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To Design and Build a Portable, Miniaturised, Multichannel FM Transmitter

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B. Eng. Electronic Engineering (LM070)

Submitted in part requirement for final year project to

University of Limerick, Limerick, Ireland

28/04/98

Abstract

The aim of the project is to develop a Miniaturised low power FM Transmitter to be used in specialised applications such as a hearing aid for a tour guiding system and room monitoring (such as a baby listening device). The overall module should be miniature to enable portability. Frequency modulation has several advantages over the system of amplitude modulation (AM) used in the alternate form of radio broadcasting. The most important of these advantages is that an FM system has greater freedom from interference and static. Various electrical disturbances, such as those caused by thunderstorms and car ignition systems, create amplitude modulated radio signals that are received as noise by AM receivers. A well-designed FM receiver is not sensitive to such disturbances when it is tuned to an FM signal of sufficient strength. Also, the signal-to-noise ratio in an FM system is much higher than that of an AM system. FM broadcasting stations can be operated in the very-high-frequency bands at which AM interference is frequently severe; commercial FM radio stations are assigned frequencies between 88 and 108 MHz and will be the intended frequency range of transmission.

The main report will reflect on 4 issues, background to frequency modulation, electronics component characteristics, basic transmitter building blocks and finally an analysis of the finished design as regards construction and performance.

Declaration

This report is presented in partial fulfilment of the requirements for the degree of Bachelor of Engineering.

It is entirely my own work and has not been submitted to any other University or higher education institution, or for any other academic award in this University. Where use has been made of the work of other people it has been fully acknowledged and fully referenced.

Signature: ______________________________

Francis Mc Swiggan

24th April 1998

Acknowledgements

I would very much like to gratefully extend my sincere thanks to all the people who gave generously their time, takes one and all. Especially my supervisor Dr. Máirtín Ó Droma for the guidance he showed me right through every stage of the project, from initial conception to final design and construction.

Richard Conway for the loan of Electronic communication Techniques at the initial stage of the project, Elfed Lewis for kindly granting me access to his Pspice lab for initial simulation work. The technicians Jimmy Kelly, James Keane, John Maurice and many more of the lads in stores who helped me with a lot of the tedious leg work involved in giving life to a design and implementation project.

To the various chancers in my course who helped me stay awake during many a long nights work. The “podger” , “Richie of the hellen” and many more too numerous to mention, thanx Lads.

Dedication

To the person who supported me through 4 years of College.

Thanks Mam.

And to the rest of my family.

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1 Frequency Modulation Background

1.1 Introduction

The comparatively low cost of equipment for an FM broadcasting station, resulted in rapid growth in the years following World War II. Within three years after the close of the war, 600 licensed FM stations were broadcasting in the United States and by the end of the 1980s there were over 4,000. Similar trends have occurred in Britain and other countries. Because of crowding in the AM broadcast band and the inability of standard AM receivers to eliminate noise, the tonal fidelity of standard stations is purposely limited. FM does not have these drawbacks and therefore can be used to transmit music, reproducing the original performance with a degree of fidelity that cannot be reached on AM bands. FM stereophonic broadcasting has drawn increasing numbers of listeners to popular as well as classical music, so that commercial FM stations draw higher audience ratings than AM stations.

The integrated chip has also played its part in the wide proliferation of FM receivers, as circuits got smaller it became easier to make a modular electronic device called the "Walkman", which enables the portability of a tape player and an AM/FM radio receiver. This has resulted in the portability of a miniature FM receiver, which is carried by most people when travelling on long trips.

1.2 Technical Background

Frequency	*Designation*	*Abbreviation*	*Wavelength*
3 - 30 kHz	Very Low frequency	VLF	100,000-10,000 m
30 - 300 kHz	Low frequency	LF	10,000 - 1,000 m
300- 3,000 kHz	Medium frequency	MF	1,000 - 100 m
30 - 30MHz	High frequency	HF	100 - 10 m
30 - 300 MHz	Very High frequency	VHF	10 - 1m
300- 3,000 MHz	Ultra-high frequency	UHF	1m - 10m
3 - 30 GHz	Super-high frequency	SHF	10cm - 1cm
30 - 300 GHz	Extremely-high frequency	EHF	1cm - 1mm

The main frequencies of interest are from 88MHz to 108MHz with wavelengths between 3.4 and 2.77 meters respectively.

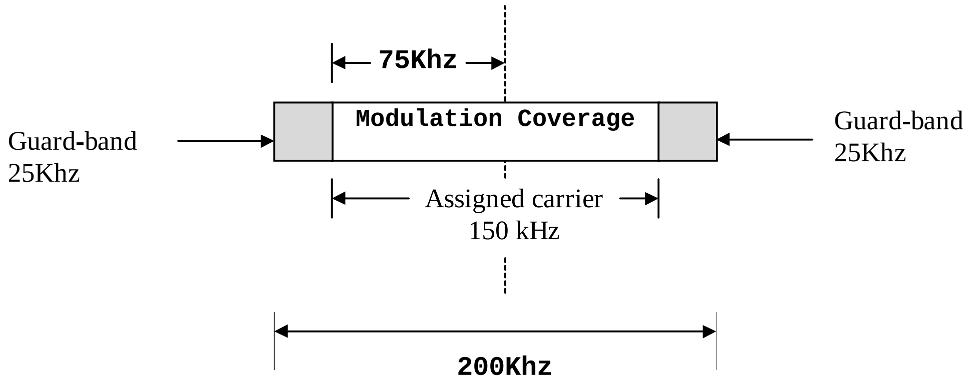


Figure 1-1

With a bandwidth of 200Khz for one station, up to 100 stations can be fitted between 88 & 108Mhz. Station 201 to 300 denote the stations, from 88.1Mhz to 107.9Mhz.

Station 201 to 220 (88Mhz to 91.2) are for non-commercial stations (educational) which could be a good area to transmit in, but in recent years the band from 88MHz to 103Mhz has been filled by a lot of commercial channels, making the lower frequencies very congested indeed.

1.2.1 Radio Frequency and Wavelength Ranges

Radio waves have a wide range of applications, including communication during emergency rescues (transistor and short-wave radios), international broadcasts (satellites), and cooking food (microwaves). A radio wave is described by its wavelength (the distance from one crest to the next) or its frequency (the number of crests that move past a point in one second). Wavelengths of radio waves range from 100,000 m (270,000 ft) to 1 mm (.004 in). Frequencies range from 3 kilohertz to 300 Giga-hertz.

1.3 Fm theory

Angle and Amplitude Modulation are techniques used in Communication to transmit Data or Voice over a particular medium, whether it be over wire cable, fibre optic or air (the atmosphere). A wave that is proportional to the original baseband (a real time property, such as amplitude) information is used to vary the angle or amplitude of a higher frequency wave (the carrier).

$$\text{Carrier} = A \quad \text{Cos} \quad \Phi(t)$$

$$\phi(t) \quad = \quad 2\pi f_C\, t \quad + \quad \alpha$$

Where A is the amplitude of the carrier and $\phi(t)$ is the angle of the carrier, which constitutes the frequency (f_C) and the phase (α) of the carrier. Angle modulation varies the angle of the carrier by an amount proportional to the information signal. Angle modulation can be broken into 2 distinct categories, frequency modulation and phase modulation. Formal definitions are given below :

Phase Modulation *(PM)* : angle modulation in which the phase of a carrier is caused to depart from its reference value by an amount proportional to the modulating signal amplitude.

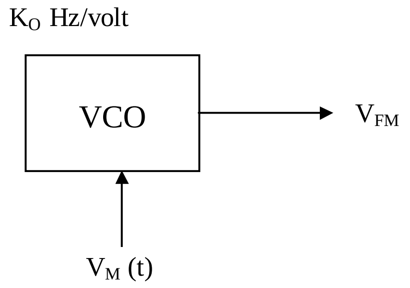
Frequency Modulation *(FM):* angle modulation in which the instantaneous frequency of a sine wave carrier is caused to depart from the carrier frequency by an amount proportional to the instantaneous value of the modulator or intelligence wave.

Phase modulation differs from Frequency modulation in one important way. Take a carrier of the form $A \, Cos(\omega_C t + \theta) = Re\{A.e^{j(\omega Ct + \theta)}\}$

Pm will have the carrier phasor in between the + and - excursions of the modulating signal. Fm modulation also has the carrier in the middle but the fact that when you integrate the modulating signal and put it through a phase modulator you get fm, and if the modulating wave were put through a differentiator before a frequency modulator you get a phase modulated wave. This may seem confusing at this point, but the above concept will be reinforced further in the sections to follow.

1.3.1 Derivation of the FM voltage equation

Consider a voltage controlled oscillator with a free running frequency of f_C , an independent voltage source with voltage $V_M(t)$ which causes the VCO to depart from f_C by an amount Δf, which is equal to the voltage of the independent source multiplied by the sensitivity of the VCO (K_O => such as the miller capacitance of a transistor). What is seen at the output of the VCO is a frequency modulated voltage. Now consider the independent voltage source as representing the amplitude of the baseband information .



$$V_{FM} = A \cos \theta(t) \qquad 1$$

$$f = f_c + \Delta f \qquad 2$$

$$\Delta f = K_o * V_m(t)$$

Above are the equations which govern the output of the VCO, f is the overall frequency of the frequency modulated output.

$$\omega = \frac{d\theta(t)}{dt} = 2\pi f \qquad 3$$

taking the angle θ(t) from equation 1 and differentiating it will give the angular velocity of the output and equate it to 2π times the effective frequency (f)

$$\frac{d\theta(t)}{dt} = 2\pi f_c + 2\pi\Delta f \qquad 4$$

$$d\theta(t) = 2\pi f_c\, dt + 2\pi\Delta f\, dt \qquad 5$$

multiply across both sides by the change in time (dt)

$$\theta(t) = 2\pi f_c \int dt + 2\pi K_o \int V_m(t)\, dt \qquad 6$$

$$V_m(t) = Vpk \quad \cos(2\pi f_m t) \qquad 7$$

$$\theta(t) = 2\pi f_c t + \frac{2\pi K_o}{2\pi f_m} Vpk \quad \sin(2\pi f_m t) \qquad 8$$

Substituting in the equation for the intelligence (baseband) voltage 7 into equation 6 and integrating gives equation 8 which is the angle of the frequency modulated wave of equation 1.

$$\theta(t) = 2\pi f_c t + \frac{K_o * Vpk}{fm} \sin(2\pi f_m t) \qquad 9$$

$$M_F = \frac{K_o * Vpk}{fm} \qquad 10$$

$$M_F = \frac{\Delta fc(pk)}{fm} \qquad 11$$

Tiding up equation 8, and setting the magnitude of the sine wave as M_F , the modulation index for frequency modulation.

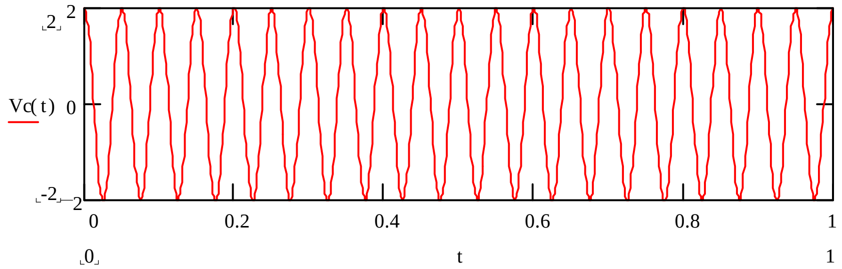
$$V_{FM} = A \cos \theta(t) = A \cos \left[2\pi f_c t + M_F \sin(2\pi f_M t)\right] \qquad 12$$

The above equation represents the standard equation for frequency modulation. The equation for the other form of angle modulation, phase modulation is rather similar but has a few subtle differences.

$$V_{PM} = A \cos \theta(t) = A \cos \left[2\pi f_c t + M_P \cos(2\pi f_M t)\right] \qquad 13$$

The difference is in the modulation Index and the phase of the varying angle inside the main brackets.

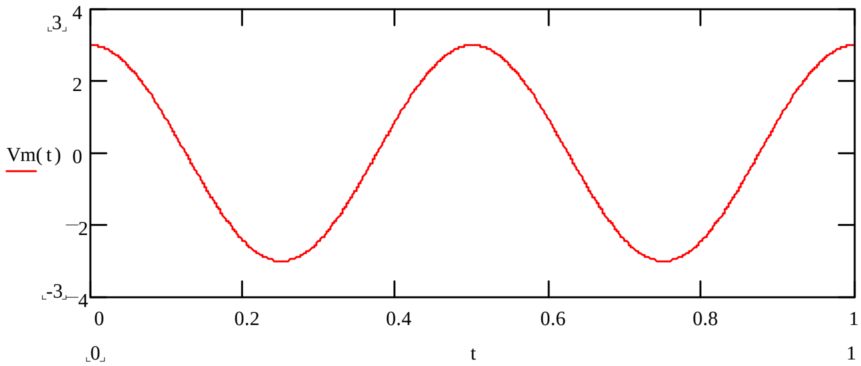
1.3.2 Angle modulation Graphs



Carrier Wave

Figure 1.3-1

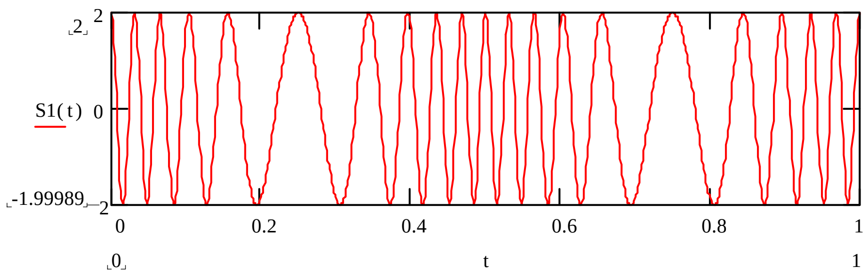
$$Vc(t) := A\cdot\cos(2\cdot\pi\cdot fc\cdot t)$$



Baseband Signal

Figure 1.3-1

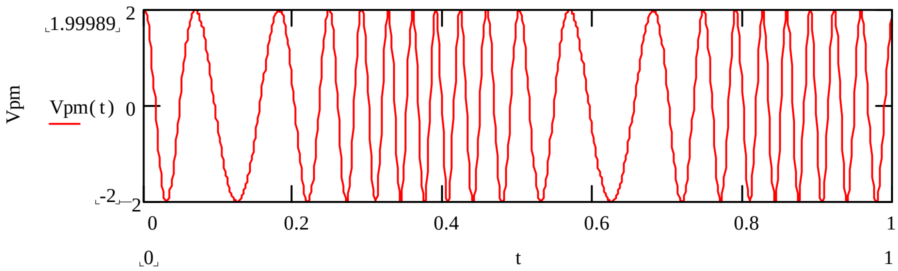
$$Vm(t) := (Vpk\cdot\cos(2\cdot\pi\cdot fm\cdot t))$$



Frequency Modulated Wave

Figure 1.3-3

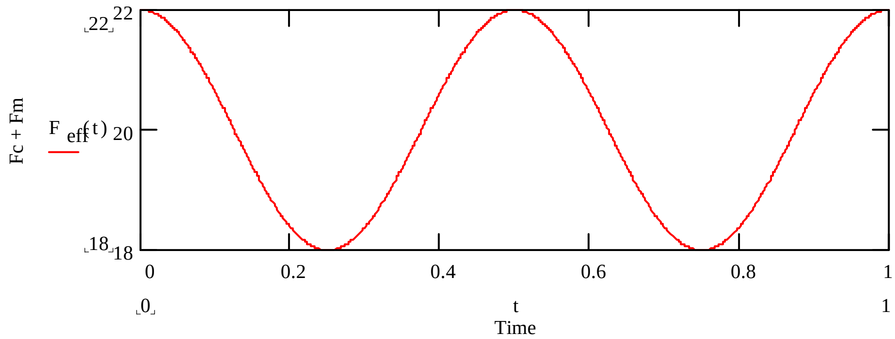
$$S(t) := A\cdot\cos\left(2\cdot\pi\cdot fc\cdot t + 2\cdot\pi\cdot Ko\cdot\int_0^t Vm(t)\,dt\right)$$



Phase Modulated Wave

Figure 1.3-4

$$Vpm(t) := A\cdot\cos(2\cdot\pi\cdot fc\cdot t + Kp\cdot Vm(t))$$



Frequency versus Time

Figure 1.3-5

$$F_{eff}(t) := fc + fm\cdot\cos\left(\frac{2\cdot\pi}{T}\cdot t\right)$$

1.3.3 Analysis of the above graphs

There are 5 significant graphs above, The carrier, the Baseband, FM signal, PM signal and the change of frequency over time. The carrier and baseband are there to show the relative scale, so a link between the carrier and Baseband can be seen.

For FM: the carrier's frequency is proportional to the baseband's amplitude, the carrier increases frequency proportional to the positive magnitude of the baseband and decreases frequency proportional to the negative magnitude of the baseband.

For PM: the carrier's frequency is proportional to the baseband's amplitude, the carrier increases frequency proportional to the positive rate of change of the baseband and decreases frequency proportional to the negative rate of change of the baseband. In other words when the baseband is a maximum or a minimum, there is Zero rate of change in the baseband, and the carrier's frequency is equal to the its free running value f_C.

In both systems the rate of modulation is equal to the frequency of modulation (baseband's frequency). The last graph shows the relationship between the frequency of FM versus Time, this relationship is used (following a limiter which makes sure the amplitude is a constant) by a discriminator at the receiver to extract the Baseband's Amplitude at the receiver, resulting in an amplitude modulated wave, the information is then demodulated using a simple diode detector. In common AM/FM receivers for an AM station to be demodulated, the limiter and discriminator can be by passed and the intermediate frequency signal can be fed straight to the diode detector.

1.3.4 Differences of Phase over Frequency modulation

The main difference is in the modulation index, PM uses a constant modulation index, whereas FM varies (Max frequency deviation over the instantaneous baseband frequency). Because of this the demodulation S/N ratio of PM is far better than FM.

The reason why PM is not used in the commercial frequencies is because of the fact that PM need a coherent local oscillator to demodulate the signal, this demands a phase lock loop, back in the early years the circuitry for a PLL couldn't be integrated and therefore FM, without the need for coherent demodulation was the first on the market. One of the advantages of FM over PM is that the FM VCO can produce high-index frequency modulation, whereas PM requires multipliers to produce high-index phase modulation. PM circuitry can be used today because of very large scale integration used in electronic chips, as stated before to get an FM signal from a phase modulator the baseband can be integrated, this is the modern approach taken in the development of high quality FM transmitters.

For miniaturisation and transmission in the commercial bandwidth to be aims for the transmitter, PM cannot be even considered, even though Narrow Band PM can be used to produce Wide band FM (Armstrong Method).

1.4 Technical terms associated with FM

Now that Fm has been established as a scheme of high quality baseband transmission, some of the general properties of FM will be looked at.

1.4.1 Capture Effect

Simply put means that if 2 stations or more are transmitting at near the same frequency FM has the ability t pick up the stronger signal and attenuated the unwanted signal pickup.

1.4.2 Modulation Index

$$M_F = \frac{\Delta fc(pk)}{fm} \qquad \text{(Was known as the modulation factor)}$$

Modulation Index is used in communications as a measure of the relative amount of information to carrier amplitude in the modulated signal. It is also used to determine the spectral power distribution of the modulated wave. This can be seen in conjunction with the Bessel function. The higher the modulation index the more side-bands are created and therefore the more bandwidth is needed to capture most of the baseband's information.

1.4.3 Deviation Ratio

The deviation can be quantified as the largest allowable modulation index.

$$D_R = \frac{\Delta fc(pk)}{fm(max)} = \frac{75KHz}{15KHz} = 5 \text{ radians}$$

For the commercial bandwidth the maximum carrier deviation is 75KHz. The human ear can pick up on frequencies from 20Hz to 20KHz, but frequencies above 15KHz can be ignored, so for commercial broadcasting (with a maximum baseband frequency of 15KHz) the deviation ratio is 5 radians.

1.4.4 Carrier Swing

The carrier swing is twice the instantaneous deviation from the carrier frequency.

$$F_{CS} = 2.\Delta F_C$$

The frequency swing in theory can be anything from 0Hz to 150KHz.

1.4.5 Percentage Modulation

The % modulation is a factor describing the ratio of instantaneous carrier deviation to the maximum carrier deviation.

$$\%\ \text{Modulation} = \frac{\Delta F_C}{\Delta F_C(pk)} \times 100$$

1.4.6 Carson's Rule

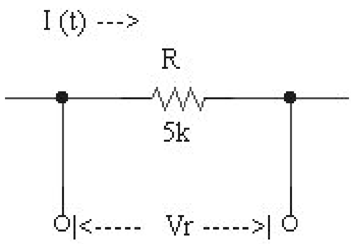
Carson's Rule gives an indication to the type of Bandwidth generated by an FM transmitter or the bandwidth needed by a receiver to recover the modulated signal. Carson's Rule states that the bandwidth in Hz is twice the sum of the maximum carrier frequency deviation and the instantaneous frequency of the baseband.

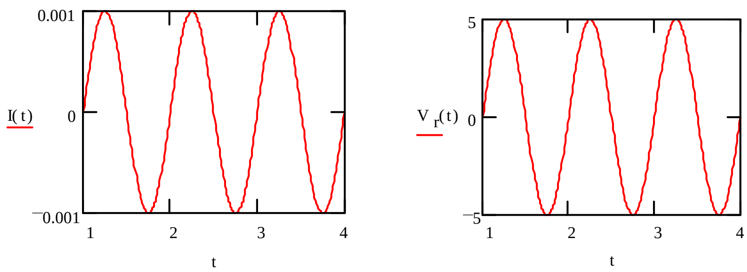
$$\text{Bandwidth} = 2\,(\Delta F_C\,(pk) + F_{M)}$$
$$= 2\,F_M\,(1 + M_F)$$

2 Electronic Components and their properties

2.1 Resistor

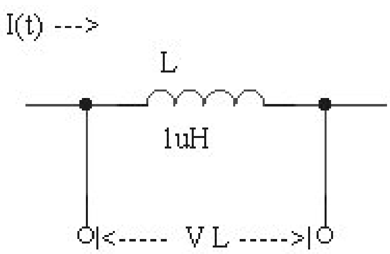
For a resistor the voltage dropped across it is proportional to the amount of current flowing on the resistor $V_R = I.R$,any current waveform through a resistor will produce the exact same voltage waveform across the resistor, although this seems trivial it is worth keeping it in mind, especially when it comes to dealing with other components such as inductors, capacitors and ordinary wire at high frequency.



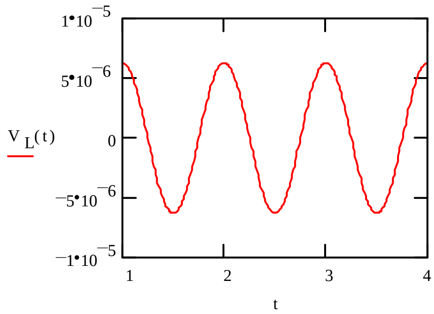


$$I(t) := 10^{-3} \cdot \sin(2 \cdot \pi \cdot t) \qquad V_r(t) := I(t) \cdot R$$

2.2 Inductor

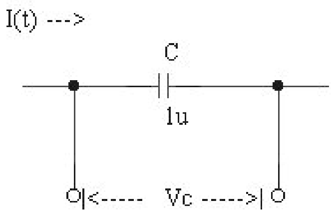


The voltage across an inductor Leads the current through it by 90°, this is due to the fact that the voltage across an inductor depends on the rate of change of current entering the inductor. The impedance of an inductor is $+j\omega L$ ($\omega = 2\pi f$), which reflects the fact that the voltage leads the current. This analysis is vital in working out the phase shift trough complicated LC networks.

