phase carrier frequency wave; i.e., square root of the product of line-to-neutral impedance and line-to-neutral admittance.)

Z = Positive- or negative-sequence characteristic impedance, line to neutral.

$$Z' = \frac{Z}{Z_0}\frac{Z_0+Z_2}{Z+Z_2}$$

Z_2 = Load impedance (to neutral) on phases 2 and 3 at coupling point. See Fig. 13.

The first term in each of these equations is a zero-sequence term. The attenuation of the zero-sequence terms is high on lines without ground wires, because of the high resistivity of the ground return path. These terms become negligible on long lines in comparison with the positive- and negative-sequence terms at a certain distance from the coupling point, and propagation takes place almost entirely between the coupled phase and the other two. The return current divides equally between the latter.

It has been noted that the attenuation per unit of distance is greater on short line-to-ground channels than on long line-to-ground channels[8]. Equations (9) and (10) provide at least a partial explanation of these results.

At the receiving point the current in the two uncoupled phases causes a loss of energy in the terminating impedances of these phases beyond the receiving point. This loss and the corresponding loss in the terminating impedances on the opposite side of the transmitting point account for the extra attenuation noticed in long line-to-ground channels as compared with phase-to-phase channels, according to Chevallier's results.

26. Characteristic Impedance of Ground-Return Circuits

The characteristic impedance of a circuit consisting of a single conductor with ground return is

$$Z_c = 138\log_{10}\frac{2h}{r} \quad (11)$$

where h is the height of the conductor above ground and r is its radius in the same units. Typical values for phase-to-ground carrier channels range from 400 to 600 ohms. The characteristic impedance of a transmission-line conductor with ground return is not greatly affected by the presence of the other conductors.

IV. NOISE VOLTAGE ON TRANSMISSION LINES

Since signal-to-noise ratio is the main criterion of the performance of a carrier transmission system, the noise level present at the receiving end of a carrier channel is equally as important to successful operation as the attenuation of the transmission path. The most important noise in a carrier system is that which originates in the power system itself; atmospheric noise is negligible, except that caused by nearby lightning strokes. The normal noise in a transmission system is the result of the presence of innumerable small arcs in dirty or defective insulators, poor joints, and the like. This condition is aggravated by wet weather, and is sometimes accompanied by corona discharge during such periods, with the result that noise usually increases to several times its normal amount. Noise also varies with the time of day under good weather conditions. Superimposed upon this normal or steady noise is the noise caused by switch operations, faults, etc.

27. Types of Noise[18,21,22]

Noise from whatever cause can be classified under two general headings: random noise and impulse noise. Random noise has a continuous frequency spectrum, containing all frequencies in equal amounts. At the output of a receiver it produces a steady hissing or rushing sound. The rms amplitude of this type of noise at the output of a receiver is proportional to the square root of the bandwidth of the receiver; i.e., the noise power is proportional to the bandwidth. The average and peak amplitudes are also proportional to the square root of the bandwidth.

Impulse noise is of far greater importance in carrier systems. It is produced when discrete, well-separated pulses exist at the input terminals of a receiver. If the pulses are irregular, the frequency spectrum is continuous and depends only slightly upon frequency. If the impulses are uniform and regularly spaced, the spectrum contains discrete frequency components separated by a frequency corresponding to the repetition rate. Power-line noise is essentially a combination of these two types of impulses, since basic repetition rates of 60 and 120 cycles are discernible along with random discrete pulses, all of irregular amplitude.

28. Response of a Receiver to Impulse Noise[18]

When a sharp impulse is applied to the input terminals of a receiver, the signal at the detector input is a damped oscillation having the natural resonant frequency of the preceding tuned circuits. The envelope of this oscillation, which represents the output of the receiver after detection, rises to a peak value at a certain time and then decays to zero. The greater the number of tuned circuits and the greater their Q, the more slowly the envelope of the oscillation rises to a maximum and the lower its peak value; i.e., the peak output is proportional to receiver bandwidth. However, the total area of the output signal envelope, and hence the average output, are independent of these factors. In practice, if the impulses are sharp and well separated, the rms output is independent of the shape of the impulses and is dependent only upon their areas, the gain of the receiver, and the square root of the receiver bandwidth. If the impulses are not well separated, so that in a receiver of a given bandwidth the resulting wave trains overlap, the response of the receiver simulates that for a combination of random noise and impulse noise. In some applications the peak output is of major importance, whereas in others the average or the rms output is the critical factor.