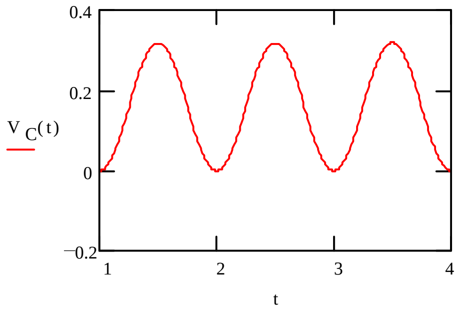


$$V_L(t) := L\cdot\frac{d}{dt}I(t)$$

2.3 Capacitor



The voltage across a capacitor lags the current through by 90^{o}, applying the same logic to the capacitor as was used for the inductor, the reason for this lag in voltage is that the voltage is proportional to the integral of current entering the capacitor. Looking at the above current plot the current will reach a maximum 90^{O} into the cycle, the voltage will reach a maximum when the area under the current's curve is added up this doesn't happen until 180^{O} into the currents cycle, giving a 90 degrees voltage lag. The Impedance of the capacitor can be found to be $-j\frac{1}{\omega C}$, which also takes into account of the capacitor's voltage lag.



$$V_C(t) := \frac{1}{C} \cdot \int_0^t I(t)\, dt$$

2.4 Resonant Circuits

In the last section the resistor, inductor & capacitor were looked at briefly from a voltage, current and impedance point of view. These components will be the basic building blocks used in any radio frequency section of any transmitter/receiver. What makes them important is there response at certain frequencies. At high low frequency the impedance of an inductor is small and the impedance of a capacitor is quite high. At high frequency the inductor's impedance becomes quite high and the capacitor's impedance drops. The resistor in theory maintains it's resistive impedance at low & high impedance. At a certain frequency the capacitor's impedance will equal that of an inductor, This is called the resonant frequency and can be calculated by letting the impedance of a capacitor to that of the inductor's and then solving for ω (angular velocity in radians per seconds) and then finding the resonant frequency Fc (it is normally represented as Fo, but in relation to FM it essentially represents the oscillator carrier frequency) in Hertz.

$$\omega_C = \frac{1}{\sqrt{LC}}$$

$$Fc = \frac{1}{2\pi\sqrt{LC}}$$

There are two configurations of RLC circuits, the series and parallel arrangements, which will now be looked at below.

2.4.1 Series resonant circuit

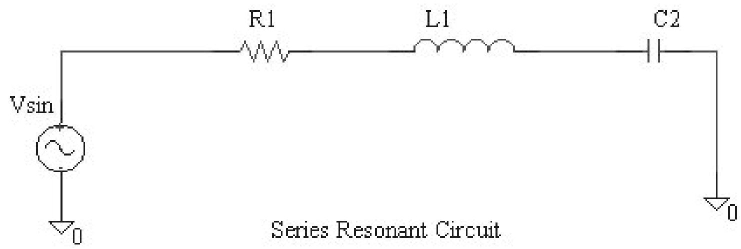


Figure 2.4-1

At low frequencies the capacitor impedance will dominate the overall impedance of the series circuit and the current is low. At high frequencies the inductor impedance will dominate and the current will also be low. But at the resonant frequency the complex impedance of the capacitor will cancel that of the inductor's and only the resistance of the resistor will remain effective, this is when the current through the circuit will be at a maximum.

$$Z(f) = r + j(2\pi f.L - 1/2\pi f.C) \text{ is a minimum at Fc}$$

2.4.2 Parallel resonant circuit

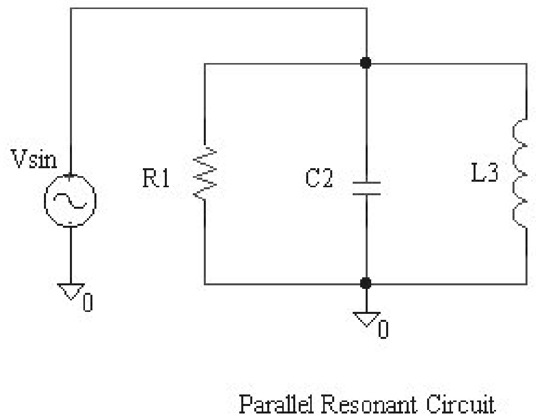


Figure 2.4-2

The parallel circuit above (known as an LC tank) takes the same advantage of the resonant frequency but this time the impedance will be at a maximum and the current will be at a minimum at F_C. This due to the fact that the minimum impedance in a parallel circuit dominates the overall impedance of the tank.

The impedance will be equal to (R // +jXL // -jXC)

$$Z = \frac{R}{1 + j\frac{R}{L}\left(\omega LC - \frac{1}{\omega}\right)}$$

now substituting $\omega_C{}^2 = \frac{1}{LC}$, cross multiply and bring the common ω_O out of the brackets to get the impedance as a function frequency.

$$Z(\omega) = \frac{R}{1 + j\left(\frac{R}{\omega_C L}\right)\left(\frac{\omega}{\omega_C} - \frac{\omega_C}{\omega}\right)}$$

now with the substitution and $\omega_O = 2\pi fo$, the parallel impedance at any frequency can be found. A factor called the Q factor can be introduced which is equal to $R/\omega_O L$.

$$Z(f) = \frac{R}{1 + jQ\left(\frac{f}{f_C} - \frac{f_C}{f}\right)}$$

at frequencies above resonance $f >> f_C$ the above equation evaluates to

$$Z(f) = -j\left(\frac{R.f_C}{Q}\right)\frac{1}{f}$$

Which is capacitive.

At frequencies above resonance $f << f_C$ the above equation evaluates to

$$Z(f) = +j\left(\frac{R}{Q.f_C}\right)f$$

Which is inductive impedance

At resonance the complex component under the line will be zero, yielding a real value of R which is purely resistive.

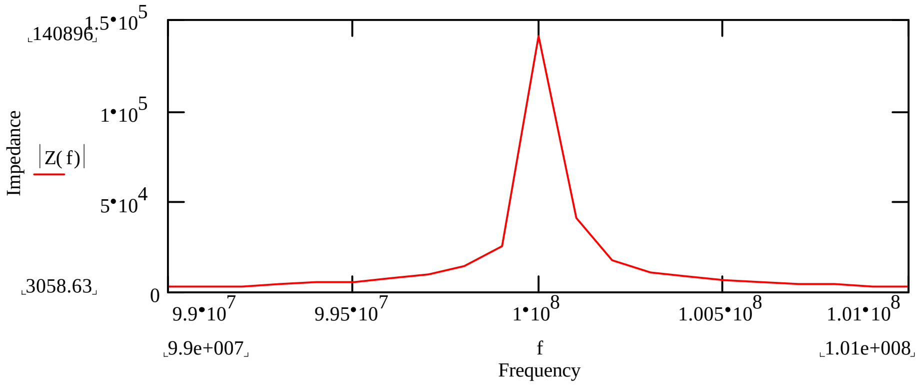


Figure 2.4-3

Impedance versus Magnitude

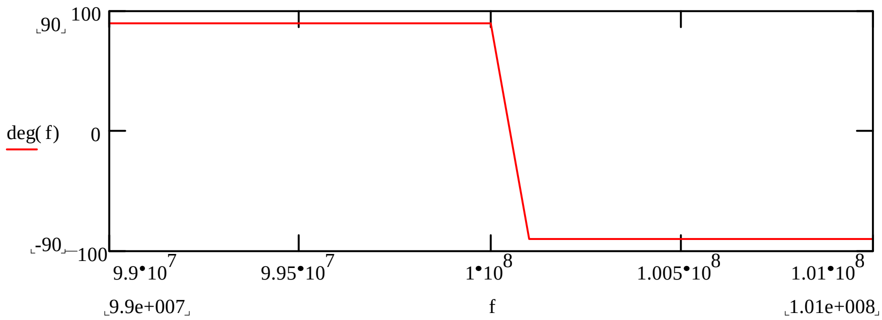


Figure 2.4-4

Phase Plot

Simulations carried out in Mathcad with values of C = 25.19pF and L = 0.1μH, the resonant (centre) frequency was found to be 100MHz. The Q has a part in finding the bandwidth, BW = F_C /Q, which was calculated to be 67KHz with a resistance R = 100KΩ.

The phase plot show's a phase of +90^O (inductive impedance) before the resonant frequency, 0^O (resistive) at resonant frequency, and -90^O (capacitive) above the resonant frequency.

2.5 The Q factor

Quality of the component has to be taken into account. The Q factor is a measure of the energy stored to that which is lost in the component due to its resistive elements at low or high frequencies. Inductors store energy in the magnetic field surrounding the device. Capacitors store energy in the dielectric between it's plates. The energy is stored in one half of an ac cycle and returned in the second half. Any energy lost in the cycle is associated with a dissipative resistance and this gives rise to the Quality factor Q. Q as stated before is the ratio of maximum energy stored to the amount lost per ac cycle. As shown in the previous section the Quality factor determines the 3db bandwidth of resonant circuits.

For a series RLC circuit at Fc

$$Q = \frac{2\pi f_C L}{R_{series}} \quad \text{or} \quad Q = \frac{1}{(2\pi f_C C)R_{series}}$$

For a parallel RLC circuit at Fc

$$Q = \frac{R_P}{2\pi f_C L}$$

In circuits where there is no R_{series} or $R_{parallel}$ (only an L and a C) the inherent resistive properties of the inductor (skin effect) and capacitor (dielectric permittivity) at high frequencies can be taken into account.

Conclusion : the higher the Q the less energy is dissipated .

2.6 High frequency response of discrete components

2.6.1 Wire

The resistance of a piece of wire decreases as the diameter of the wire increases, $R = \rho \frac{L}{A}$, where ρ is the resistivity, L the length of the wire and A is its cross sectional area. But beyond a particular frequency the resistance of the wire increases, strong magnetic fields are built up at the centre of the wire due to high frequency ,this force pushes the majority of the charge carriers (electrons) away from the centre and towards the outside of the wire. So now there is less available cross sectional area for the carriers have to move along the wire, therefore the resistance increases at high frequencies. This phenomenon is known as the "skin effect", when the magnetic field at the centre increases and local inductive reactance takes over.

Analysing the skin effect further, it is understood that AC current distributes itself across the cross sectional area of the wire in a parabolic shape, simply put means that the majority of the carrier lie in the outside, while few remain at the centre of the wire. The outside region where most of the electrons reside can be defined as the distance in from the outside where the number of electrons has dropped to $(2.7183)^{-1}$ = 36.8 % of the electrons on the outside.

2.6.2 Inductor

Since wire is the main ingredient of inductors and since the resistance of wire increases with increasing frequency, therefore the losses of an inductor will increase with increasing frequency (as characteristic resistance increases). The amount of loss for a given inductor through dissipation the inverse of the Q factor.

$$\text{Dissipation} = Q^{-1} = \frac{R_{series}}{2\pi f L}$$

Therefore, since R_{series} increases with frequency, therefore the Q factor will decrease with increasing frequency. Initially, the Q factor of the inductor increases at the same rate as the frequency changes and this continues as long as the series resistance remains at the DC value. Then, at some frequency that depends on the wire diameter and also on the manner of the windings, the Skin effect sets in and the series resistance starts to climb. However not at the same rate as the frequency does, and so the Q continues to rise, but not as steeply as before. As the frequency increases further, a stray capacitance begins to build up between adjacent turns. Along with the inductance a parallel resonant circuit is formed and the resulting resonant frequency causes the Q factor to start decreasing.

2.6.3 Capacitor

The resistive element in a capacitor at a high frequency is brought about by the material in between the plates of the capacitor, which inherently controls the permittivity and then also the conductive properties of the capacitor at high frequencies. The dissipation factor of the capacitor is also the inversely associated with the Q factor. The efficiency in capacitors at high frequencies are generally better than the inductor as regards the Q factor, but other considerations such as the added series inductance of the leads and the internal capacitor plates will greatly effect the efficiency of the capacitor. Good RF techniques are usually used to combat this by keeping the leads short when soldering a capacitor into a circuit.

2.7 Temperature stability of the Tank

The temperature coefficient (TC) of a device is the relative change in one of its parameters per degree Celsius or Kelvin. The units are usually in parts variation per million per degrees Celsius (ppm/^{O}C). Taking the case of an oscillator (with an LC tank) the TC is the fractional change of frequency over the centre frequency per 1^{O}C temperature change. Usually the TC for any given component or system is given, to find the change in frequency for a given temperature change, simply multiply the TC by the temperature change and the centre frequency (frequency the oscillator should be running at).

$$\frac{\Delta fc}{fc} = TC \quad x \quad \Delta T$$

An oscillator will always change frequency due to temperature change, because its components have non-zero temperature coefficients. The most likely offender would be the capacitor. The capacitance is normally worked out by C = (ε.A) / d, where ε is the permittivity of the dielectric between a capacitor's plates, A is the common surface area that the plates overlap across the dielectric and d is the distance between the plates. One of the best tuning capacitors available is the silvered mica capacitor (often called the chocolate drop, because of it's smooth brown oval appearance). The variation of centre frequency of an oscillator will now be looked at with respect with capacitance change.

$fc = \frac{1}{2\pi}(LC)^{\frac{1}{2}}$, now differentiate with respect to C and then solve for dfc by multiplying across by dC. Then dividing across by fc will yield $\frac{dfc}{fc} = \frac{\Delta fc}{fc} = -\frac{1}{2}\frac{\Delta C}{C}$, if the capacitance change due to temperature or any other ageing effects is less than 10%. Looking at the equation, it becomes apparent if a 2% increase in capacitance occurs, then a 1% decrease in centre frequency shall take place. This seems trivial but when large frequencies are involved, i.e. 100MHz a 1% change is -1MHz, which is a change of 5 channels down in the commercial bandwidth.

The junction capacitance of the transistor (section 2.9) which also sets the center frequency, is also a major source of frequency instability due to temperature change

2.8 Discrete components to be considered for use in a High frequency circuit

Fixed and variable resistors form the basic components in any electronic circuit, therefore they shall be the first component that will looked at, followed by Capacitors and finally Inductors.

2.8.1 Resistors

The three main factors when choosing a resistor for an intended application are

- Tolerance
- Power Rating
- Stability

The Table below gives a standard overview of the types of resistors used and their specification's

	Thick Film	Metal Film	Carbon Film	Wire-wound
Max. Value	1MΩ	10MΩ	10MΩ	22KΩ
Tolerance	±1% to ±5	±1% to ±5	±1% to ±5	±1% to ±5
Power Rating	0.1 to 1 Watt	0.125 to 0.75W	0.125 to 2W	2.5 W
Temp. Coeff.	±100 to 200ppm/^{O}C	±50 to 200ppm/^{O}C	0 to 700ppm/^{O}C	±30 to 500ppm/^{O}C
Stability	V. Good	V. Good	V. Good	V. Good
Typical Use	for accurate work	Accurate work	General purpose	for low values

2.8.2 Capacitors

Capacitors as mentioned before in a previous section are made up of two conducting plates with a dielectric in between. The most important factors when choosing a capacitor are

- Leakage resistance
- Polarised / non-Polarised
- Temperature Coefficient

	Silvered Mica	**Ceramic**	**Electrolytic**	**Tantalum**	**Polystyrene**
Range	2.2pF to 10nF	1nF to 100nF	0.1μF to 47mF	1μF to 100μF	22pF to 0.1μF
Tolerance	± 1%	-20% to 80%	-10% to 50%	±20%	±1%
Temp. Coeff.	+35ppm/^{o}C	+20% t -80%	±1500 ppm/^{o}C	±500 ppm/^{o}C	-150ppm/^{o}C
Leakage resistance	Very High	High	Very Low	Low	Very High
Stability	Excellent	Good	Fair	Good	Excellent

2.8.2.1 Silver Mica

These capacitors have excellent stability and a low temperature coefficient, and are widely used in precision RF 'tuning' applications

2.8.2.2 Ceramic types

these low cost capacitors offer relatively large values of capacitance in a small low-inductance package. They often have a very large and non-linear temperature coefficients. They are best used in applications such as RF and HF coupling or decoupling, or spike suppression in digital circuits, in which large variations of value are of little importance

2.8.2.3 Electrolytic Types

These offer large values at high capacitance density; they are usually polarised and must be installed the correct way round. Aluminium foil types have poor tolerances and stability and are best used in low precision applications such as smoothing filtering, energy storage in PSU's, and coupling and decoupling in audio circuits.

2.8.2.4 Tantalum types

Offer good tolerance, excellent stability, low leakage, low inductance, and a very small physical size, and should be used in applications where these features are a positive advantage.

2.8.2.5 Poly Types

Of the four main 'poly' types of capacitor, polystyrene gives the best performance in terms of overall precision and stability. Each of the others (polyester, polycarbonate and polypropylene) gives a roughly similar performance and is suitable for general purpose use. 'Poly' capacitors usually use a layered 'Swiss-roll' form of construction. Metallised film types are more compact that layered film-foil types, but have poorer tolerances and pulse ratings than film-foil types. Metallised polyester types are sometimes known as 'green-caps'

2.8.2.6 Trimmer capacitors

Polypropylene capacitors are ideal variable capacitors, a fact due to the polypropylene dielectric having a high insulation resistance with a low temperature coefficient. The polypropylene variable capacitor comes in a 5mm single turn package, which is suitable for mounting directly on to a PCB. The typical range of capacitance involved would be from 1.5pF to 50pF.

2.8.3 Inductors

There are two types of inductors that can be discussed, and they are

- Manufactured inductor
- Self made inductor

2.8.3.1 Manufactured inductor

When choosing an inductor from a manufacturer, the core in the coil and the over all Q factor will have to be taken into account. The core should preferably be made of soft ferrite which will in turn minimise the energy losses of the inductor and therefore increase the Q factor. The ferrite core can be adjusted to give a slight change in inductance

2.8.3.2 Self Made inductor

Inductors can be easily wound around air cored formers, there are a number a various manufactured air cored formers on the market. Self made inductors are very useful when a particular inductance is desired.

$$L = N^2\left(\frac{d^2}{18d + 40b}\right)$$

where L = inductance in μH

d = diameter, in inches

b = coil length, inches

N = number of turns

$$N = \frac{\sqrt{L(18d + 40b)}}{d}$$

2.9 NPN Transistor

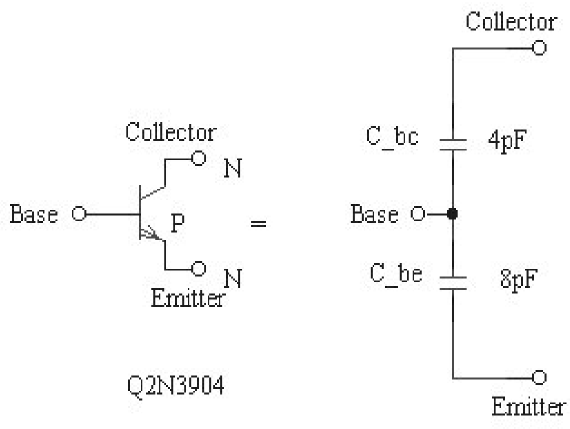


Figure 2.9-1

PNP bipolar and P channel J-Fets are widely used at low frequencies, the preference for high frequency systems lies with the NPN and N channel J-Fets. This is due to the electrons being the majority carriers in both the BJT's and J-Fet's conduction channel.

The NPN BJT is the most commonly used and for the rest of this discussion will be the transistor that will be focused on.

- The bias current acts as a controlled flow source which steadily opens up the collector emitter channel enabling charge carriers to flow, this can be analogous to a slues gate, this rate of flow is controlled by the current gain $\beta = I_C/I_B$.
- Transistors are non-linear especially when biased in the saturation region.
- The Input impedance drops as the biasing current being sinked to the collector increases.
- As the base current increases to allow more collector current through, the current gain β also increases.
- The collector-emitter voltage has a maximum value that cannot be exceeded at an instant in time.

2.9.1 High Frequency Response

The most interesting property is the junction capacitance from the base to emitter and base-collector, the Figure 2.9-2 shows that for the 2N3904, the base-emitter capacitance is larger than the base-collector, because of heavier extrinsic doping and it's forward biasing the depletion region is naturally smaller than the base-collector's. As the frequencies are increased the two capacitances will drop. Because the capacitors are effectively in series, the smaller one dominates (base-collector capacitance). The capacitance is also influenced by the rate of change in base current magnitudes.

A resistance exists of typically in the order of tens of ohms at the base, this parasitic is caused by impure contact between the base's polysilicon to silicon junction. This coupled with the r'e resistance and the current gain makes up the input resistance of the transistor. $Rin = \beta(R_{base} + r'e)$; as stated previously the r'e will inevitably drop as the frequency increases, therefore Rin (base) will inevitably be equal to $\beta(R_{base})$. This makes the system rather unstable, as R_{base} is essentially parasitic impedance. To increase stability RE, (which is normally RF bypassed), will have to be introduced.

Another inherent flaw which might be used to some advantage in the high frequency response of the NPN model, is that of output collector signals are be fed back to the base. This increases the likelihood of continuous oscillation at high frequencies. The importance of this flaw can be seen when oscillators will be discussed in section 3.5.

2.10 Transistor Amplifiers

Now that the basic electronic components have been considered, a look at the 3 transistor amplifiers is worthwhile prelude to the next section, which contain references and examples of these amplifiers. The three amplifiers are called Common Emitter, Common collector and Common Base.

2.10.1 Common Emitter

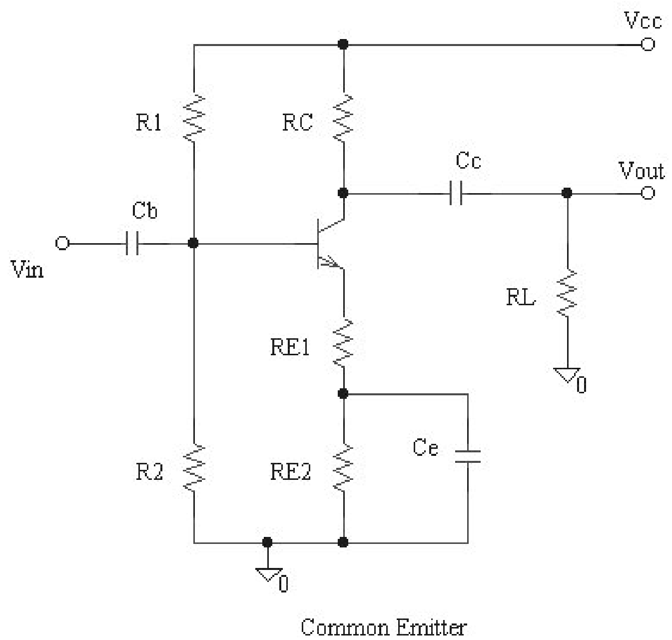


Figure 2.10-1

r'c and r'e are the junction resistances at the collector and emitter respectively. r'c is seen as infinite (reverse bias junction), r'c is equal to the threshold voltage V_T divided by the emitter current.

$I_C = I_B + I_E$, I_B is relatively small compared to I_B the base current $\Rightarrow I_C \approx I_E$.

All capacitor's used here are DC opens and AC shorts. The supply ideally has no impedance and therefore no voltage dropped across it. So it is an AC ground

DC Analysis

Voltage at the base, $Vb = \left(\frac{R2}{R1+R2}\right)Vcc$, Voltage at the Emitter, Ve = Vb - 0.7.

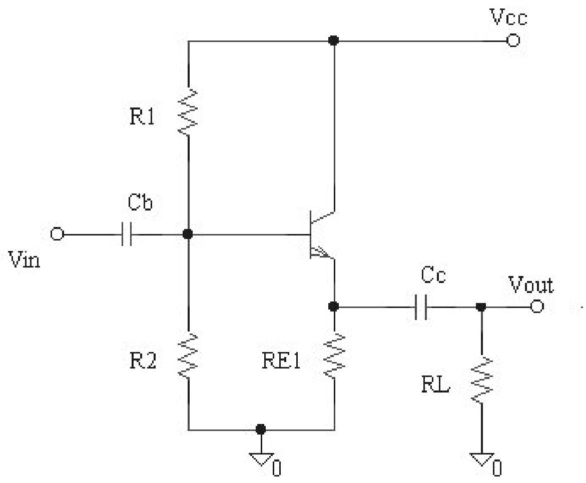
Emitter Current, Ie = Ve / (RE1 + RE2) ≈ Ic, Voltage at collector, Vc = Vcc - (Ic.RC)

Voltage across the collector and emitter, Vce = Vc - Ve

AC Analysis

- $Rin_{(base)} = \beta(RE1 + r'e)$; Input Impedance, $Rin = R1//R2//Rin_{(base)}$.
- Output impedance , Rout = (RC // r'c) ; r'c >>RC, therefore Rout ≈ RC.
- Voltage gain, Av = RC / (RE1 + r'e), note RE1 is not bypasses because it is more independent of temperature change than r'e and therefore increasing stability against temperature change.
- Current gain, $Ai = I_C / I_B = \beta$
- Power Gain, Ap =Av * Ai

2.10.2 Common Collector (emitter follower)



Common Collector or Emitter Follower

Figure 2.10-2

DC analysis is similar to the common emitter.

AC Analysis

- Input impedance is the same as the common emitter.
- Output impedance, Rout =RE1 // r'e ; RE1 >> r'e ⇒ Rout ≈ r'e (quite low!)
- Voltage gain, Av = RE / (RE1 + r'e) ; RE1 >> r'e ⇒ Av ≈ 1
- Current Gain, Ai = I_E / I_B ≈ β
- Power Gain, Ap ; Same as common emitter.

2.10.3 Common Base

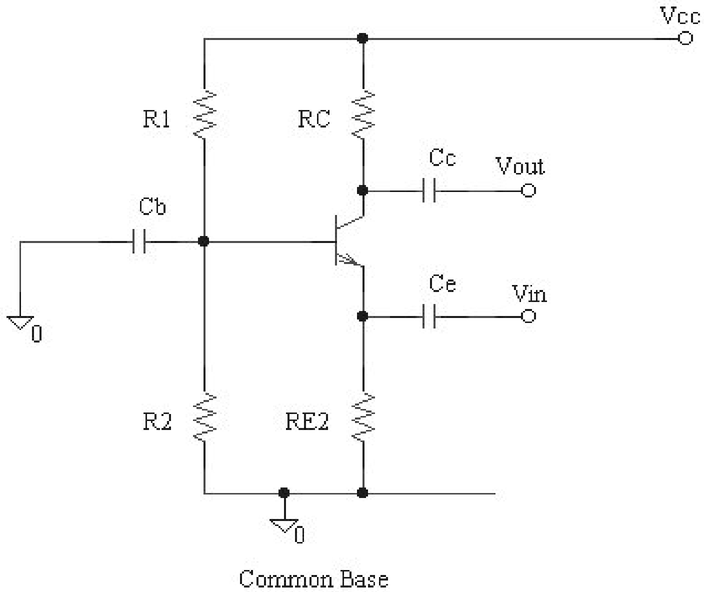


Figure 2.10-3

DC analysis is similar to the common emitter.

AC Analysis

- Input Impedance, Rin =RE1 // r'e ; RE1 >> r'e ⇒ Rout ≈ r'e
- Output impedance, Rout = (RC // r'c) ; r'c >>RC, therefore Rout ≈ RC.
- Voltage gain, Av = RC / r'e
- Current Gain, Ai = I_C / I_E ≈ 1
- Power Gain, Ap = Av * Ai ≈ Av *1 ⇒ Ap ≈ Av.

3 Basic Building blocks for an FM transmitter

3.1 Introduction

When creating a system for transmitting a frequency modulated wave a number of basic building blocks have to be considered, the diagram below gives a very broad impression of the transmitter and it's individual parts.

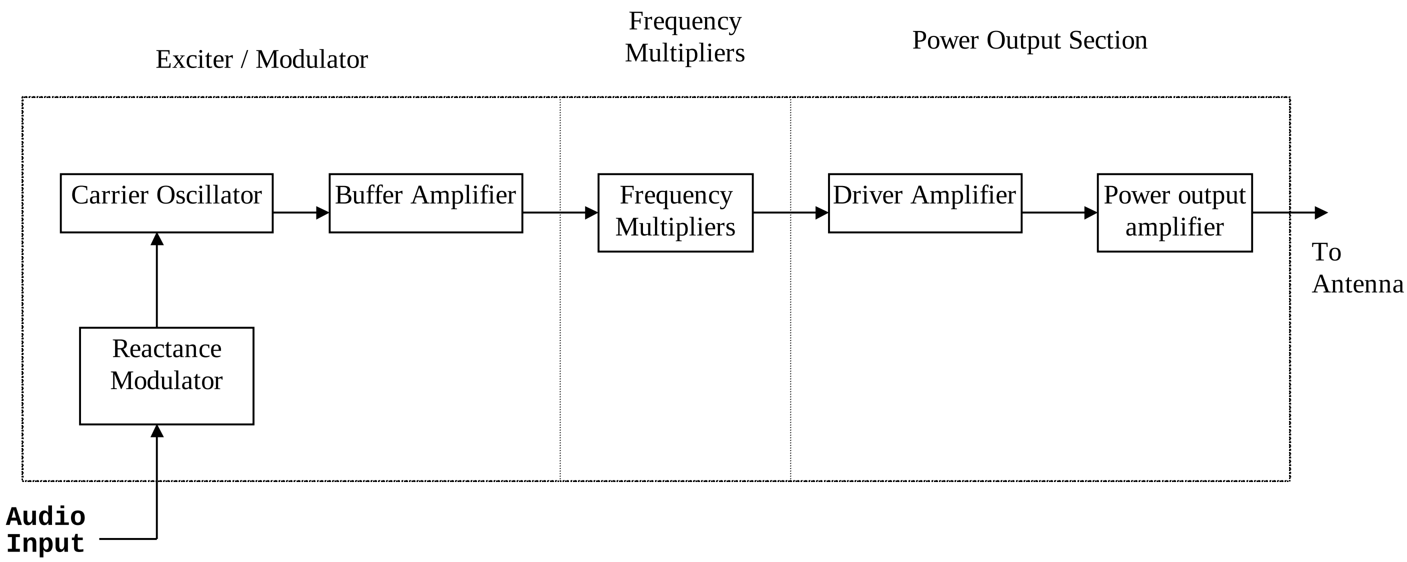


Figure 3.1

3.2 General Overview

3.2.1 Exciter /Modulator

- Carrier Oscillator generates a stable sine wave for the carrier wave. Linear frequency even when modulated with little or No amplitude change
- Buffer amplifier acts as a high impedance load on oscillator to help stabilise frequency.
- The Modulator deviates the audio input about the carrier frequency. The peak + of audio will give a decreased frequency & the peak - of the audio will give an increase of frequency

3.2.2 Frequency Multipliers

- Frequency multipliers tuned-input, tuned-output RF amplifiers. In which the output resonance circuit is tuned to a multiple of the input .Commonly they are *2 *3*4 & *5 .

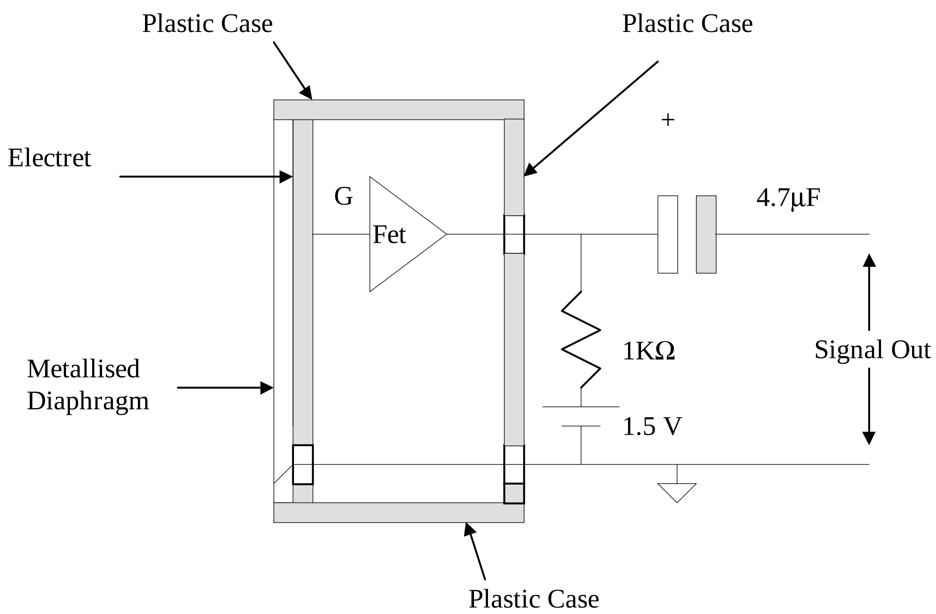
3.2.3 Power output section

- This develops the final carrier power to be transmitter.

Also included here is an impedance matching network, in which the output impedance is the same as that on the load (antenna).

3.3 The Microphone

Microphones are acoustic to electrical transducers. The four best known variations of these are the moving coil (‘dynamic’), ribbon, piezo-electric (‘crystal’), and electret (‘capacitor’). The electret type will be discussed because of there incredibly small size and high performance at audio frequencies.



Figure

A light weight metallised diaphragm forms one plate of a capacitor and the other plate is fixed, the capacitance thus varies in sympathy with the acoustic signal. The capacitance acquires a fixed charge, via a high value resistor (input impedance of FET) and since the voltage across a capacitor is equal to its charge divided by its capacitance, it will have a voltage output which is proportional to the incoming audio (baseband).

The fixed plate at the back is known as Electret which holds an electrostatic charge (dielectric) that is built in during manufacture and can be held for about 100 years. The IGFET (needs to be powered by a 1.5 volt battery via a 1KΩ resistor) output is then coupled to the output by an electrolytic capacitor.

3.4 Pre-emphasis

Improving the signal to noise ratio in FM can be achieved by filtering, but no amount of filtering will remove the noise from RF circuits. But noise control is achieved in the low frequency (audio) amplifiers through the use of a high pass filter at the transmitter (pre-emphasis) and a low pass filter in receiver (de-emphasis) The measurable noise in low- frequency electronic amplifiers is most pronounced over the frequency range 1 to 2KHz. At the transmitter, the audio circuits are tailored to provide a higher level, the greater the signal voltage yield, a better signal to noise ratio. At the receiver, when the upper audio frequencies signals are attenuated t form a flat frequency response, the associated noise level is also attenuated.

3.5 The Oscillator

The carrier oscillator is used to generate a stable sine-wave at the carrier frequency, when no modulating signal is applied to it . When fully modulated it must change frequency linearly like a voltage controlled oscillator. At frequencies higher than 1MHz a Colpitts (split capacitor configuration) or Hartley oscillator (split inductor configuration) may be deployed.

A parallel LC circuit is at the heart of the oscillator with an amplifier and a feedback network (positive feedback). The Barkhausen criteria of oscillation requires that the loop gain be unity and that the total phase shift through the system is 360^{o}. I that way an impulse or noise applied to the LC circuit is fed back and is amplified (due to the fact that in practice the loop gain is slightly greater than unity) and sustains a ripple through the network at a resonant frequency of $\dfrac{1}{2\pi\sqrt{LC}}$ Hz.

The Barkhausen criteria for sine-wave oscillation maybe deduced from the following block diagram

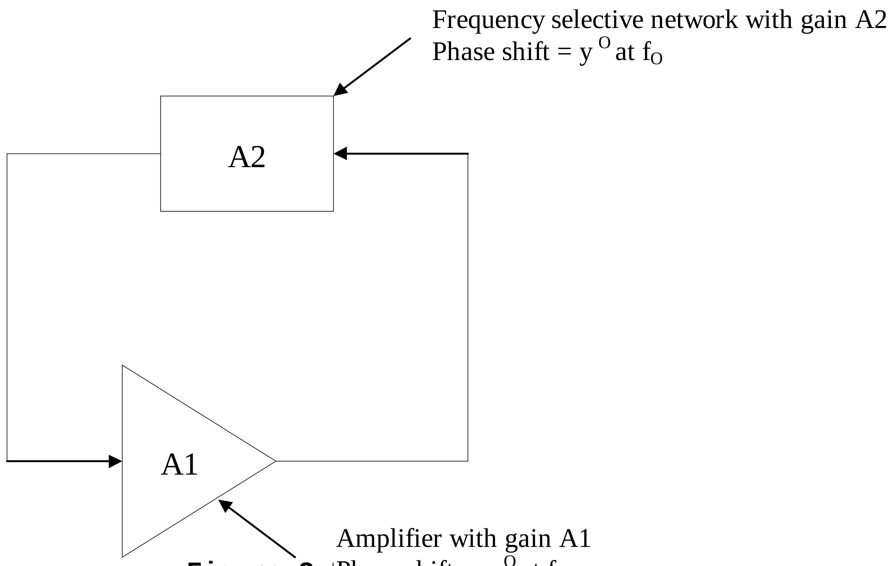


Figure 3.

Condition for oscillation

$x^o + y^o = 0^o$ or 360^o

Condition for Sine-wave generation

A1 * A2 = 1

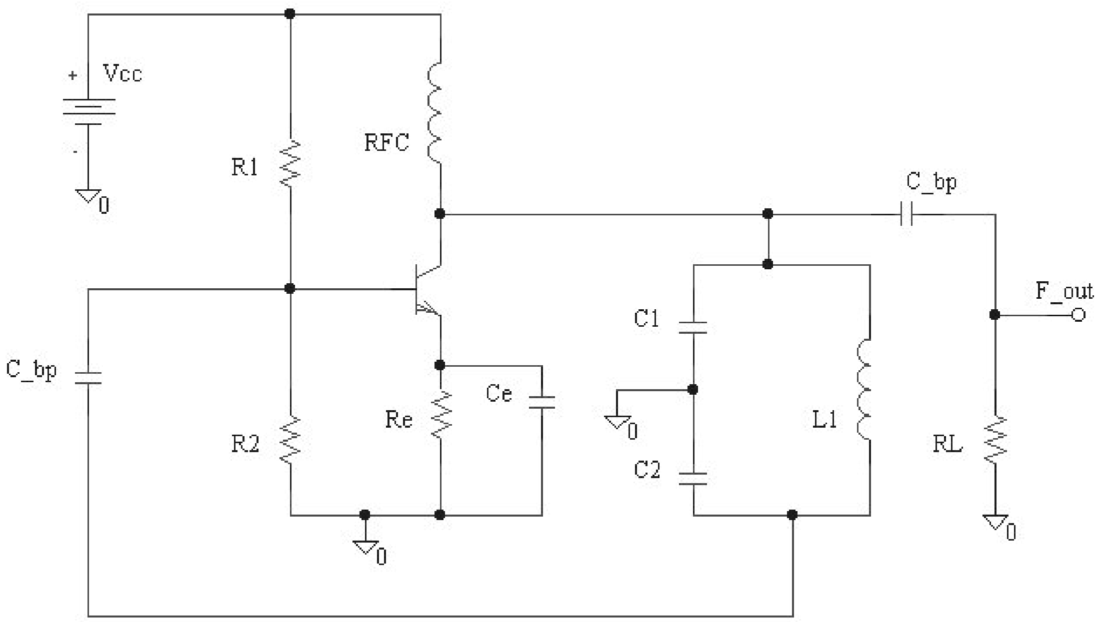


Figure 3.5-2

The above circuit diagram is an example of a colpitts oscillator, an LC (L1, C1 &C2) tank is shown here which is aided by a common emitter amplifier and a feedback capacitor (C_fb) which sustains oscillation. From the small signal analysis in order for oscillation to Kick off and be sustained $Gm * RL = \frac{C2}{C\ 1}$ the frequency of the oscillator is found to be $\frac{1}{2\pi \sqrt{LC_*}}$, where C_* is $\frac{C1 * C2}{C1 + C2}$.

3.6 Reactance modulator

The nature of FM as described before is that when the baseband signal is Zero the carrier is at it's "carrier" frequency, when it peaks the carrier deviation is at a maximum and when it troughs the deviation is at its minimum. This deviation is simply a quickening or slowing down of frequency around the carrier frequency by an amount proportional to the baseband signal. In order to convey the that characteristic of FM on the carrier wave the inductance or capacitance (of the tank) must be varied by the baseband. Normally the capacitance of the tank is varied by a varactor diode. The varactor diode (seen below) when in reverse bias has a capacitance across it proportional to the magnitude of the reverse bias applied to it. The formula for working out the instantaneous capacitance is shows that as the reverse bias is increased the capacitance is decreased.

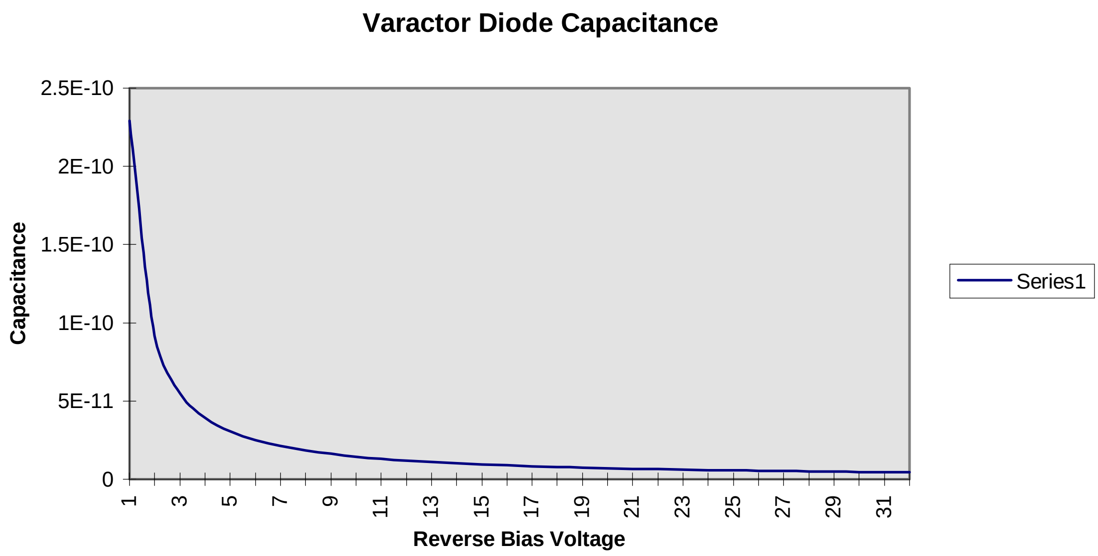


Figure 3.6-1

$$C_D = \frac{C_O}{\sqrt{1 + 2|V_R|}}$$ where C_O is the capacitance at zero Reverse bias voltage

Applying this to an LC tank : as the capacitance decreases the frequency increases. So placing a fixed reverse bias on the varactor will yield a fixed capacitance which can be placed in parallel capacitor and inductor. A bypass capacitor can be used to feed the baseband voltage to the varactor diode, the sine-wave baseband voltage has the effect of varying the capacitance of the varactor up and down from the level set by the fixed reverse voltage bias. As the baseband peaks the varactor's capacitance is at a minimum and the overall frequency will increase, applying this logic to when the baseband troughs the frequency will decrease. Looking at the three cases for the varactor diode, Maximum capacitance, Nominal capacitance set by V_bias (no modulation) and Minimum capacitance and observing the frequency will show that by modulating the reactance of the tank circuit will bring about Frequency Modulation

$$F_{NOM} = \frac{1}{2\pi\sqrt{L(C1 + Cd_{NOM})}}$$

with no baseband influence (the carrier frequency)

$$F_{MIN} = \frac{1}{2\pi\sqrt{L(C1 + Cd_{MAX})}} \qquad F_{MAX} = \frac{1}{2\pi\sqrt{L(C1 + Cd_{MIN})}}$$

with peak negative baseband influence   with peak positive baseband influence

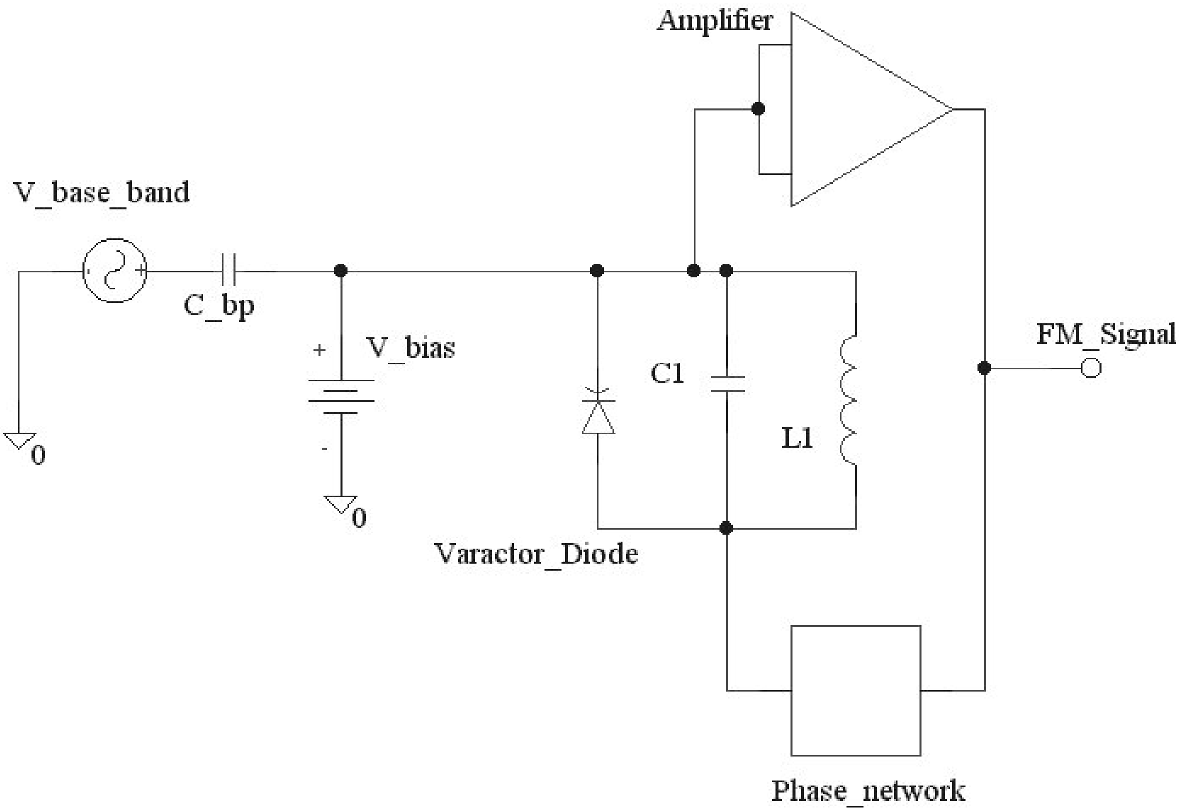


Figure 3.6-2

The diagram below show's a proposed modulation scheme, with the amplifier and phase network discussed earlier in the oscillator section.

3.7 Buffer Amplifier

The buffer amplifier acts as a high input impedance with a low gain and low output impedance associated with it. The high input impedance prevents loading effects from the oscillator section, this high input impedance maybe looked upon as R_L in the analysis of the Colpitts Oscillator. The High impedance R_L helped to stabilise the oscillators frequency.

Looking at the Buffer amplifier as an electronic block circuit, it may resemble a common emitter with low voltage gain or simply an emitter follower transistor configuration.