Thus, in specifying the characteristics of noise on transmission lines, it is necessary to state not only the relative amounts of random and impulse noise in a given bandwidth, but also the peak values of the impulses (or their statistical distribution) and the duration and spacing of the individual impulses. In order to evaluate the effect of this noise upon a given receiving system, it is necessary to know the receiver bandwidth and gain and the particular application involved. The number of tuned circuits and their Q determine the receiver bandwidth at a given frequency.

29. Measurement of Carrier-Frequency Noise on Power Lines

The accurate measurement of all the characteristics of carrier-frequency noise on transmission lines requires a considerable amount of equipment, including a noise meter of definite bandwidth, capable of measuring peak impulse amplitudes, and an oscilloscope to indicate the spacing and the duration of the impulses. In order to be significant, readings should be taken over a period of time sufficient to include both fair and rainy weather. As a result, actual test data of this type on carrier-frequency noise is extremely limited, and no typical figures for "quiet" or "noisy" lines have been established.

Figure 14 shows the results of one set of measurements made during fair weather on a 132-kv line that can be classified only as "relatively noisy" for carrier. These measurements were made with a Stoddart Type URM-6 noise meter. The curves show that the peak values of noise on this line are far in excess of the average values, indicating that average-reading instruments do not give a true indication of the probable interfering effects of noise for all applications. A graphic record of quasi-peak values* over an extended time, including periods of rainy weather, gave the curves of Fig. 15, which indicate a relatively great increase in the noise under some conditions, with maximum quasi-peak values exceeding 100,000 microvolts for approximately 3 percent of the time.

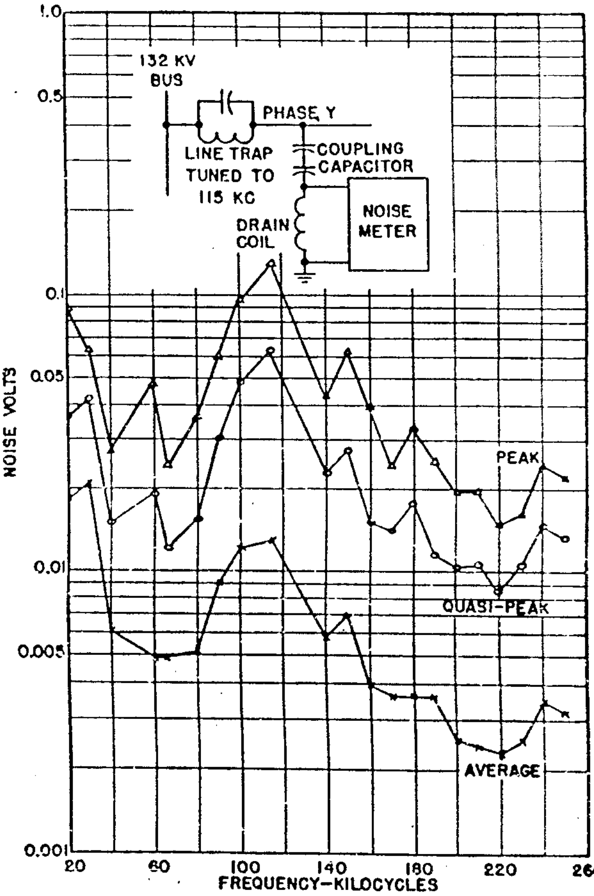


Fig. 14—Noise voltages as a function of frequency on a relatively noisy 132-kv line. The rise in the vicinity of 115 kc is probably accounted for by the presence of the line trap.

*Quasi-peak readings are based on a fast detector output circuit charging time and a slow discharging time, and hence are a function of the peak amplitudes as well as the spacing of the impulses. The times are chosen so that the quasi-peak readings are approximately proportional to the interfering properties of impulse noise in aural reception of a-m signals.