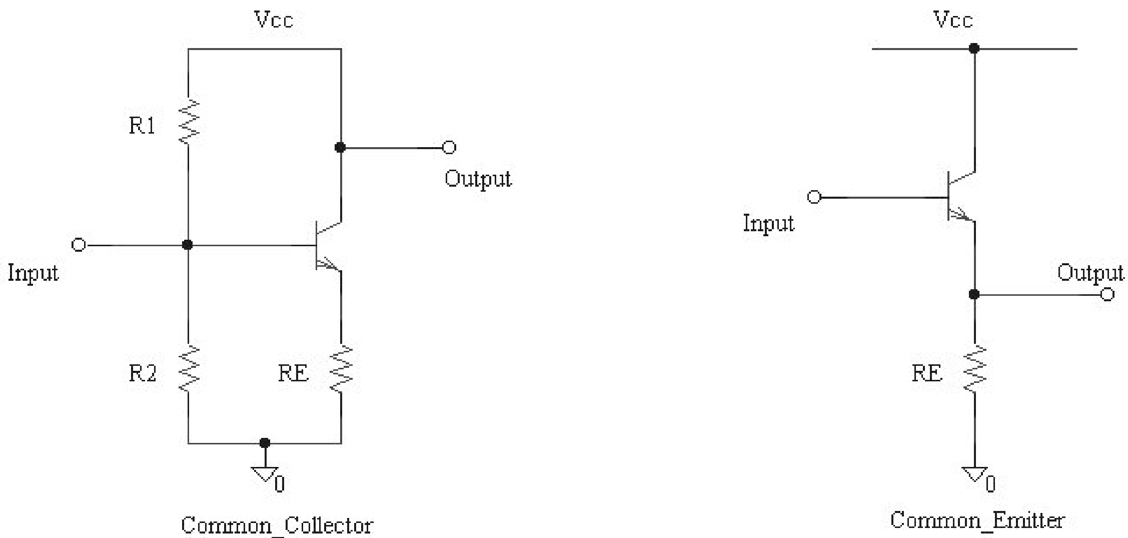


Figure 3.7-1

3.8 Frequency Multipliers

Frequency modulation of the carrier by the baseband can be carried out with a high modulation index, but this is prone to frequency drift of the LC tank, to combat this drift, modulation can take place at lower frequencies where the Q factor of the tank circuit is quite high (i.e. low bandwidth or less carrier deviation) and the carrier can be created by a crystal controlled oscillator. At low frequency deviations the crystal oscillator can produce modulated signals that can keep an audio distortion under 1%. This narrow-band angle modulated wave can be then multiplied up to the required transmission frequency, the deviation brought about by the baseband is also multiplied up, which means that the percentage modulation and Q remain unchanged. This ensures a higher performance system that can produce a carrier deviation of ±75Khz.

Frequency multipliers are tuned input, tuned output RF amplifiers, where the output resonant tank frequency is a multiple of the input frequency. The diagram of the simple multiplier below shows the output resonant parallel LC tank which is a multiple of the input frequency.

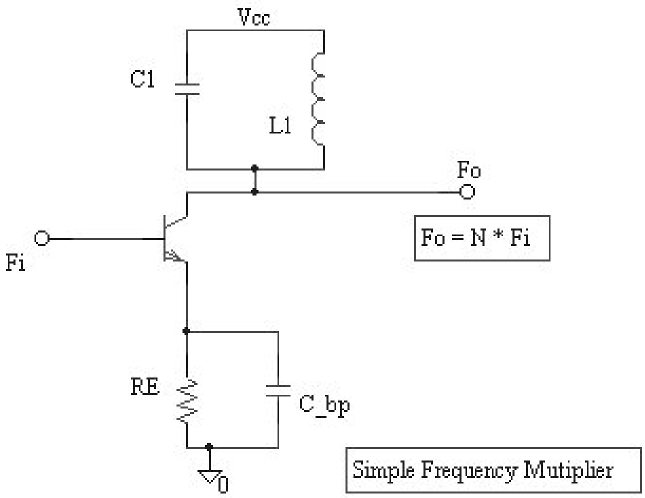


Figure 3.8-1

The circuit above is good for low multiplying factors (i.e. *2), for triplers and especially quadruplers, current idlers are used to improve efficiency. These series resonant LC's help in the output filtering of the input, but more importantly they aid in the circulation of harmonic currents to enhance the transistor's non-linearity. The idlers can be tuned to fi, 2fi , N-1(fi), the final output tank is tuned to fo = N(fi).

Other devices can be used instead of the transistor, one of which is called a Step Recovery Diode (SRD) or snap diode : it accumulates part of the input cycle and then releases it with a snap. The circuit efficiency or power loss is proportional to 1/N as opposed to $1/N^2$ for a good transistor multiplier. Of course the transistors current gain will make up for some of the loss provided by the transistor multiplier circuit.

So for high efficiency transistor power amplifiers, it is important to realise that most of the non-linearity is provided in the collector-base junction (varactor diode behaviour) and not the base-emitter, in order to maintain a high current gain.

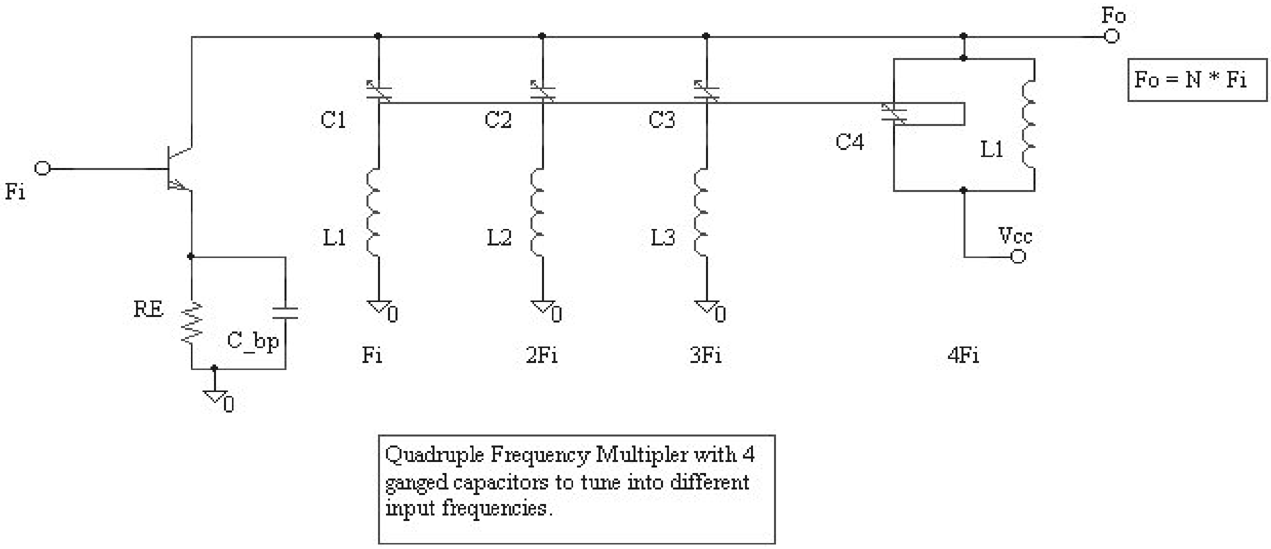


Figure 3.8-2

The above multiplier circuit is a quadrupler and is used in very complex transmitter systems, because of its size and relative complexity it will not be included in the final design for the project, but it is worth noticing how it increases efficiency compared with the first simpler Class-C operation multiplier circuit.

3.9 Driver Amplifier

The driver amplifier can be seen to do the same function as the buffer amplifier, i.e. a high input impedance, low gain (close to unity) and low output impedance between the frequency multiplier and power output stages of the transmitter. The circuitry is the same as discussed in the Buffer amplifier description.

3.10 Power Output Amplifier

The power amplifier takes the energy drawn from the DC power supply and converts it to the AC signal power that is to be radiated. The efficiency or lack of it in most amplifiers is affected by heat being dissipated in the transistor and surrounding circuitry. For this reason , the final power amplifier is usually a Class-C amplifier for high powered modulation systems or just a Class B push-pull amplifier for use in a low-level power modulated transmitter. Therefore the choice of amplifier type depends greatly on the output power and intended range of the transmitter.

3.11 Antenna

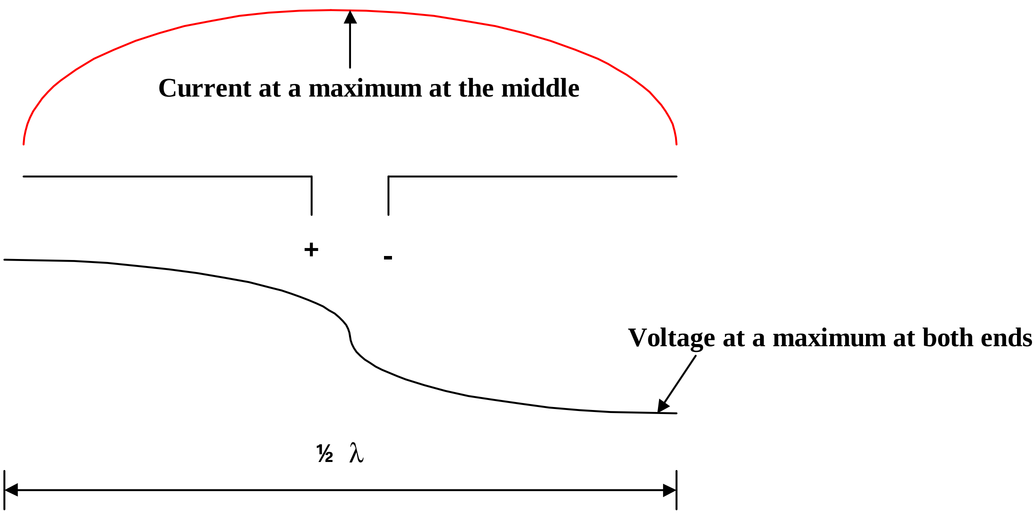


Figure 3.11-1

The final stage of any transmitter is the Antenna, this is where the electronic FM signal is converted to electromagnetic waves, which are radiated into the atmosphere. Antennas can be Vertically or Horizontally polarised, which is determined by their relative position with the earth's surface (i.e. antenna parallel with the ground is Horizontally polarised). A transmitting antenna that is horizontally polarised transmits better to a receiving antenna that is also horizontally polarised, this is also true for vertically polarised antennas. One of the intended uses for the transmitter is as a tour guiding aid, where a walkman shall be used as the receiver, for a walkman the receiving antenna is the co-axial shielding around the earphone wire. The earphone wire is normally left vertical, therefore a vertically polarised whip antenna will be the chosen antenna for this particular application.

3.11.1 Radiation Resistance

The power radiated by an antenna is given by the Poynting vector theorem ρ = E X H watts/m^2. Getting the cross product of the E (electric field strength) and H (magnetic field strength) fields ,multiply it by a certain area ($\pi.r^2$) and equating the resulting power to $I^2.Rr$, Rr the radiation resistance maybe obtained.

$$I^2 Rr \quad = \quad Power \quad = \quad 80.\pi^2.I^2 \quad \left(\frac{dl}{\lambda}\right)^n$$

$$Rr \quad = \quad 80.\pi^2 \quad \left(\frac{dl}{\lambda}\right)^n$$

Where dl is the length of the antenna, λ is the wavelength and n is an exponent value that can be found by using (dl/λ) on the y-axis and then n can be found on the x-axis.

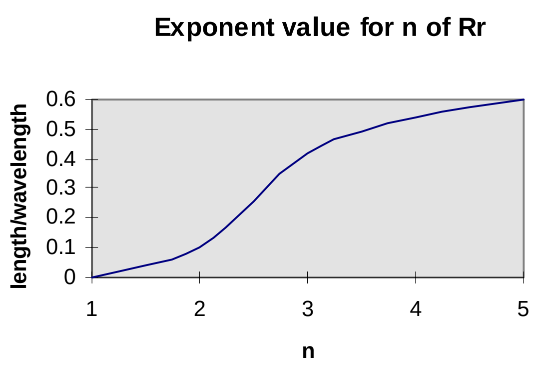


Figure 3.11-2

Taking a centre fed dipole with a length of approximately half a wavelength, due to a capacitive effect at the ends of the antenna the overall length in practice is shorter (95% of the theoretical length). For dl half the wavelength, n is found to be 3.2.

$Rr = 789.5 * (0.5 * .95)^{3.2} = 72.9 \approx 73\Omega$

For an end fed half wavelength making a few elementary changes to the above equation, i.e. making the length 97.5% and halving and then negating the exponent to give n = -1.6 which results in the radiation resistance equal to $789.5 * (0.5 * .975)^{-1.6} = 2492 \approx 2.5K\Omega$

3.11.2 Power transfer

Maximum power transfer between the antenna and the electronics circuitry will have to be looked at in order to produce an antenna that will be efficient in transmitting an audio signal to a receiver. Taking the case of the receiver with an antenna of impedance Zin connected with the input terminal, which is terminated with a resistor Rg. The maximum power transfer theorem shows that with a voltage induced in the antenna the current flowing into the receiver will be I = V / (Zin + Rg). The power transferred will be I^2 .Rg, differentiating the power with respect to Rg and letting the derivative equal to Zero for max. power transfer, it is shown that Zin + Rg = 2Rg, which means that Rg will be equal to Zin.

3.11.3 Reciprocity

The theorem for reciprocity states that if an emf is applied to the terminals of a circuit A and produces a current in another circuit B, then the same emf applied to terminals B, will produce the same current at the terminals of circuit A. Simply put means that every antenna will work equally well for transmitting and receiving. So applying the same logic of max. power transfer at the receiver to a transmitter circuit, the output impedance of the transmitter must match the input impedance of the antenna, which can be taken as the radiation resistance of the antenna.

Now that a qualitative view of some of the characteristics of an antenna have been looked at, it is now time to look at some of the basic types of antenna that can be considered for this project

3.11.4 Hertz Dipole

The Hertz Antenna provides the best transmission of electromagnetic waves above 2 MHz, with a total length of ½ the wavelength of the transmitted wave.

Placing the + and - terminals in the middle of the antenna and ensuring that the impedance at the terminals is high (typically 2500Ω) and the impedance at the open ends is low (73Ω). This will ensure that the voltage will be at a minimum at the terminal and at a maximum at the ends, which will efficiently accept electrical energy and radiate it into space as electromagnetic waves. The gap at the centre of the antenna is negligible for frequencies above 14Mhz.

3.11.5 Monopole or Marconi Antenna

Gugliemo Marconi opened a whole new area of experimentation by popularising the vertically polarised quarter wave dipole antenna, it was theorised that the earth would act as the second quarter wave dipole antenna. Comparing the signal emanating from the quarter wave antenna in µV/m, it has been shown experimentally that for a reduction in the antenna from $\lambda/2$ to $\lambda/4$ a reduction of 40 % (in µV/m) takes place, for a reduction $\lambda/4$ to $\lambda/10$ a reduction of only 5% (in µV/m). This slight reduction of .05 in transmitted power for a decrease of .75 in antenna length seems impressive, but their is a decrease in the area of coverage.

When considering an antenna type and size for this project 2 things have to be taken into account, the frequency of transmission and the portability of the antenna.

Transmitting in a frequency range of 88 to 108 MHz, the mean frequency is $(88 * 108)^{1/2}$ = 97.5MHZ. Rounding this off to 100MHz, calculating the wavelength gives $(3*10^8 / 100*10^6)$ yields a wavelength of approximately 3 metres. $\lambda/2 = 1.5$ m ; $\lambda/4 =$ .75m ;$\lambda/10 = 30$cm

The above analysis concludes that the use of an adjustable end fed whip antenna with an affective length of 30 to 75 cm could be used with considerable affect.

3.12 Impedance matching

Between the final power amplifier of the transmitter and the antenna, an impedance matching network maybe be considered. One of the possible surprises in power amplifiers is the realisation that output impedance matching is not based on the maximum power criteria. One reason for this, is the fact that matching the load to the device output impedance results in power transfer at 50% efficiency.

An impedance matching system maybe merely a special wide-band transformer which is used for broadband matching (i.e. between 88 & 108Mhz), which maybe a two pole LC band-pass or low pass resonant circuits to minimise noise and spurious signal harmonics. The purpose of the impedance matching network is to transform a load impedance to an impedance appropriate for optimum circuit operation. Detailed analysis and calculations will be used latter on when evaluating the final design of the system.

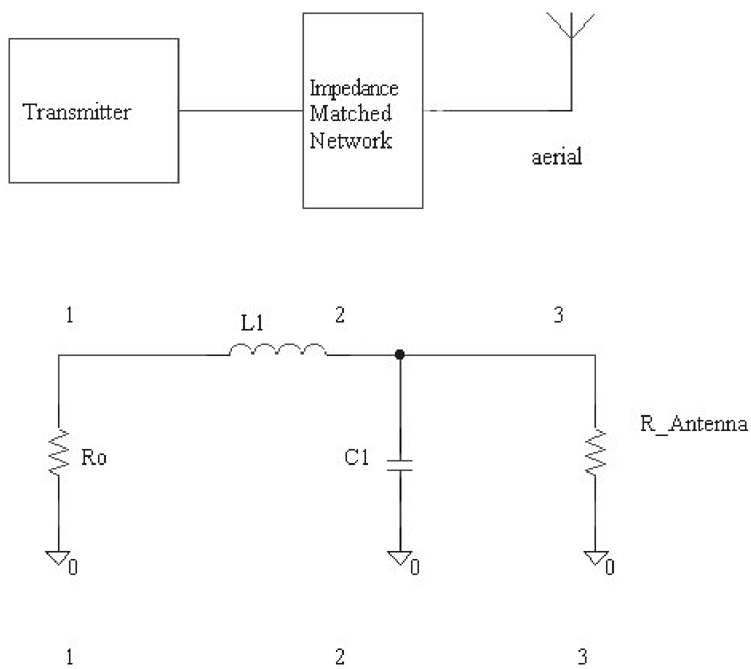


Figure 3.12-1

Here are a few equations that determine the inductance and capacitor values of Figure 3.12-1, when R_L (resistance of the antenna) and Ro (the output impedance of the antenna) are known.

$$Q = \sqrt{(RL/Ro) \quad -1}$$

$$X_L = \sqrt{Ro.RL - Ro^2}$$

$$Xc = {}^{Ro.R_L}\!/_{X_L}$$

The use of this matching network is predicted on the fact that Ro < R_L according to the equation for calculating the inductance $X_{L.}$ This method of matching is similar to the so called quarter wave transformer for transmission lines.

4 Designs Under consideration

4.1 Introduction

Considering all the factors Ranging from frequency modulation theory (section 1) to electronic component properties (section 2) and then individual transmitter stages (section 3), it is now possible to have a look at complete FM transmitter designs. There are four different possible designs covered in this chapter, each includes a diagram and a brief explanation on how it works and discussion on whether it meets the criteria of this project, i.e. (miniature, low powered and Multichannel)

4.2 Phase locked loop

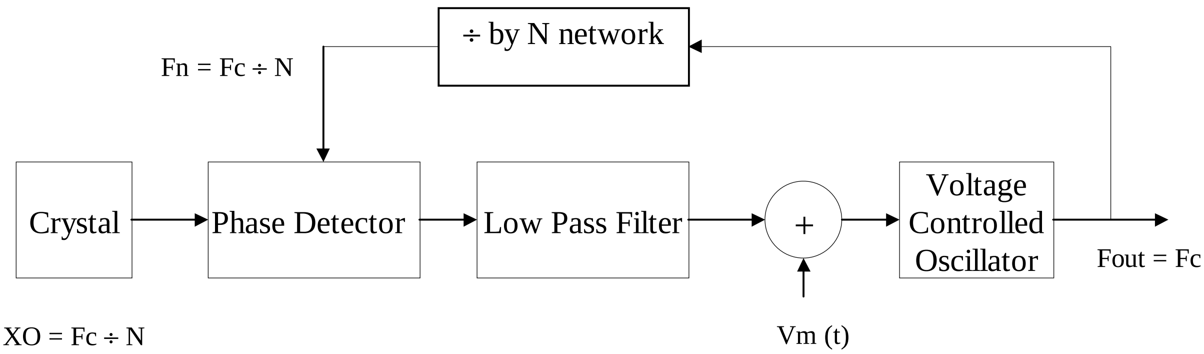


Figure 4.2.0

The above transmitter block gives a conceptual feel for this type of PLL implementation. LC tuned VCO's have good deviation sensitivity, but poor stability with respect to frequency drifts due to the ageing affects and non-zero temperature coefficients of the inductor and capacitor. This is where the feedback stability of a PLL comes into play, by dividing the output carrier frequency by a factor which will make it equal to a reference frequency such as a crystal oscillator. The divide by N network also plays a part in minimising interference from the crystal oscillator (XO). The low pass filter prevents feedback of modulated frequencies and eliminates the possibility of the loop locking to a side band.

The overall system is quite stabile, which is good, but the circuitry involved is quite large even if the devices were all integrated circuits it still would be quite bulky and rather complicated. The complication would be in making the system multi-channelled. For multi-channel capability it would mean changing the divide by N factor and including another block to change the reference frequency XO to equal that of the feedback network in order for the tracking of phase to be possible when the VCO drifts from its centre frequency.

4.3 Stand Alone VCO

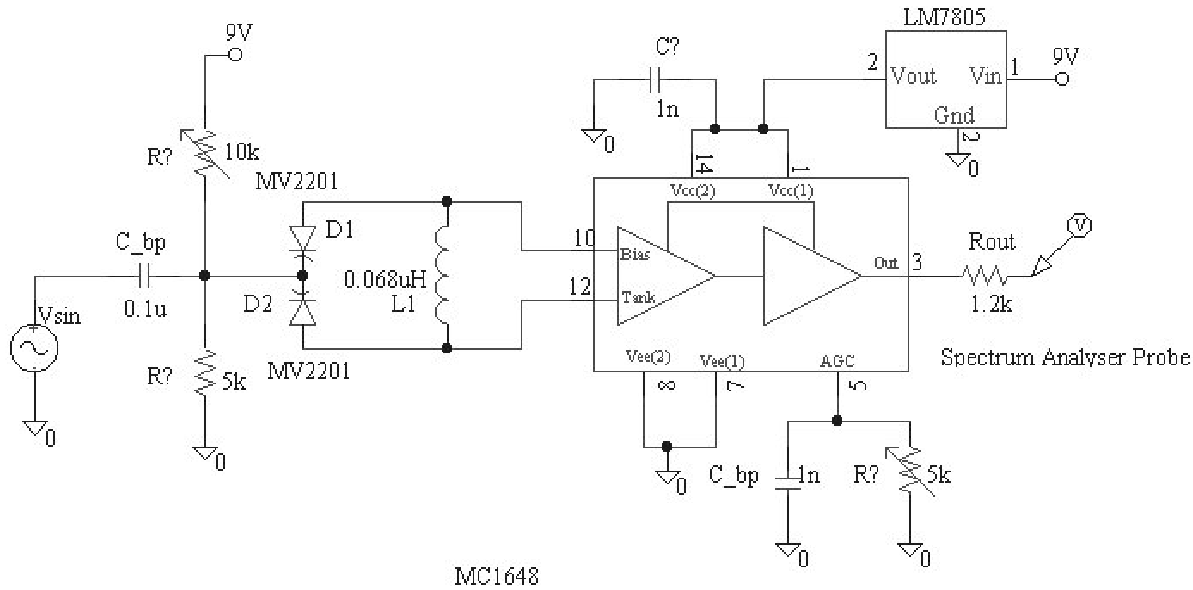


Figure 4.3-1

Note : the 7805 is a 5 volt regulator, which enables the MC1648 to be powered from a 9 Volt battery.

This was a design that built and tested for the feasibility of using a VCO on it's own for wide-band frequency modulation of an audio input. The MC1648 (Appendix C) is the voltage-controlled oscillator used at the heart of this modulation scheme. The frequency of the tank is controlled by the resonant frequency of L1, D1 and D2. D1 and D2 (MV1404 was used) are both varactor diodes, which as seen before in section 3.6, will have a nominal value of capacitance when a certain reverse dc bias is applied to it, the 10K (variable) / 5K potentiometer takes care of this bias voltage. D1 and D2 are effectively in parallel and their effective capacitance is added together. To change the output carrier frequency the 10 variable resistor is varied. A signal generator was used to simulate the audio baseband, its voltage was varied from .5 to 1.5 volts and its frequency was varied from 200Hz to 10Khz. A 1.2K resistor was used in conjunction with a probe, which was connected into a spectrum analyser. The results were as expected (from the data-sheet), for a 5.5 volt bias applied, 100MHz was seen on the analysers, screen and side-bands were also seen as a result of the voltage generator, the side-bands increased as the baseband voltage and frequency was increased, which shows Carson's Rule in practice. A Walkman radio receiver was set to 100MHz and the voltage generator's signal could be successfully demodulated. As the voltage was increased at the signal generator, the sound in the receiver's earphone became louder and as the generators frequency was increased, the sound increased in pitch, proving that modulation and demodulation had taken place.

This is a rather interesting design, but it has to be considered as only a functional block and not a complete transmitter. To make it into a transmitter an audio amplifier section needs to be inserted in order to interface with a microphone for audio modulation and possibly a Class-C output amplifier terminated with an impedance matched network before going into an antenna. If this were done, the transmitter would take up quite a considerable area and this is only considering single-channel transmission. Making the transmitter multi-channel would add on extra circuitry due to the fact that a more stable method than just tweaking the variable resistor will have to be found and also the class-c's output tank resonant frequency would also have to be changed. For this reason alone, it cannot be considered as part of a final working design.

4.4 Two transistor Design

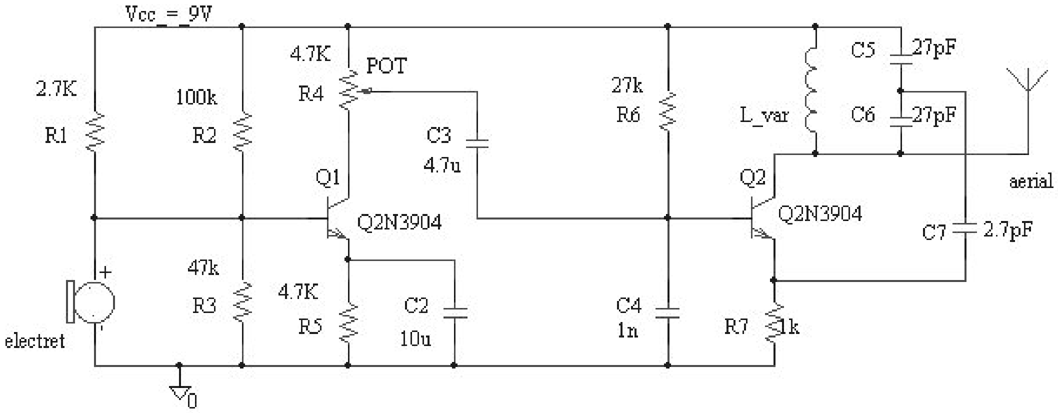


Figure 4.4-1

This simple FM transmitter is built around two amplifiers, Q1 is a common emitter with a dc gain of 1 and an Ac gain that can be set by the potentiometer R4, this will amplify the signal from the Electret and pass it on to the next stage by coupling capacitor C3.

Q2 is at the heart of the RF section, because of C4 (which ac grounds the base) and the feedback cap C7 (that splits the capacitance C5 and C6) the RF section is a colpitts oscillator in the common-base mode. The inductance of L_var and the effective capacitance of C5 and C6 work out the centre frequency. The base-collector junction capacitance (which acts like a varactor diode) is varied as the amplified base-band signal changes it's reverse-bias voltage, this capacitance will inevitably be part of the over all tuning capacitance of the resonant tank. The Antenna, (very short end fed wire) can be resistively matched by an ordinary low-value resistor.

This is quite an effective little transmitter that can be easily made and has a range of about 60 feet indoors.

4.5 One transistor design

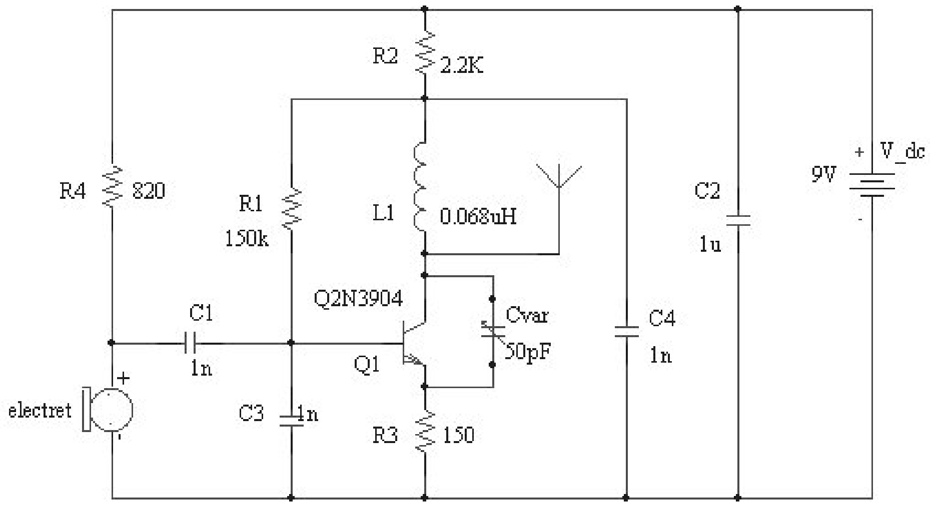


Figure 4.5-1

The transmitter above makes use of the old 'reflex' technique dating from the time when active devices were expensive and were sometimes made to perform two functions at the same time.

In this case the transistor Q1 is acting as an audio amplifier for the signal from the Electret microphone. The amplified signal appears at the collector, R2 being the collector load resistor. R1 provides bias and DC feedback to set the collector to about 3.4V producing a simple common emitter amplifier.

At the same time the transistor is operating as a common base oscillator at VHF, the base being grounded to RF by c3. RF feedback to the emitter through TC1 sustains oscillation. The frequency is determined by L1, TC1, stray capacity and the collector base capacity of Q1. Now the collector base junction is reverse biased and looks like a variable capacitance diode. Since the amplified audio appears across this diode its capacity, and hence the VHF frequency, will swing in sympathy with the audio input. Output is taken direct from the collector using a short aerial. The frequency of oscillation is set with TC1. The RF power INPUT to the transistor is only about 8mW from a total power to the bug of about 25mW

5 Final Design , Construction and Assembly

5.1 Introduction

This chapter will discuss the final design in detail and give instructions on how it can be built and implemented.

5.2 Final Circuit Design

After considering 4 basic designs, it was concluded that a mixture of the 3rd and 4th design's be implemented for transmitting human speech across the commercial bandwidth (88 MHz to 108MHz). The design had to be portable, low powered and be able to have the capability of transmitting to more than 1 channel.

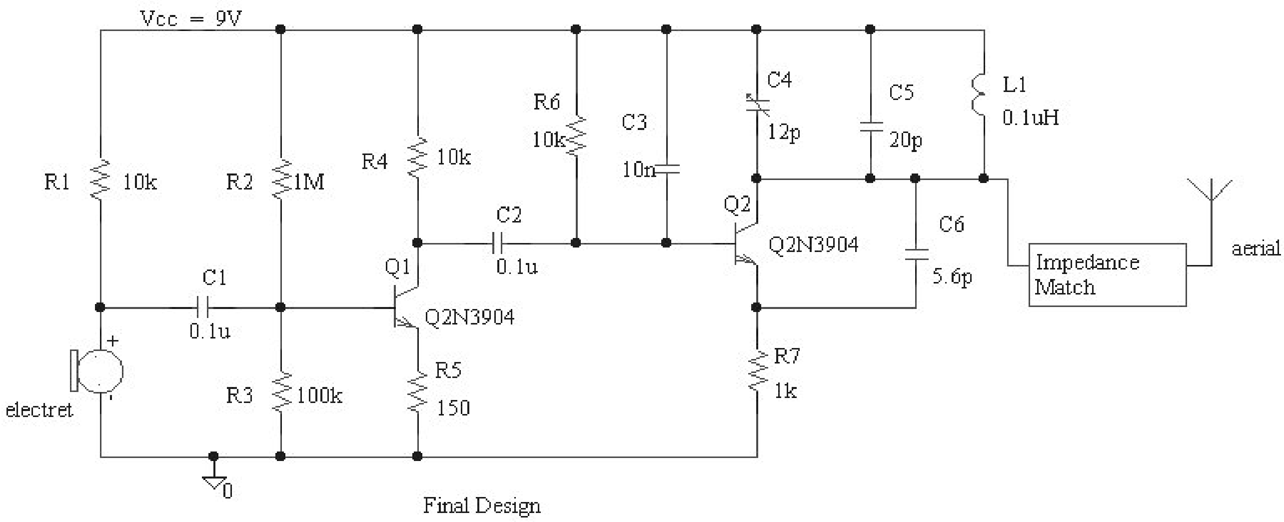


Figure 5.2-1

Although the variable and shunt capacitors C4 and C5 are set up to transmit from 88 to 108Mhz, the transmitter only has an effective tuning range 6 MHz (30 out of the 100 channels) this is due in part to the feedback capacitor C6 being at the right impedance for positive feedback to occur.

5.3 Oscillator analysis

When analysing circuitry with transistor a few theoretical assumptions will have to be made. The main assumption made for the analysis would be the small signal resistance r'e = VT/IE ≈ 20Ω at start of oscillation and rising to about 28Ω after initial oscillations through the system. When the power to the circuit is turned on, unit step is applied to the tank circuit, The capacitor charges up and then releases its charge into the inductor, when the inductor finishes absorbing the charge it's magnetic field will break down and releases the charge back into capacitor and the cycle happens all over again at the resonant frequency of the tank.

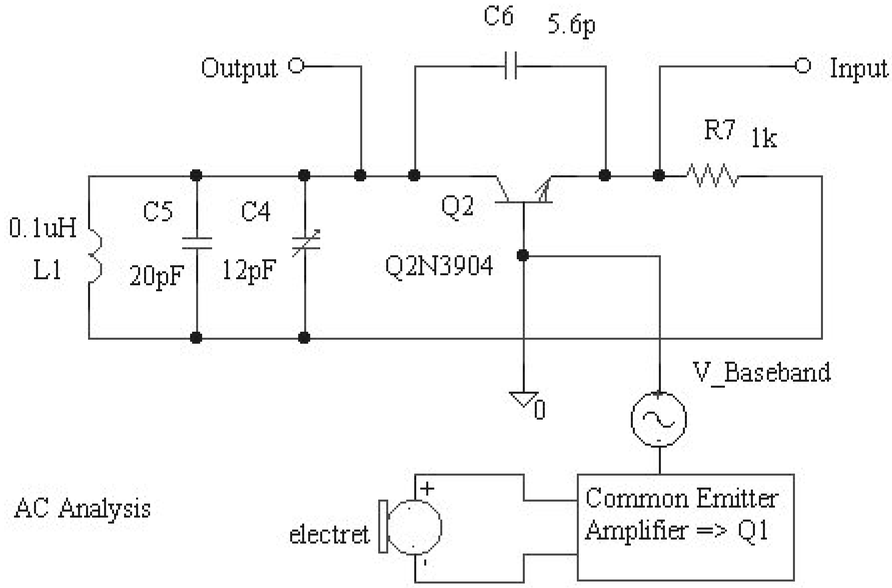


Figure 5.3-1

According to the Barkhausen criteria for sine wave oscillation (section 3.5) of the tank the amplifier (common base, section 2.10.3) and feedback must has a loop gain of unity. Taking 100MHz as the resonant frequency of the tank

L = 0.1µH C4 + C5 = 25.33pF (theoretical value),

- The overall impedance of the tank will be Rc = XL//XC ≈ 31.4
- The amplifier gain (Av) is Rc/r'e = 31.4/20 ≈ 1.25.
- The feedback fraction (B) will be equal to Vin/Vout , consider that Ic ≈ Ie, then B = (RE/ RE+X_{C6}) ≈ .77

- The closed loop gain of the system will therefore equal to $A_L = Av * B = 1.57 * 0.77 = 1.2$, when oscillation is sustained r'e will rise to about 28 and the overall loop gain will fall to unity.

The second Barkhausen Criterion states that in order for sustained oscillation to happen the phase shift through the network at the resonant frequency will have to be zero, L1, C4 and C5 will yield 0^O at the resonant frequency. There is a $- 90^O$ phase shift through C6. The emitter–collector channel has an interesting property when it comes to current and voltage, the current entering the emitter leads the voltage across the collector, hence a $+ 90^O$ phase shift. Putting the capacitor and transistor together there will be a 0^O between the input and output nodes.

5.4 Components List

5.4.1 Resistors

R1	10KΩ	Carbon Film	Bias for the Electret microphone
R2	1MΩ	Carbon Film	DC bias for the Base of Q1
R3	100KΩ	Carbon Film	DC bias for the Base of Q1
R4	150Ω	Carbon Film	Sets the DC & AC gain of Q1
R5	10KΩ	Carbon Film	Sets the DC & AC gain of Q1
R6	10KΩ	Carbon Film	Forms a HPF with C2 (pre-emphasis)
R7	1KΩ	Carbon Film	Sets the gain for the oscillator.

5.4.2 Capacitors

C1	0.1µF	Non-polarised tantalum	Audio coupling capacitor
C2	0.1µF	Non-polarised tantalum	Forms HPF with R6 .
C3	10nF	Ceramic	AC grounds the base of Q2.
C4	1 to 12pF	Silver Mica	Tuning cap for Multichannel
C5	20pF	Ceramic	Shunt capacitor for tuning
C6	5.6pF	Ceramic	Feedback for oscillation.

5.4.3 Inductor

L1	0.1µH	Toko

5.4.4 Transistors

Q1 & Q2	2N3904	TO-92

5.4.5 Microphone

Miniature Electret

Tie Clip Electret microphone

5.4.6 Input - Out connections

Input is provided for an external tie clip Electret microphone, which will disconnect the fixed small Electret microphone. In order to achieve this a 3.5 mm jack is used which has a multiplex feature for switching between different references, which is dependant on the socket being empty or filled by the 3.5mm jack.

5.5 Construction and assembly

One of the most versatile properties of this design is this design is that the parts are very easily obtained, the circuit can even be built on ordinary vero-board. One thing that had to be observed was to keep the leads of the devices small and compact the circuitry. With a simple carbon film 70Ω resistor incorporated into the straight wire antenna.The PCB design sank 8.37mA at 9v dc battery and yielded an output power of 8mW.Type of power source used was a Duracell Alkaline Manganese Dioxide 9volt PP3 battery. The duration for effective transmission is 14 Hours of continuous use.

5.5.1 Pcb Layout

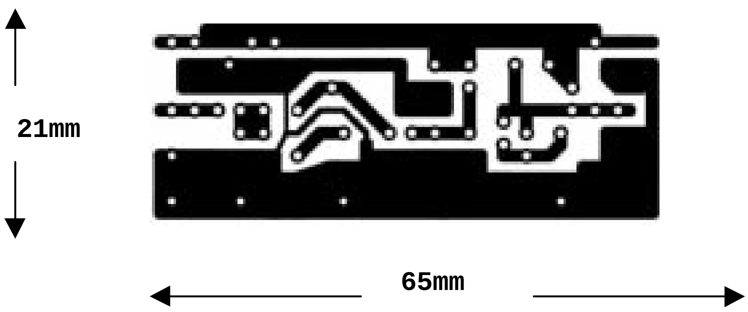


Figure 5.5-1

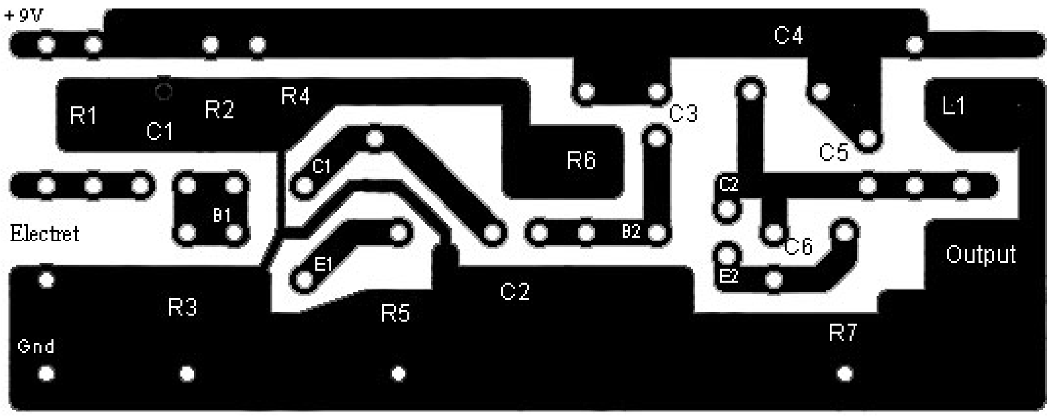


Figure 5.5-2

The PCB design above was designed using easy-pc a dos based PCB design package. Figure 5.5-2 shows the under side of the double sided board, most of the underside is devoted to a ground plane, the topside (component side) is totally devoted to a ground plane. A foil isn't needed for the component side, all that needs to be done is drill the under side and use a track cutter on the component side to enlarge the spacing from hole edge to the ground plane. Figure 5.5-3 displays where the components fit in on the board.

Note: the capacitor C4 is mounted on to the PCB on it's side to allow for access to the tuning circuitry. Also the RF section (between C3 and just before the output is coated with household cling film and then wrapped with aluminium cooking foil. This will prevent any stray signal feedback from interfering with modulation.

The PCB board slots into a handheld instrumentation case 90*65*25(obtained from RS), a total of 4 holes were drilled, two on top (for the miniature electret and a 3.5mm socket for antenna), one at the side (for a switch) and one in the front (for tuning the capacitor).

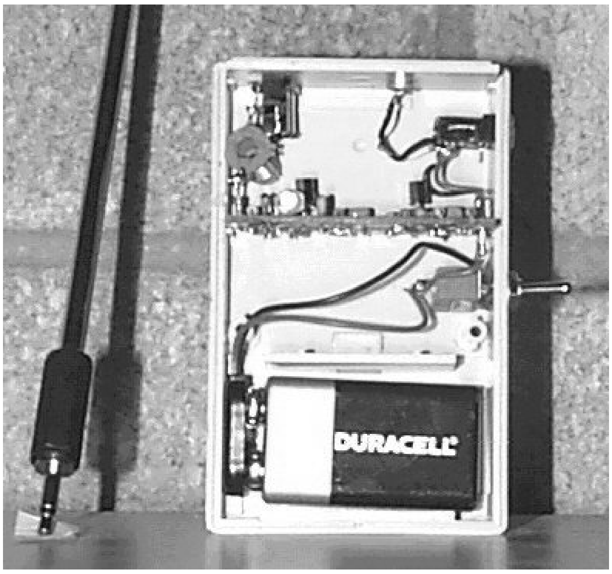


Figure 5.5-1

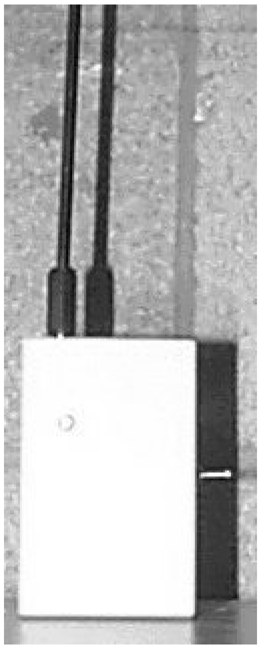


Figure 5.5-2

Figure 5.5-1 & 2 shows the assembled transmitter.

5.6 Antenna Considerations

The antenna used for the project was an end fed whip antenna with a fully extended length of 75cm. From section 3.11.1 which dealt about radiation resistance of an antenna. Taking again the frequency to be about 100Mhz, which yields a wavelength of 3 metres, the radiation resistance is calculated to be about R_L=4.6KΩ. The output impedance of the oscillator is seen as the impedance of the tank in parallel with the impedance of the feedback capacitor plus the bias resistor R7 which at 100MHz works out to be Ro = 30Ω. To match the output with the input impedance of the antenna a simple shunt resistor of 4.570KΩ maybe placed between the output and the input to the antenna. Or a more elaborate scheme maybe employed by using an LC bandpass filter to make the antenna look like it has the same impedance as the output of the transmitter (see section 3.12).

Q can be calculated as 12, X_L as 370 and Xc as 373. At a frequency of 100MHz XL is equal to 0.589μH and Xc is equal to 4.2pF. The network's bandwidth can be calculated as frequency over quality factor, which is calculated as 8.3Mhz. This quite a sophisticated way of matching the load, but it does have it's downside, especially when the transmitter is multi-channelled, is that the frequency is not a constant. Taking the frequency as 106MHz, the inductance will have to be 0.55μH and the capacitance as 4.02pF. So choosing an inductance of 0.6μH and a capacitance of 4.1pF could possibly match the transmitter with the antenna.

The method chosen for matching the impedance of the antenna was the resistor placed between the output node and the antenna. This yielded a respectable range of 80 feet in a household environment and about 50 feet inside a lab.

An extension to the antenna was discussed and implemented using a thin coaxial cable with it's outer conductor grounded to the board, which tended to change the centre frequency of the transmitter.

5.7 Overall frequency of the transmitter

The frequency stability of due to ageing effects and the non-zero temperature coefficients of the components (see section 2.7) tends to vary the frequency of the transmitter, so mild adjust of the receiver is called for every so often, this will be ok for analogue FM receivers, but for digital receivers (which are slowly becoming popular) this can be quite tedious as was found during testing.

6 Test and Results

6.1 Introduction

This section will discuss some of the more detailed tests carried out on the final circuit which was discussed in section 5. Graphs and pictures will be used to aid in the final analysis of the Design.

6.2 Equipment used

The equipment used in analysing circuitry is vital in yielding the correct information about the advantages and disadvantages of any design. During the course of final test the equipment used were a spectrum analyser, a frequency meter, digital multi-meter, an analogue and digital FM radio receiver was used.

6.2.1 Spectrum Analyser

A spectrum analyser is exactly what its name implies, it shows the frequency response over a specified width in the frequency domain. The spectrum analyser that was used, was the MS610C from Antristu. The Antristu has a dynamic range of 9Khz to 2 GHz. The spectrum analyser was used to view the varying effects of the carrier when it was modulated by the baseband audio signal. The signal strength was also measured using the analyser, using a conversion formula that will be shown later.

6.2.2 Frequency Meter

The frequency meter used was the GX240 from ITT instruments. It has a maximum frequency of 120MHz. This was used to check the output frequency of the oscillator. To effectively use it, an impedance of 20Ω was attached to the meter's probe and placed at the output of the transmitter, this was used to properly match the 50Ω transmission line of the probe with that of the transmitter output.

6.2.3 Radio Receiver

An analogue (dial turn) and a digital (push-button) receiver was used in demodulating the modulated carrier wave generated by the transmitter.

6.3 Spectrum Analyser test

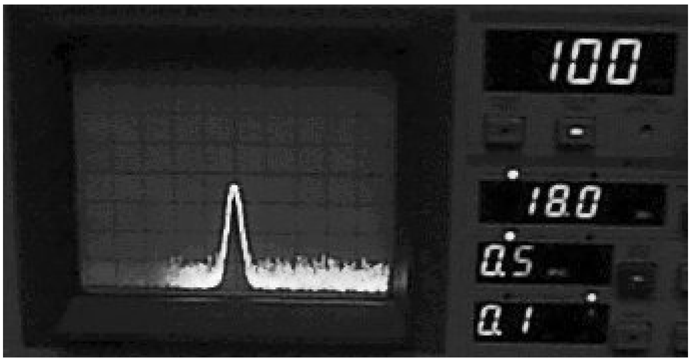


Figure 6.3-1

Figure 6.3-1 shows the unmodulated carrier spectrum at 100MHz

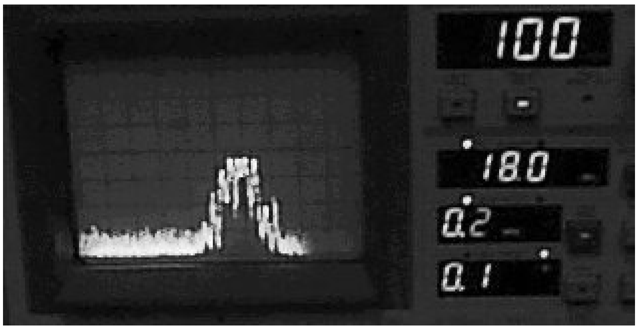


Figure 6.3-2

Figure 6.3-2 show's the spectrum of the carrier being modulated by a baseband audio signal.