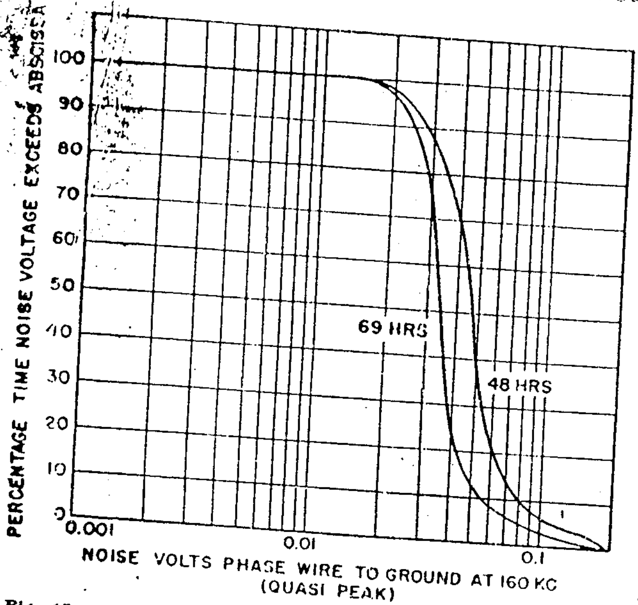


Fig. 15—Quasi-peak noise voltages at 160 kc taken over two extended periods on the same 132-kv line. Both periods included rainy as well as fair weather.

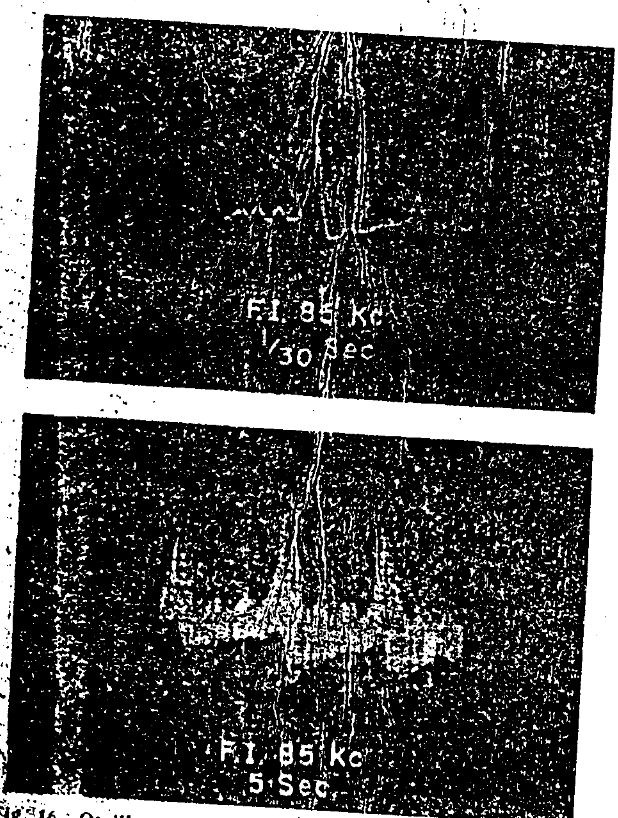


Fig. 16—Oscilloscope patterns of carrier noise at 85 kc with 1/30 second and 5 second exposures.

Cathode-ray oscilloscope patterns of the output of the detector of the noise meter used are shown in Fig. 16. These indicate irregularity in the pulses, even with 1/30-second exposures. The five-second exposure shows that although the basic system frequency is present in the amplitudes, the pulses occur almost at random throughout the cycle. It must be remembered that these pulses have undergone a smoothing and rounding effect as a result of the action of the tuned circuits in the noise meter and that the actual pulses at the input of the meter were sharper and of shorter duration than those shown in the photographs.

V. COUPLING AND TUNING EQUIPMENT AND CIRCUITS

In the early days of power-line carrier it was universal practice to couple the carrier equipment to the power line by a method known as antenna coupling. In this method the carrier equipment was connected to an isolated conductor, several spans long, on the same tower with the circuit to which coupling was to be effected. Eventually it was realized that the energy which found its way into the power line was transferred mainly through the capacitance between the antenna and the line, and this led to the development of compact capacitor units for coupling purposes. Such coupling capacitors are safer, easier to install, and are a more efficient coupling means than antennas. They also have the advantage that they can be used simultaneously in conjunction with potential devices to supply a voltage proportional to line voltage for the operation of protective relays and indicating instruments.

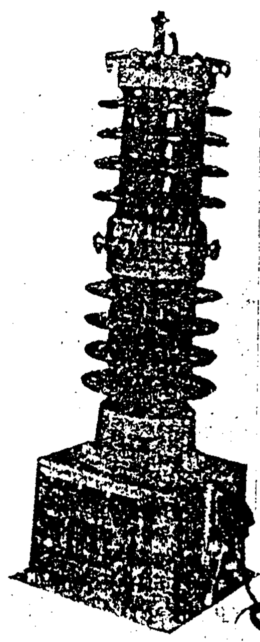


Fig. 17—Typical carrier coupling capacitor. This unit is rated at 115 kv and has a total capacitance of .00187 mfd.