6.4 Power Output

The y-axis of the spectrum analyser measures the power of the output wave in milli-decibels, the formulae for converting either way from mill-watts to milli-decibels are given below

$$P_{dBm} = 10.\log_{10}(P_{mW}) \qquad P_{mW} = 10^{\frac{P_{dBm}}{10}}$$

All measures were taken with a 20 Ω impedance onto the end of the probe to ensure the matching of the 50Ω transmission line to the output impedance of the oscillator.

Supply Voltage (Volts)	Supply Current (mA)	Output Power (dBm)	Output Power (mW)
9	8.37	8.1	6.456542
8	7.31	7.1	5.128614
7	6.24	6	3.981072
6	5.23	3.5	2.238721
5	4.27	1.88	1.5417
4	3.33	-1.42	0.721107

Table 6.4-1

Table 6.4-1 shows the supply current and output power as a result of the dc supply of the PP3 battery decreased during continuous operation over 18 hours.

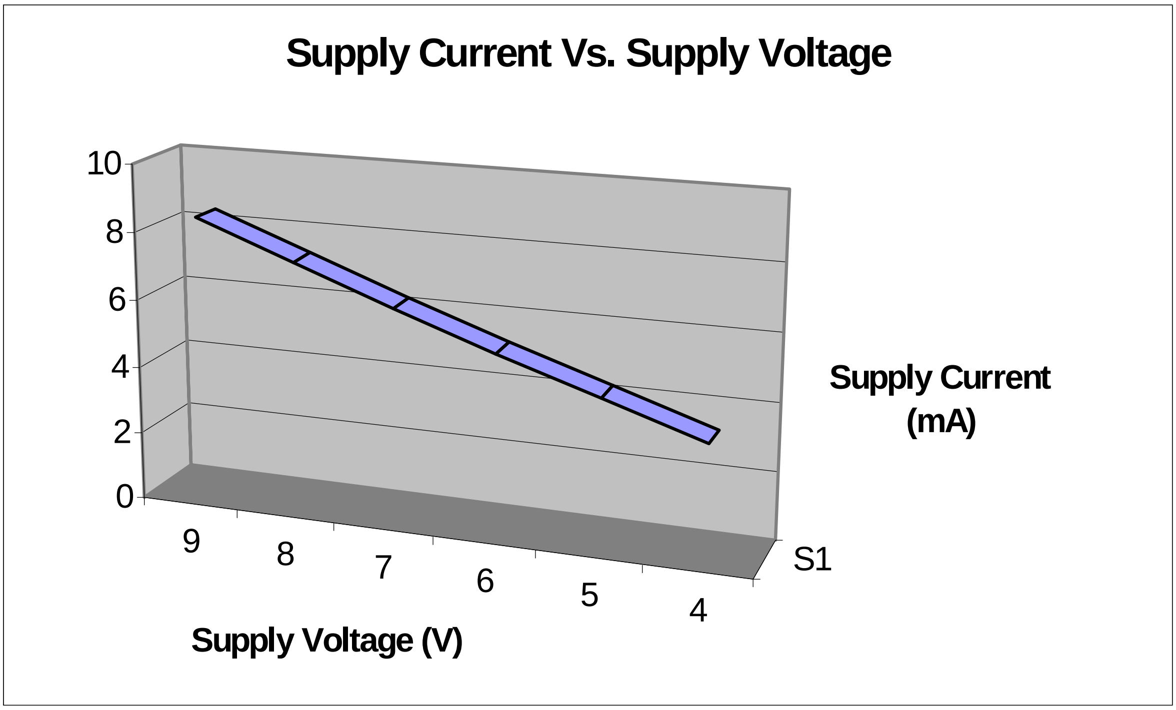


Figure 6.4-1

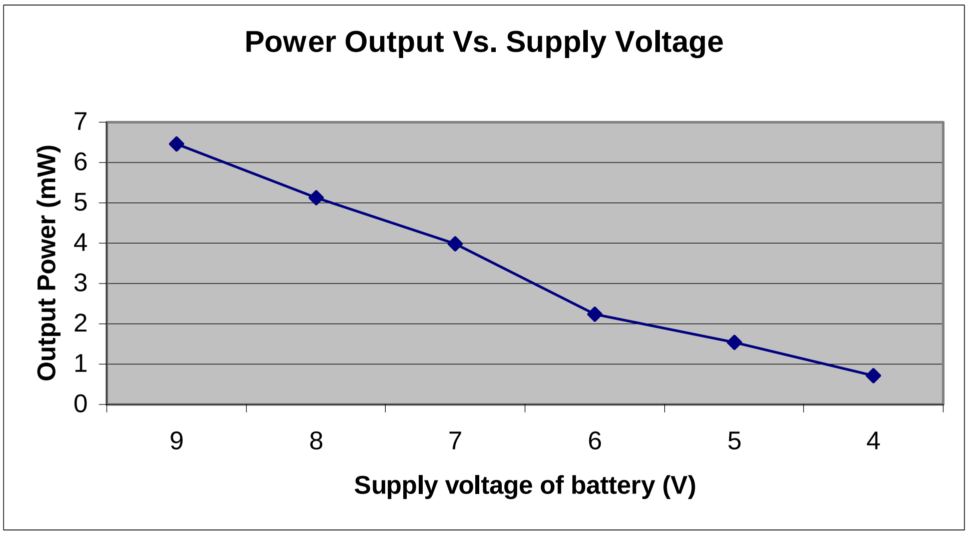


Figure 6.4-2

It can be easily seen from figure 6.4-1 & 2 that the supply current decreases linearly as the supply voltage supplied by the battery decreases, also the output power is decreases as voltage supply to the transmitter decreases. The typical voltage input from the battery source will be 9.2V in the first 30 minutes of operation decreasing to the 8.8volts to 7 volts band in the next 14 hours of continuous operation and decreasing to the 7 to 4 volts band of operation in the following 6 hours. This last band of voltage operation produced fairly low levels of transmission. The Middle band 7 to 8.8 volts emitted a very strong transmission indeed, which lasted for 14 hours, this is the main feature that make the PP3 such a popular battery for intensive dc supply to "walkie talkie's", because of this a stable transmission over a long period of time can be sustained.

7 Final Discussion and Conclusions

7.1 Introduction

The fundamental aim of this project was to design a miniaturised low powered, multi-channel FM transmitter for modulation across the commercial Band. This final section shall begin with a detailed discussion of the main topics from initial approach to final design and implementation. Also in this section a number of conclusions shall be drawn from the approach that I have taken and finally recommendations on how what should have been done.

7.2 Report Overview

When considering a design for frequency modulation, a number of key elements have to be considered, such as a good understanding of the concept of modulation schemes and the electronic circuitry that goes into creating the scheme.

In sections 1 and 2 of this report the theory of frequency modulation was covered, then a broad overview of electronic components and their various properties were considered. Section 3 introduced the various building blocks and considered their possible use in the final design. In section 4 a quantitative overview of some of the possible designs that were considered in the progress of the project was given. A detailed description of how each design works is given here, along with why the design was or was not chosen. The final design was discussed in full, in section 5 along with details of construction and assembly, pictures of the final circuit enclosed in a handheld plastic casing are shown here to give the reader of the report a general feel for what the final transmitter looks like.

Finally Section 6 gives the results of the various tests used on the design

7.3 Discussion

The design chosen was miniature, low powered and tuneable to different frequencies. The parts used are very common and the circuit is very easily constructed. The circuit was firsts build on vero-board and worked rather well without applying any real effective RF techniques, all that had to be adhered to was to keep the leads short and compact the circuitry as much as possible. The PCB, as expected performed exceedingly well, but more of an better attempt had to be made in matching the antenna (section: 3.12 & 5.6) and shielding the RF section from the output as the PCB layout was a lot more efficient in radiating power out. Unwanted Electro-magnetic radiation had to be stopped from destructively interfering with the carrier modulation. To keep the design simple and easy to construct it was decided to just wrap household cling-film, electrically protecting the circuitry from the Aluminium foil that was used to electro-magnetically shield the RF stage.

The effective range of the transmitter was 80 feet in a household environment and about 50 feet in a lab environment, which makes the transmitter ideal aiding Tourists hear a Tour Guide within a densely packed room. The Antenna co-axial extension lead helps in raising the antenna over people's heads and transmitting to their FM receivers (otherwise the water content in the human body would bounce the electromagnetic waves). Because of the 14 hours of effective transmission given by the PP3 and the power consumption of the battery, the device could be used as a baby monitoring device within the home, with a range of about 80 feet, it reaches out even to the garden.

7.4 Conclusions

The Miniaturised FM transmitter is essentially a Design and Implementation project. To approach a project like this a parallel path has to be taken in regards to the Theory and the practical circuitry, for a successful conclusion in any project these path's must meet, and this only happens when they are fully understood. This is why a good grounding in the basics of Communication theory and Analogue design must be achieved before ever approaching a project like this. To start off looking at block diagrams of basic transmitter was a must, even if it seemed abstract and obscure the underlying meaning of each block can be found out one by one. Which is what made the overall project challenging and rewarding.

7.5 Recommendations

The design used for this project is essentially quite a simple one, and it is this simplicity which partly brings it down when it comes to the overall reliable performance. The main area of instability is in the oscillator part of the circuit. Shielding (section 5.5) the oscillator helped in part to counteract this.

After learning a lot from this project, there would have been a few things that could have been done to the final design to improve it's performance.

- Use negative temperature coefficients to compensate for typically positive-temperature-coefficient tuned circuits.
- Follow the oscillator with a buffer amplifier to reduce the effects of load changes.

The standalone Vco design of section 4.3, is worth a look at, maybe if the MC1648 soic 4 pin package was used in conjunction with other stages as mentioned in section 4.3, the stability of the oscillator could have been greatly increased.

References

1. Modern electronic communication / Miller, Gary M
2. Electronic communications: modulation and transmission / Schoenbeck, Robert
3. Electronics : circuits, amplifiers, and gates / Bugg, D. V. (David Vernon)
4. Antennas / Connor, Frank Robert
5. Fundamentals of reliable circuit design / Xlander, Mel
6. Electronic devices / Floyd, Thomas L
7. Electronic Communication Techniques / Young, Paul H.

Appendix A

$t := 0, .001 .. 1$ $\quad Ko := 4$ $\quad Vpk := 3$ $\quad fm := 2$ $\quad fc := 20$

$Kp := 2$ $\quad f := 1, 1.001 .. 30$

$\pi := 3.14$

$A := 2$ $\quad Vm(t) := (Vpk \cdot \cos(2 \cdot \pi \cdot fm \cdot t))$ $\quad Vc(t) := A \cdot \cos(2 \cdot \pi \cdot fc \cdot t)$

Modulation Index

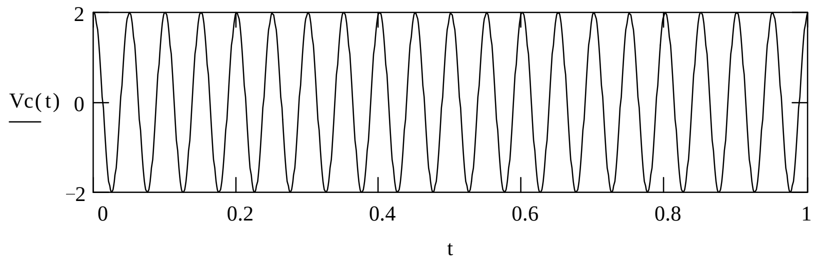
$mf := Ko \cdot \frac{Vpk}{fm}$ Index for FM $\qquad mp := Kp \cdot Vpk$ Index for PM

$$S(t) := A \cdot \cos\left(2 \cdot \pi \cdot fc \cdot t + 2 \cdot \pi \cdot Ko \cdot \int_0^t Vm(t)\, dt\right)$$

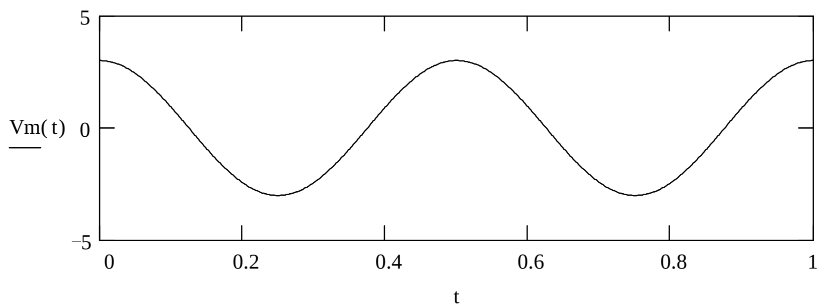
S(t) & S1(t) are the same frequency modulated waves only represented differently, i.e. S(t) shows the Information wave being integrated.

$$S1(t) := A \cdot \cos(2 \cdot \pi \cdot fc \cdot t + mf \cdot \sin(2 \cdot \pi \cdot fm \cdot t))$$

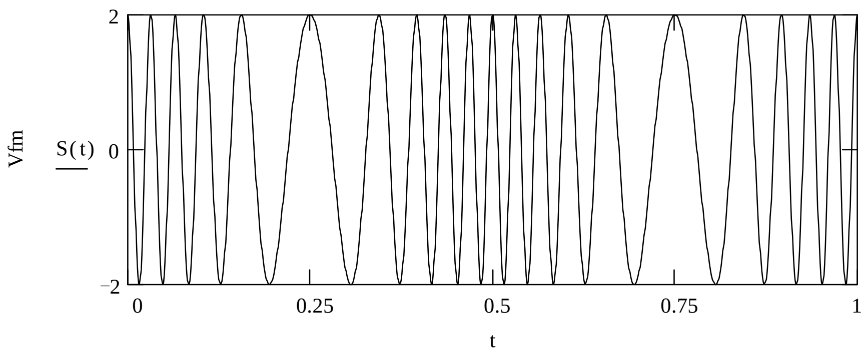
$$Vpm(t) := A \cdot \cos(2 \cdot \pi \cdot fc \cdot t + Kp \cdot Vm(t))$$



Carrier Wave

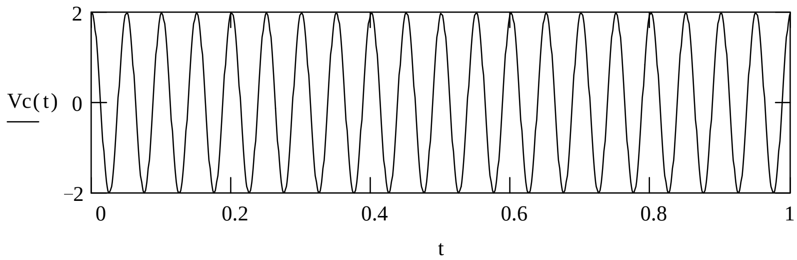


Wave Conataining Information

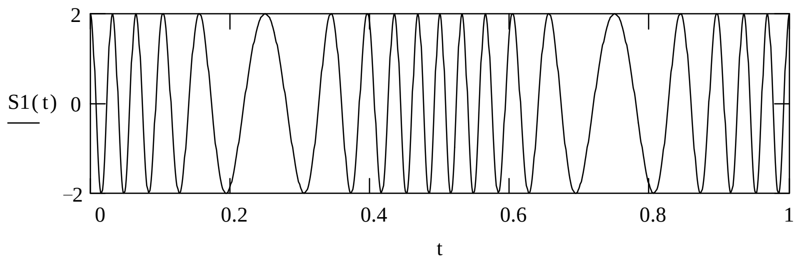


Frequency Modulated Wave

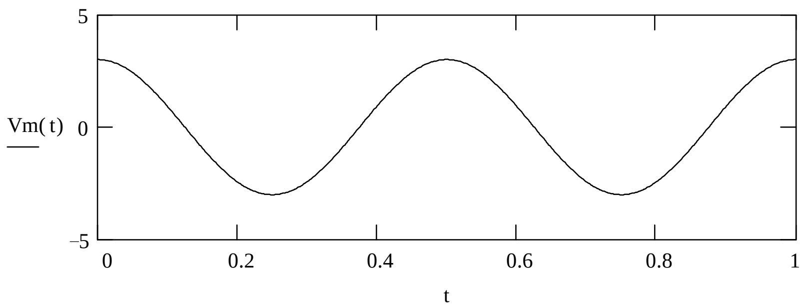
Graphs comparing the Phase & Frequency modulation schemes, along side the Carrier & Information Waves.



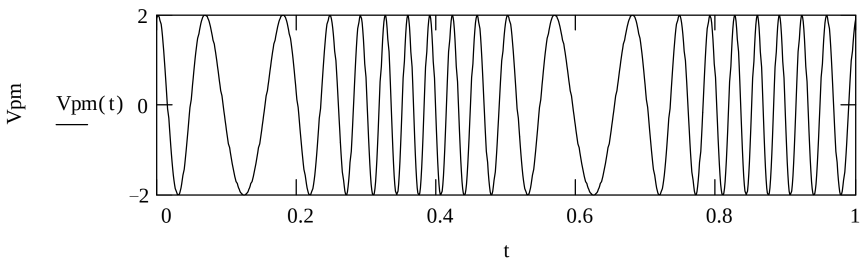
Carrier Wave



Frequency Modulated wave



Wave Containing Information

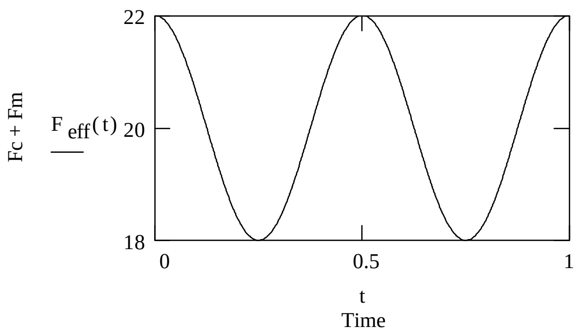


Phase Modulated Wave

This page is concerned with Demodulation !
Which means graphing the frequency change with respect to time.

$$T := \frac{1}{fm}$$

$$F_{eff}(t) := fc + fm \cdot \cos\left(\frac{2 \cdot \pi}{T} \cdot t\right)$$



Frequency versus Time

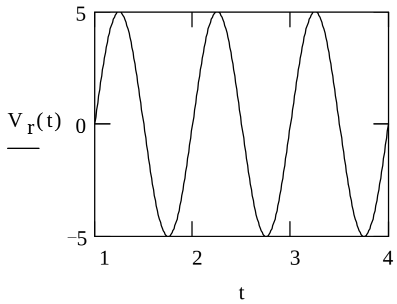
$$N := 8$$

Using Mathcad to enter equations for voltage & current across a resistor, capacitor & inductor. And to Create Elementary Graphs showing the voltage and current graphs for each of these devices.

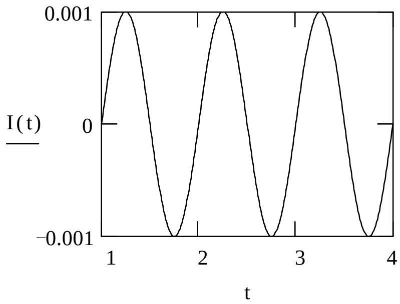
$t := 1, 1.01 .. 4 \qquad f := \frac{1}{2} \qquad I(t) := 10^{-3} \cdot \sin(2 \cdot \pi \cdot t)$

For a Resistor : Voltage and Current are in Phase!

$R := 5 \cdot 10^{3} \qquad V_r(t) := I(t) \cdot R$



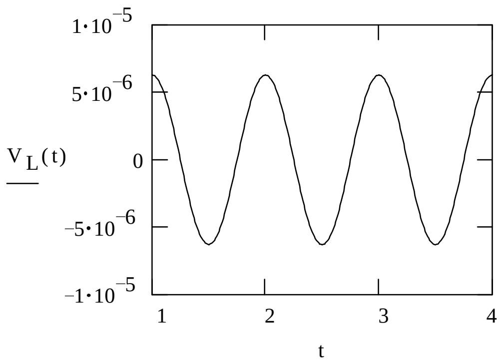
Voltage Through a Resistor



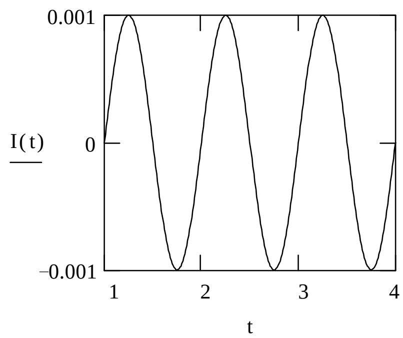
Instantaneous Current

For an Inductor : Voltage leads current by 90 degrees

$L := 1 \cdot 10^{-3} \qquad V_L(t) := L \cdot \frac{d}{dt} I(t)$



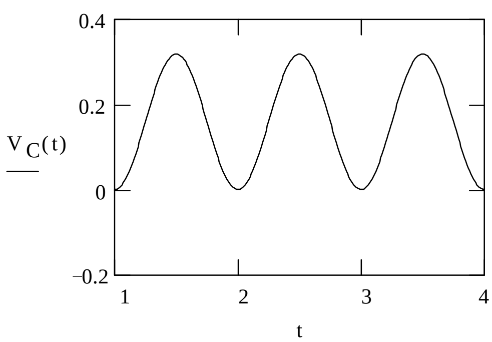
Volatage Through an Inductor



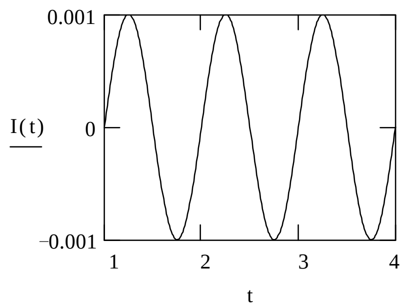
Instantaneous Current

For a Capacitor : Volatge Lags Current by 90 degrees.

$C := 1 \cdot 10^{-3} \qquad V_C(t) := \frac{1}{C} \cdot \int_0^t I(t)\, dt$



Voltage Through a Capitor



Instantaneous Current

The following equations show's the Impedance and Phase plots for a Parallel LCR tank. Note that at resonance (Xc = XL) the impedance is at a Maximum and the Phase is 0 degress

$L := 0.1 \cdot 10^{-6}$ $\qquad$ $C := 25.319 \cdot 10^{-12}$

$R := 100 \cdot 10^{3}$

$$Q := \frac{R}{2 \cdot \pi \cdot f_o \cdot L} \qquad f_o := \frac{1}{2 \cdot \pi \cdot \sqrt{L \cdot C}}$$

$$BW := \frac{f_o}{Q}$$

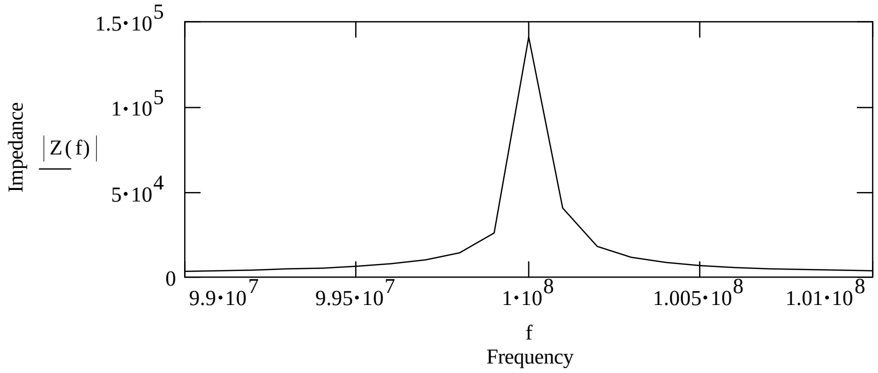
$BW = 6.286 \cdot 10^{4}$

$$Z(f) := \frac{R}{j \cdot Q \cdot \left(\frac{f}{f_o} - \frac{f_o}{f} \right)}$$

$f := 99 \cdot 10^{6}, 99.1 \cdot 10^{6} .. 101 \cdot 10^{6}$

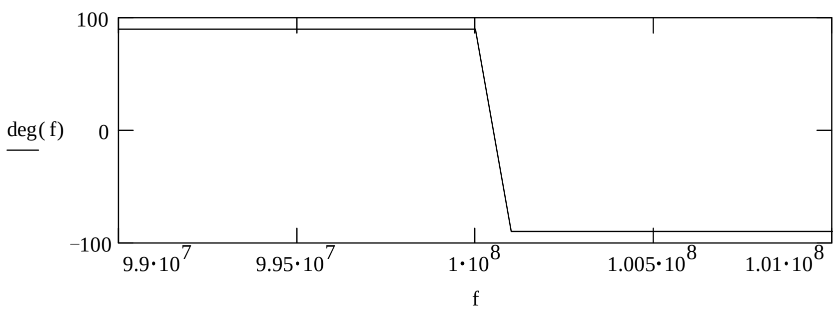
$Q = 1.591 \cdot 10^{3}$

$f_o = 1 \cdot 10^{8}$



Impedance of Parrallel LC tank Vs Freq

$$deg(f) := \left(\frac{180}{\pi} \cdot arg(Z(f)) \right)$$



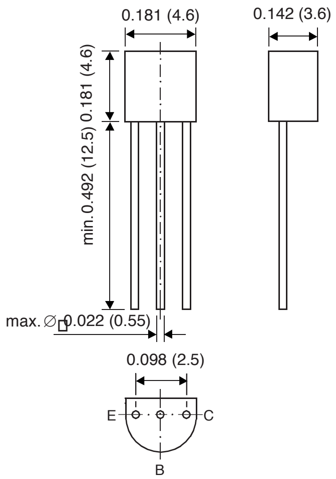
Phase of Parrallel LC Tank

Appendix B

2N3904

SMALL SIGNAL TRANSISTORS (NPN)

TO-92



Dimensions in inches and (millimeters)

FEATURES

- ♦ NPN Silicon Epitaxial Planar Transistor for switching and amplifier applications.
- ♦ As complementary type, the PNP transistor 2N3906 is recommended.
- ♦ On special request, this transistor is also manufactured in the pin configuration TO-18.
- ♦ This transistor is also available in the SOT-23 case with the type designation MMBT3904.

MECHANICAL DATA

Case: TO-92 Plastic Package
Weight: approx. 0.18g

MAXIMUM RATINGS AND ELECTRICAL CHARACTERISTICS

Ratings at 25°C ambient temperature unless otherwise specified

	SYMBOL	VALUE	UNIT
Collector-Base Voltage	V_{CBO}	60	V
Collector-Emitter Voltage	V_{CEO}	40	V
Emitter-Base Voltage	V_{EBO}	6.0	V
Collector Current	I_C	200	mA
Power Dissipation at $T_A = 25°C$ at $T_C = 25°C$	P_{tot}	625 1.5	mW W
Thermal Resistance Junction to Ambient Air	$R_{\theta JA}$	250[(1)]	°C/W
Junction Temperature	T_j	150	°C
Storage Temperature Range	T_S	– 65 to +150	°C

NOTES:
(1) Valid provided that leads are kept at ambient temperature.

2N3904

ELECTRICAL CHARACTERISTICS

Ratings at 25°C ambient temperature unless otherwise specified

	SYMBOL	MIN.	MAX.	UNIT
Collector-Base Breakdown Voltage at I_C = 10 μA, I_E = 0	$V_{(BR)CBO}$	60	–	V
Collector-Emitter Breakdown Voltage at I_C = 1 mA, I_B = 0	$V_{(BR)CEO}$	40	–	V
Emitter-Base Breakdown Voltage at I_E = 10 μA, I_C = 0	$V_{(BR)EBO}$	6	–	V
Collector Saturation Voltage at I_C = 10 mA, I_B = 1 mA at I_C = 50 mA, I_B = 5 mA	V_{CEsat} V_{CEsat}	– –	0.2 0.3	V V
Base Saturation Voltage at I_C = 10 mA, I_B = 1 mA at I_C = 50 mA, I_B = 5 mA	V_{BEsat} V_{BEsat}	– –	0.85 0.95	V V
Collector-Emitter Cutoff Current V_{EB} = 3 V, V_{CE} = 30 V	I_{CEV}	–	50	nA
Emitter-Base Cutoff Current V_{EB} = 3 V, V_{CE} = 30 V	I_{EBV}	–	50	nA
DC Current Gain at V_{CE} = 1 V, I_C = 0.1 mA at V_{CE} = 1 V, I_C = 1 mA at V_{CE} = 1 V, I_C = 10 mA at V_{CE} = 1 V, I_C = 50 mA at V_{CE} = 1 V, I_C = 100 mA	h_{FE} h_{FE} h_{FE} h_{FE} h_{FE}	40 70 100 60 30	– – 300 – –	– – – – –
Input Impedance at V_{CE} = 10 V, I_C = 1 mA, f = 1 kHz	h_{ie}	1	10	kΩ
Voltage Feedback Ratio at V_{CE} = 10 V, I_C = 1 mA, f = 1 kHz	h_{re}	$0.5 \cdot 10^{-4}$	$8 \cdot 10^{-4}$	–
Gain-Bandwidth Product at V_{CE} = 20 V, I_C = 10 mA, f = 100 MHz	f_T	300	–	MHz
Collector-Base Capacitance at V_{CB} = 5 V, f = 100 kHz	C_{CBO}	–	4	pF
Emitter-Base Capacitance at V_{EB} = 0.5 V, f = 100 kHz	C_{EBO}	–	8	pF

2N3904

ELECTRICAL CHARACTERISTICS

Ratings at 25°C ambient temperature unless otherwise specified

	SYMBOL	MIN.	MAX.	UNIT
Small Signal Current Gain at V_{CE} = 10 V, I_C = 1 mA, f = 1 kHz	h_{fe}	100	400	–
Output Admittance at V_{CE} = 1 V, I_C = 1 mA, f = 1 kHz	h_{oe}	1	40	µS
Noise Figure at V_{CE} = 5 V, I_C = 100 µA, R_G = 1 kΩ, f = 10 … 15000 Hz	NF	–	5	dB
Delay Time (see Fig. 1) at I_{B1} = 1 mA, I_C = 10 mA	t_d	–	35	ns
Rise Time (see Fig. 1) at I_{B1} = 1 mA, I_C = 10 mA	t_r	–	35	ns
Storage Time (see Fig. 2) at $–I_{B1}$ = I_{B2} = 1 mA, I_C = 10 mA	t_s	–	200	ns
Fall Time (see Fig. 2) at $–I_{B1}$ = I_{B2} = 1 mA, I_C = 10 mA	t_f	–	50	ns

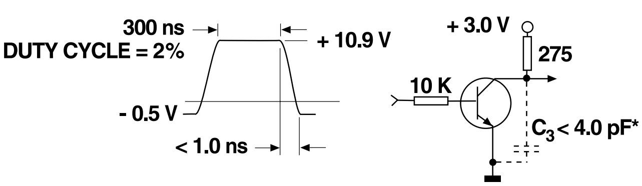


Fig. 1: Test circuit for delay and rise time
* total shunt capacitance of test jig and connectors

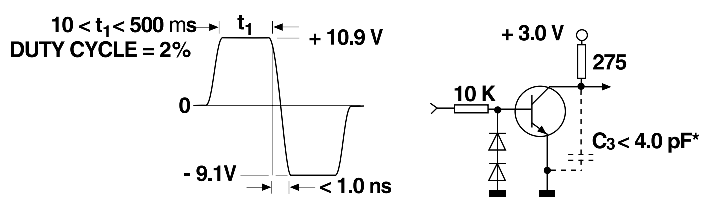


Fig. 2: Test circuit for storage and fall time
* total shunt capacitance of test jig and connectors

Appendix C

Voltage Controlled Oscillator

Consider MC12148 for New Designs

MC1648

VOLTAGE CONTROLLED OSCILLATOR

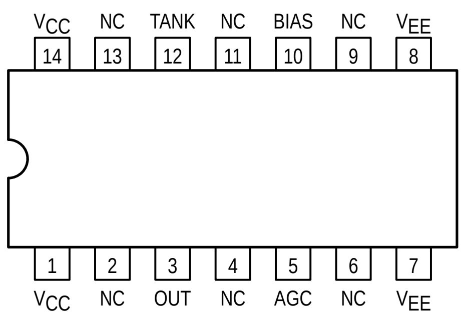
The MC1648 requires an external parallel tank circuit consisting of the inductor (L) and capacitor (C). **For Maximum Performance $Q_L \geq 100$ at Frequency of Operation.**

A varactor diode may be incorporated into the tank circuit to provide a voltage variable input for the oscillator (VCO). The MC1648 was designed for use in the Motorola Phase–Locked Loop shown in Figure 9. This device may also be used in many other applications requiring a fixed or variable frequency clock source of high spectral purity. (See Figure 2)

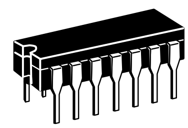
The MC1648 may be operated from a +5.0Vdc supply or a –5.2Vdc supply, depending upon system requirements.

NOTE: The MC1648 is NOT useable as a crystal oscillator.

Pinout: 14–Lead Package (Top View)

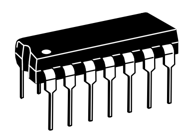


Pin assignment is for Dual–in–Line Package.
For PLCC pin assignment, see the MC1648 Non–Standard Pin Conversion Table below.

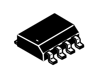


L SUFFIX
14–LEAD CERAMIC PACKAGE
CASE 632–08

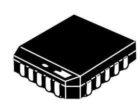
Not Recommended for New Designs



P SUFFIX
14–LEAD PLASTIC PACKAGE
CASE 646–06



D SUFFIX
8–PIN PLASTIC SOIC PACKAGE
CASE 751–05



FN SUFFIX
20–LEAD PLCC PACKAGE
CASE 775–02

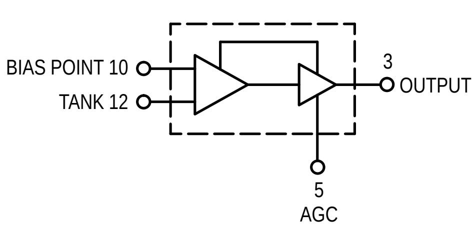
MC1648 NON–STANDARD PIN CONVERSION DATA

Package	TANK	V_{CC}	V_{CC}	OUT	AGC	V_{EE}	V_{EE}	BIAS
8 D	1	2	3	4	5	6	7	8
14 L,P	12	14	1	3	5	7	8	10
20FN	18	20	2	4	8	10	12	14

*NOTE – All unused pins are not connected.

Supply Voltage	GND Pins	Supply Pins
+5.0Vdc	7,8	1,14
–5.2Vdc	1,14	7,8

LOGIC DIAGRAM



- Input Capacitance = 6.0pF (TYP)
- Maximum Series Resistance for L (External Inductance) = 50Ω (TYP)
- Power Dissipation = 150mW (TYP)/Pkg (+5.0Vdc Supply)
- Maximum Output Frequency = 225MHz (TYP)

V_{CC1} = Pin 1
V_{CC2} = Pin 14
V_{EE} = Pin 7

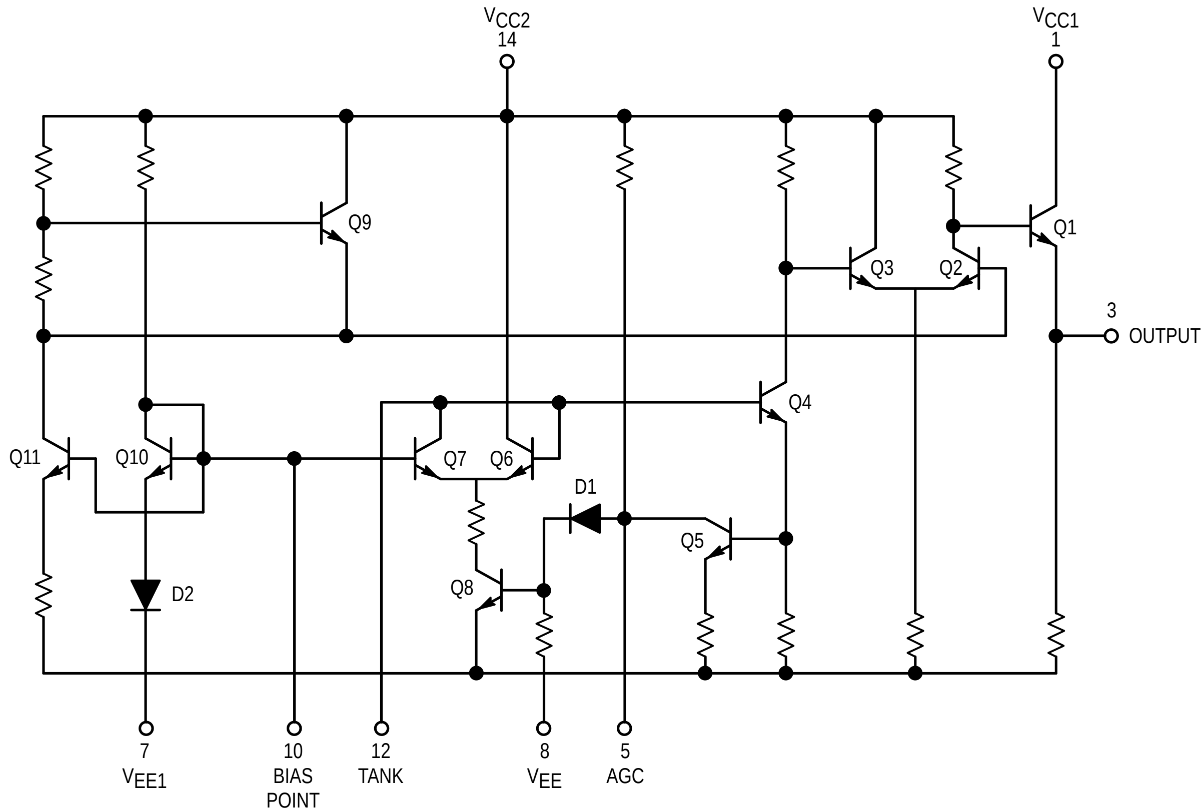


Figure 1. Circuit Schematic

TEST VOLTAGE/CURRENT VALUES

@ Test Temperature	(Volts)			mAdc
	V_{IHmax}	V_{ILmin}	V_{CC}	I_L
MC1648				
–30°C	+2.0	+1.5	+5.0	–5.0
+25°C	+1.85	+1.35	+5.0	–5.0
+85°C	+1.7	+1.2	+5.0	–5.0

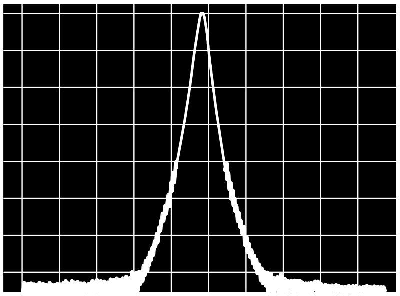
Note: SOIC “D” package guaranteed –30°C to +70°C only

ELECTRICAL CHARACTERISTICS (Supply Voltage = +5.0V)

Symbol	Characteristic	–30°C		+25°C		+85°C		Unit	Condition
		Min	Max	Min	Max	Min	Max		
I_E	Power Supply Drain Current	–	–	–	41	–	–	mAdc	Inputs and outputs open
V_{OH}	Logic “1” Output Voltage	3.955	4.185	4.04	4.25	4.11	4.36	Vdc	V_{ILmin} to Pin 12, I_L @ Pin 3
V_{OL}	Logic “0” Output Voltage	3.16	3.4	3.2	3.43	3.22	3.475	Vdc	V_{IHmax} to Pin 12, I_L @ Pin 3
V_{BIAS}[1]	Bias Voltage	1.6	1.9	1.45	1.75	1.3	1.6	Vdc	V_{ILmin} to Pin 12

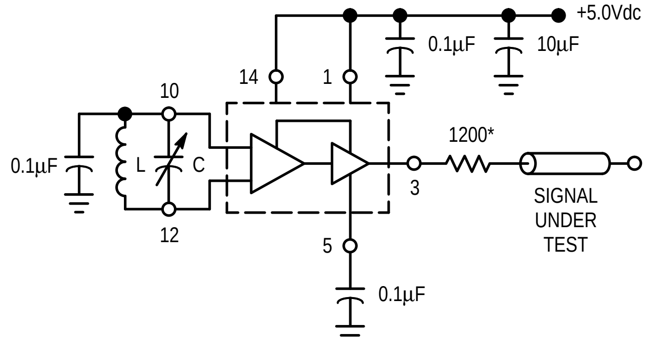
Symbol	Characteristic	Min	Typ	Max	Min	Typ	Max	Min	Typ	Max	Unit	Condition
V_{P-P}	Peak–to–Peak Tank Voltage	–	–	–	–	400	–	–	–	–	mV	See Figure 3
Vdc	Output Duty Cycle	–	–	–	–	50	–	–	–	–	%	
f_{max}[2]	Oscillation Frequency	–	225	–	200	225	–	–	225	–	MHz	

1. This measurement guarantees the dc potential at the bias point for purposes of incorporating a varactor tuning diode at this point.
2. Frequency variation over temperature is a direct function of the ΔC/Δ Temperature and ΔL/Δ Temperature.



B.W. = 10 kHz
Center Frequency = 100 MHz
Scan Width = 50 kHz/div
Vertical Scale = 10 dB/div

L: Micro Metal toroid #T20–22, 8 turns #30 Enameled Copper wire.
C = 3.0–35pF



* The 1200 ohm resistor and the scope termination impedance constitute a 25:1 attenuator probe. Coax shall be CT–075–50 or equivalent.

Figure 2. Spectral Purity of Signal Output for 200MHz Testing

TEST VOLTAGE/CURRENT VALUES

@ Test Temperature	(Volts) V_{IHmax}	(Volts) V_{ILmin}	(Volts) V_{EE}	mAdc I_L
MC1648				
–30°C	–3.2	–3.7	–5.2	–5.0
+25°C	–3.35	–3.85	–5.2	–5.0
+85°C	–3.5	–4.0	–5.2	–5.0

Note: SOIC "D" package guaranteed –30°C to +70°C only

ELECTRICAL CHARACTERISTICS (Supply Voltage = –5.2V)

Symbol	Characteristic	–30°C Min	–30°C Max	+25°C Min	+25°C Max	+85°C Min	+85°C Max	Unit	Condition
I_E	Power Supply Drain Current	–	–	–	41	–	–	mAdc	Inputs and outputs open
V_{OH}	Logic "1" Output Voltage	–1.045	–0.815	–0.96	–0.75	–0.89	–0.64	Vdc	V_{ILmin} to Pin 12, I_L @ Pin 3
V_{OL}	Logic "0" Output Voltage	–1.89	–1.65	–1.85	–1.62	–1.83	–1.575	Vdc	V_{IHmax} to Pin 12, I_L @ Pin 3
V_{BIAS}[1]	Bias Voltage	–3.6	–3.3	–3.75	–3.45	–3.9	–3.6	Vdc	V_{ILmin} to Pin 12

Symbol	Characteristic	–30°C Min	–30°C Typ	–30°C Max	+25°C Min	+25°C Typ	+25°C Max	+85°C Min	+85°C Typ	+85°C Max	Unit	Condition
$V_{P–P}$	Peak–to–Peak Tank Voltage	–	–	–	–	400	–	–	–	–	mV	See Figure 3
Vdc	Output Duty Cycle	–	–	–	–	50	–	–	–	–	%	
f_{max}[2]	Oscillation Frequency	–	225	–	200	225	–	–	225	–	MHz	

1. This measurement guarantees the dc potential at the bias point for purposes of incorporating a varactor tuning diode at this point.
2. Frequency variation over temperature is a direct function of the ΔC/Δ Temperature and ΔL/Δ Temperature.

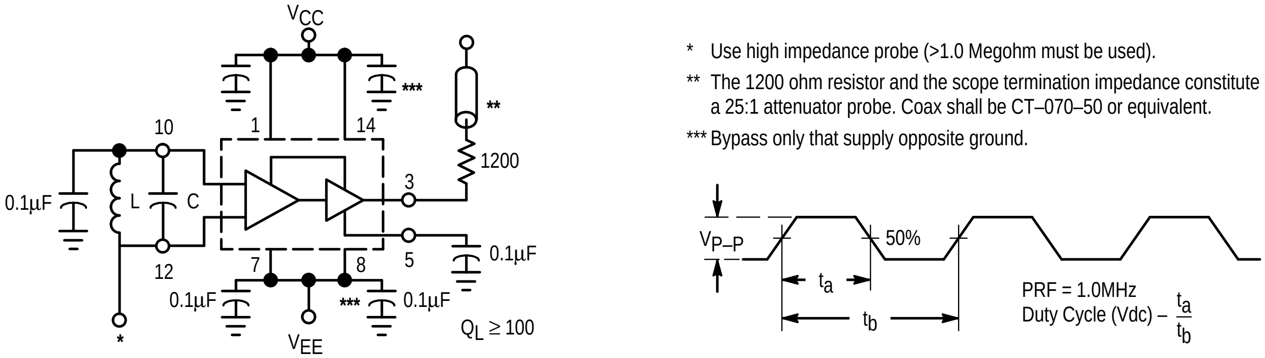


Figure 3. Test Circuit and Waveforms

OPERATING CHARACTERISTICS

Figure 1 illustrates the circuit schematic for the MC1648. The oscillator incorporates positive feedback by coupling the base of transistor Q6 to the collector of Q7. An automatic gain control (AGC) is incorporated to limit the current through the emitter–coupled pair of transistors (Q7 and Q6) and allow optimum frequency response of the oscillator.

In order to maintain the high Q of the oscillator, and provide high spectral purity at the output, transistor Q4 is used to translate the oscillator signal to the output differential pair Q2 and Q3. Q2 and Q3, in conjunction with output transistor Q1, provides a highly buffered output which produces a square wave. Transistors Q9 and Q11 provide the bias drive for the oscillator and output buffer. Figure 2 indicates the high spectral purity of the oscillator output (pin 3).

When operating the oscillator in the voltage controlled mode (Figure 4), it should be noted that the cathode of the varactor diode (D) should be biased at least "2" V_{BE} above V_{EE} (≈1.4V for positive supply operation).

When the MC1648 is used with a constant dc voltage to the varactor diode, the output frequency will vary slightly because of internal noise. This variation is plotted versus operating frequency in Figure 5.

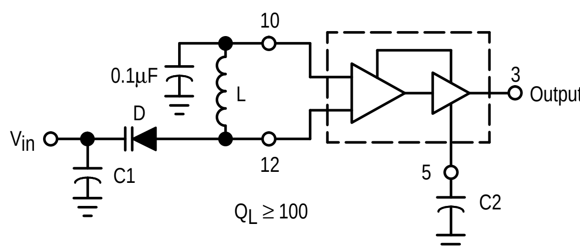
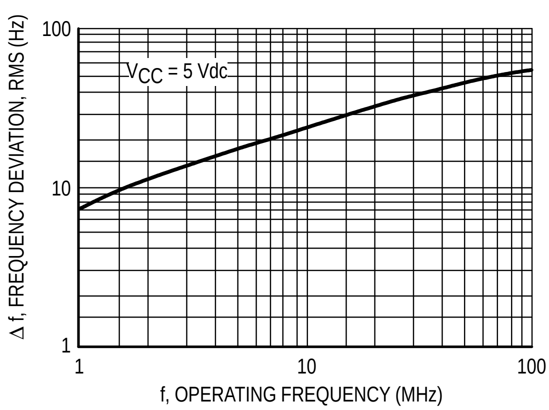
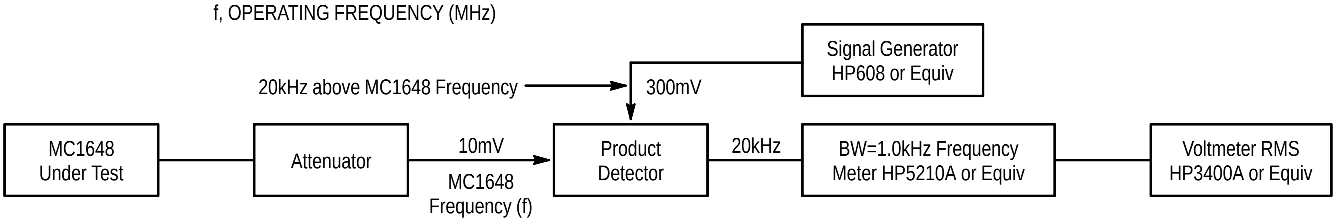


Figure 4. The MC1648 Operating in the Voltage Controlled Mode

Oscillator Tank Components (Circuit of Figure 4)

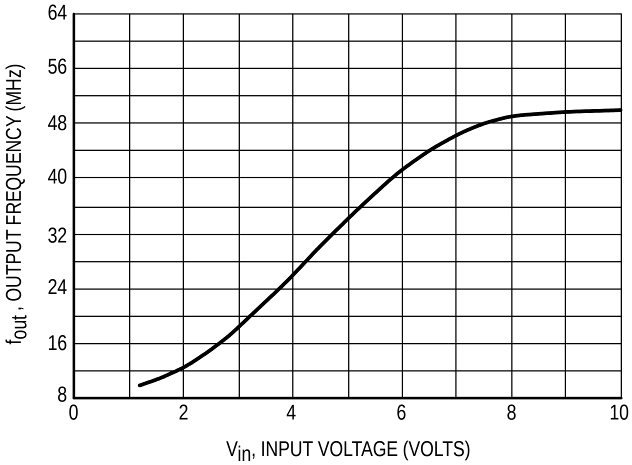
f MHz	D	L µH
1.0–10	MV2115	100
10–60	MV2115	2.3
60–100	MV2106	0.15



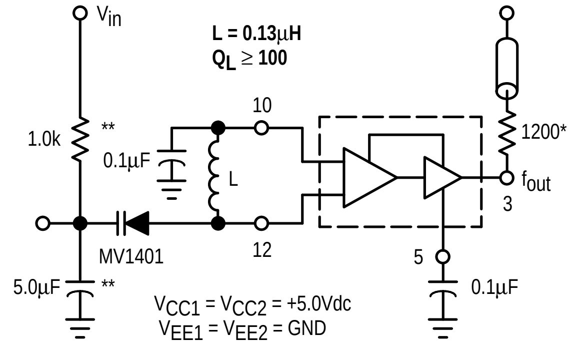


$$\text{Frequency Deviation} = \frac{\text{(HP5210A output voltage) (Full Scale Frequency)}}{\text{1.0Volt}}$$

Figure 5. Noise Deviation Test Circuit and Waveform



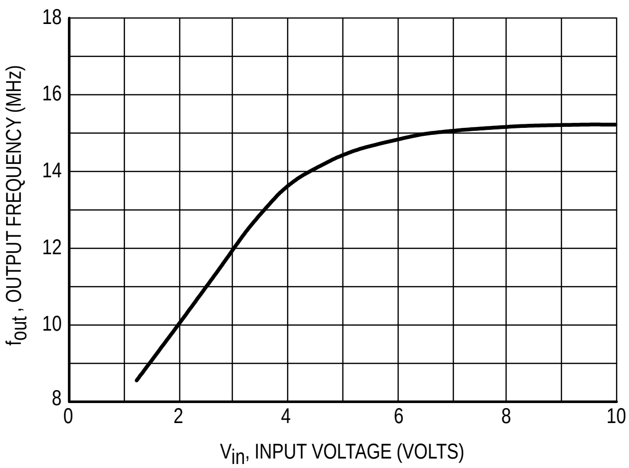
L: Micro Metal Toroidal Core #T44–10, 4 turns of No. 22 copper wire.



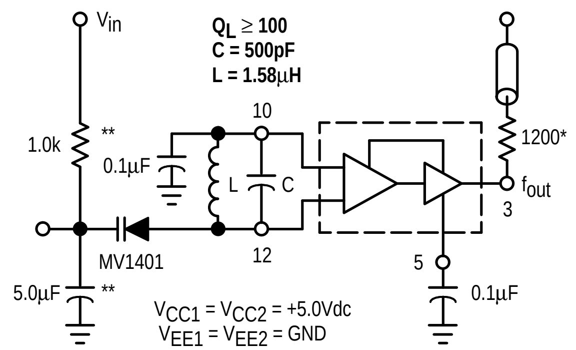
* The 1200 ohm resistor and the scope termination impedance constitute a 25:1 attenuator probe. Coax shall be CT–070–50 or equivalent. NOT used in normal operation.

** Input resistor and cap are for test only. They are NOT necessary for normal operation.

Figure 6



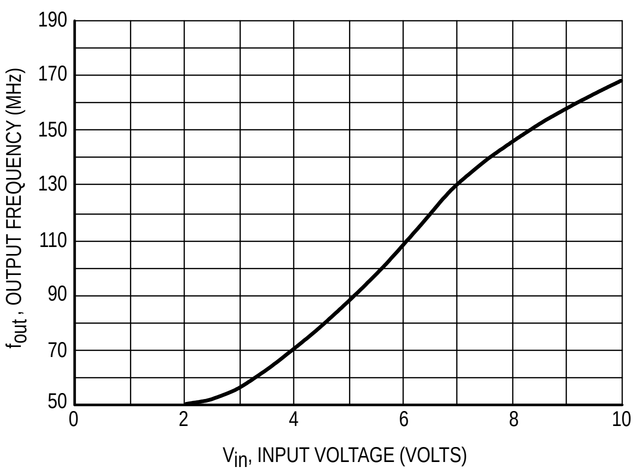
L: Micro Metal Toroidal Core #T44–10, 20 turns of No. 22 copper wire.



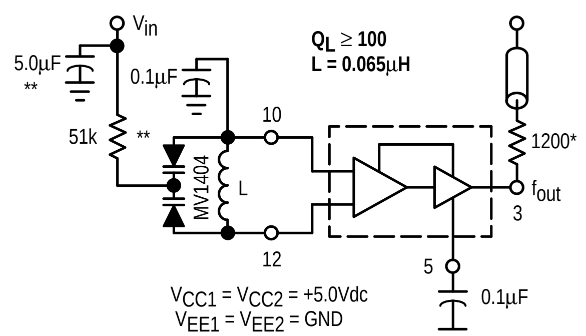
* The 1200 ohm resistor and the scope termination impedance constitute a 25:1 attenuator probe. Coax shall be CT–070–50 or equivalent. NOT used in normal operation.

** Input resistor and cap are for test only. They are NOT necessary for normal operation.

Figure 7



L: Micro Metal Toroidal Core #T30–12, 6 turns of No. 22 copper wire.



* The 1200 ohm resistor and the scope termination impedance constitute a 25:1 attenuator probe. Coax shall be CT–070–50 or equivalent. NOT used in normal operation.

** Input resistor and cap are for test only. They are NOT necessary for normal operation.

Figure 8

Typical transfer characteristics for the oscillator in the voltage controlled mode are shown in Figure 6, Figure 7 and Figure 8. Figure 6 and Figure 8 show transfer characteristics employing only the capacitance of the varactor diode (plus the input capacitance of the oscillator, 6.0pF typical). Figure 7 illustrates the oscillator operating in a voltage controlled mode with the output frequency range limited. This is achieved by adding a capacitor in parallel with the tank circuit as shown. The 1.0kΩ resistor in Figure 6 and Figure 7 is used to protect the varactor diode during testing. It is not necessary as long as the dc input voltage does not cause the diode to become forward biased. The larger–valued resistor (51kΩ) in Figure 8 is required to provide isolation for the high–impedance junctions of the two varactor diodes.

The tuning range of the oscillator in the voltage controlled mode may be calculated as:

$$\frac{f_{max}}{f_{min}} = \frac{\sqrt{C_D(max) + C_S}}{\sqrt{C_D(min) + C_S}}$$

where $$f_{min} = \frac{1}{2\pi\sqrt{L(C_D(max) + C_S)}}$$

CS = shunt capacitance (input plus external capacitance)

CD = varactor capacitance as a function of bias voltage

Good RF and low–frequency bypassing is necessary on the power supply pins. (See Figure 2)

Capacitors (C1 and C2 of Figure 4) should be used to bypass the AGC point and the VCO input (varactor diode), guaranteeing only dc levels at these points.

For output frequency operation between 1.0MHz and 50MHz a 0.1µF capacitor is sufficient for C1 and C2. At higher frequencies, smaller values of capacitance should be used; at lower frequencies, larger values of capacitance. At high frequencies the value of bypass capacitors depends directly upon the physical layout of the system. All bypassing should be as close to the package pins as possible to minimize unwanted lead inductance.

The peak–to–peak swing of the tank circuit is set internally by the AGC circuitry. Since voltage swing of the tank circuit provides the drive for the output buffer, the AGC potential directly affects the output waveform. If it is desired to have a sine wave at the output of the MC1648, a series resistor is tied from the AGC point to the most negative power potential (ground if +5.0 volt supply is used, –5.2 volts if a negative supply is used) as shown in Figure 10.

At frequencies above 100 MHz typ, it may be desirable to increase the tank circuit peak–to–peak voltage in order to shape the signal at the output of the MC1648. This is accomplished by tying a series resistor (1.0kΩ minimum) from the AGC to the most positive power potential (+5.0 volts if a +5.0 volt supply is used, ground if a –5.2 volt supply is used). Figure 11 illustrates this principle.

APPLICATIONS INFORMATION

The phase locked loop shown in Figure 9 illustrates the use of the MC1648 as a voltage controlled oscillator. The figure illustrates a frequency synthesizer useful in tuners for FM broadcast, general aviation, maritime and landmobile communications, amateur and CB receivers. The system operates from a single +5.0Vdc supply, and requires no internal translations, since all components are compatible.

Frequency generation of this type offers the advantages of single crystal operation, simple channel selection, and elimination of special circuitry to prevent harmonic lockup. Additional features include dc digital switching (preferable over RF switching with a multiple crystal system), and a broad range of tuning (up to 150MHz, the range being set by the varactor diode).

The output frequency of the synthesizer loop is determined by the reference frequency and the number programmed at the programmable counter; $f_{out} = Nf_{ref}$. The channel spacing is equal to frequency (f_{ref}).

For additional information on applications and designs for phase locked–loops and digital frequency synthesizers, see Motorola Brochure BR504/D, Electronic Tuning Address Systems, (ETAS).

Figure 10 shows the MC1648 in the variable frequency mode operating from a +5.0Vdc supply. To obtain a sine wave at the output, a resistor is added from the AGC circuit (pin 5) to V_{EE}.

Figure 11 shows the MC1648 in the variable frequency mode operating from a +5.0Vdc supply. To extend the useful range of the device (maintain a square wave output above 175Mhz), a resistor is added to the AGC circuit at pin 5 (1.0 kohm minimum).

Figure 12 shows the MC1648 operating from +5.0Vdc and +9.0Vdc power supplies. This permits a higher voltage swing and higher output power than is possible from the MECL output (pin 3). Plots of output power versus total collector load resistance at pin 1 are given in Figure 13 and Figure 14 for 100MHz and 10MHz operation. The total collector load includes R in parallel with R_p of L1 and C1 at resonance. The optimum value for R at 100MHz is approximately 850 ohms.

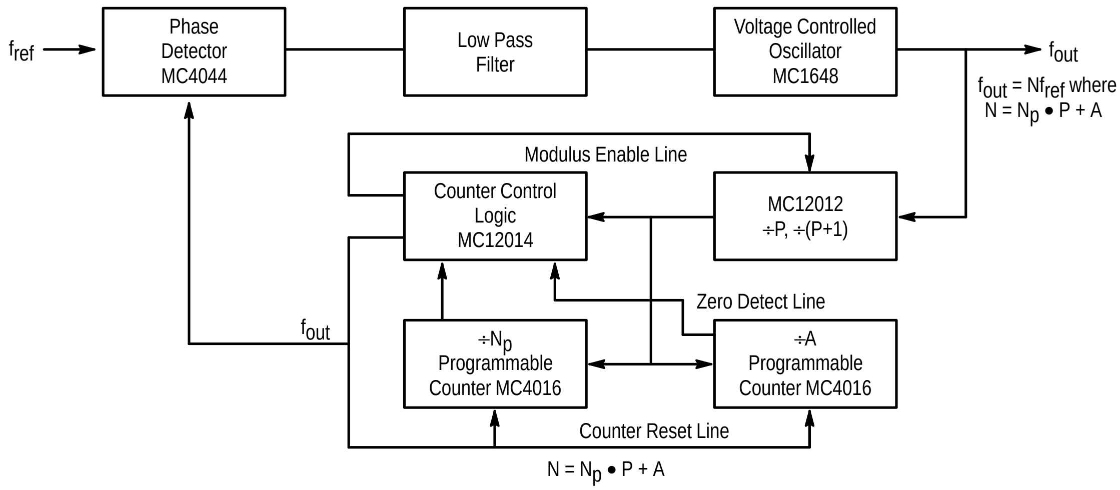


Figure 9. Typical Frequency Synthesizer Application

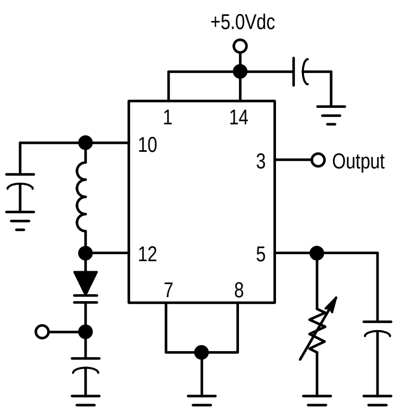


Figure 10. Method of Obtaining a Sine–Wave Output

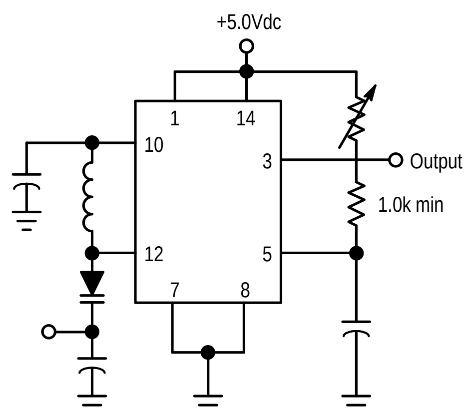


Figure 11. Method of Extending the Useful Range of the MC1648 (Square Wave Output)

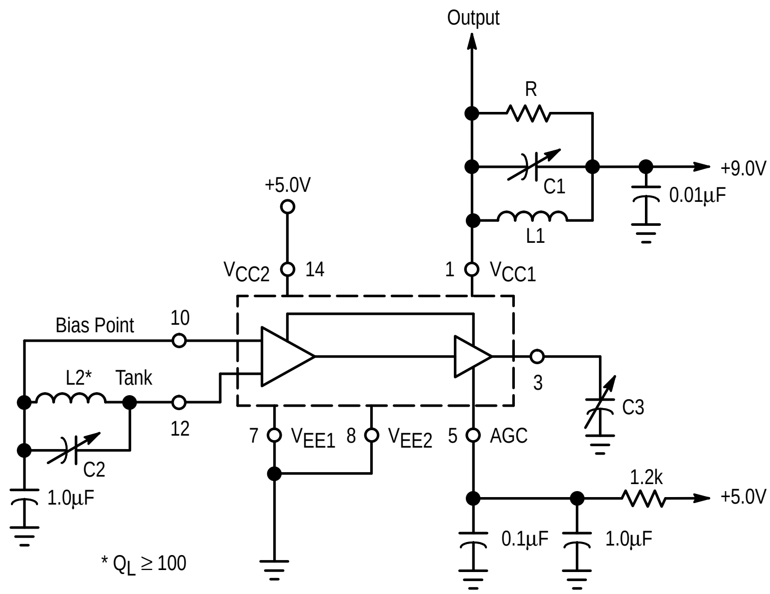
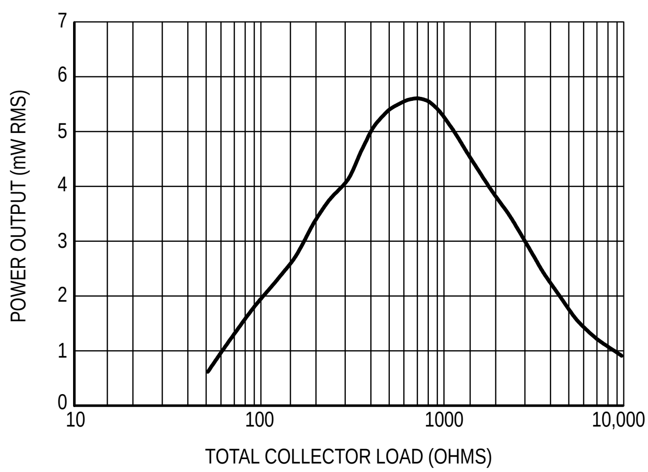
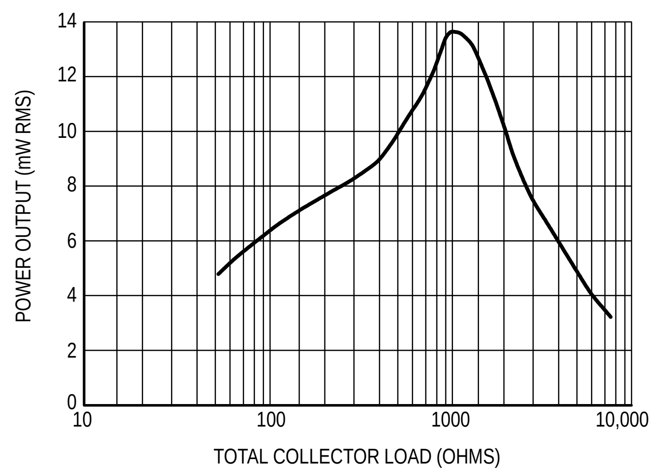


Figure 12. Circuit Used for Collector Output Operation



See test circuit, Figure 12, f = 100MHz
C3 = 3.0–35pF
Collector Tank
L1 = 0.22µH C1 = 1.0–7.0pF
R = 50Ω–10kΩ
R_P of L1 and C1 = 11kΩ @ 100MHz Resonance
Oscillator Tank
L2 = 4 turns #20 AWG 3/16” ID
C2 = 1.0–7.0pF

Figure 13. Power Output versus Collector Load

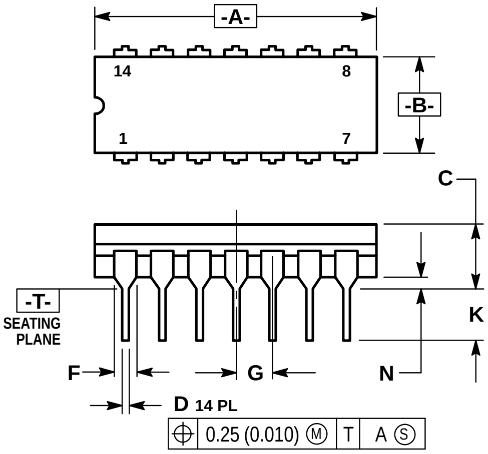


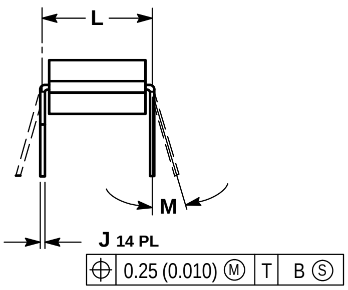
See test circuit, Figure 12, f = 10MHz
C3 = 470pF
Collector Tank
L1 = 2.7µH C1 = 24–200pF
R = 50Ω–10kΩ
R_P of L1 and C1 = 6.8kΩ @ 10MHz Resonance
Oscillator Tank
L2 = 2.7µH
C2 = 16–150pF

Figure 14. Power Output versus Collector Load

OUTLINE DIMENSIONS

L SUFFIX
CERAMIC PACKAGE
CASE 632–08
ISSUE Y

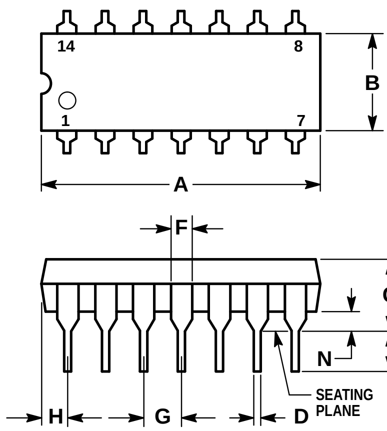


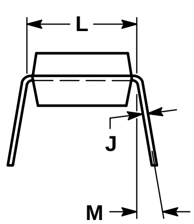


NOTES:
1. DIMENSIONING AND TOLERANCING PER ANSI Y14.5M, 1982.
2. CONTROLLING DIMENSION: INCH.
3. DIMENSION L TO CENTER OF LEAD WHEN FORMED PARALLEL.
4. DIMESNION F MAY NARROW TO 0.76 (0.030) WHERE THE LEAD ENTERS THE CERAMIC BODY.

	INCHES		MILLIMETERS	
DIM	MIN	MAX	MIN	MAX
A	0.750	0.785	19.05	19.94
B	0.245	0.280	6.23	7.11
C	0.155	0.200	3.94	5.08
D	0.015	0.020	0.39	0.50
F	0.055	0.065	1.40	1.65
G	0.100 BSC		2.54 BSC	
J	0.008	0.015	0.21	0.38
K	0.125	0.170	3.18	4.31
L	0.300 BSC		7.62 BSC	
M	0°	15°	0°	15°
N	0.020	0.040	0.51	1.01

P SUFFIX
PLASTIC PACKAGE
CASE 646–06
ISSUE L

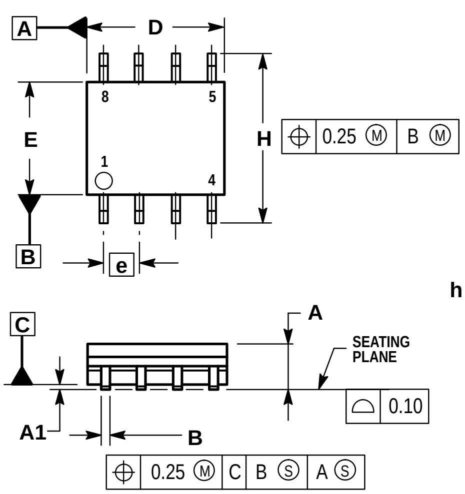


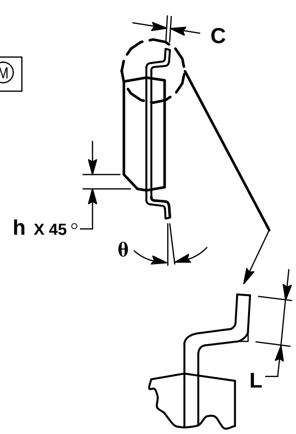


NOTES:
1. LEADS WITHIN 0.13 (0.005) RADIUS OF TRUE POSITION AT SEATING PLANE AT MAXIMUM MATERIAL CONDITION.
2. DIMENSION L TO CENTER OF LEADS WHEN FORMED PARALLEL.
3. DIMENSION B DOES NOT INCLUDE MOLD FLASH.
4. ROUNDED CORNERS OPTIONAL.

	INCHES		MILLIMETERS	
DIM	MIN	MAX	MIN	MAX
A	0.715	0.770	18.16	19.56
B	0.240	0.260	6.10	6.60
C	0.145	0.185	3.69	4.69
D	0.015	0.021	0.38	0.53
F	0.040	0.070	1.02	1.78
G	0.100 BSC		2.54 BSC	
H	0.052	0.095	1.32	2.41
J	0.008	0.015	0.20	0.38
K	0.115	0.135	2.92	3.43
L	0.300 BSC		7.62 BSC	
M	0°	10°	0°	10°
N	0.015	0.039	0.39	1.01

D SUFFIX
PLASTIC SOIC PACKAGE
CASE 751–05
ISSUE R



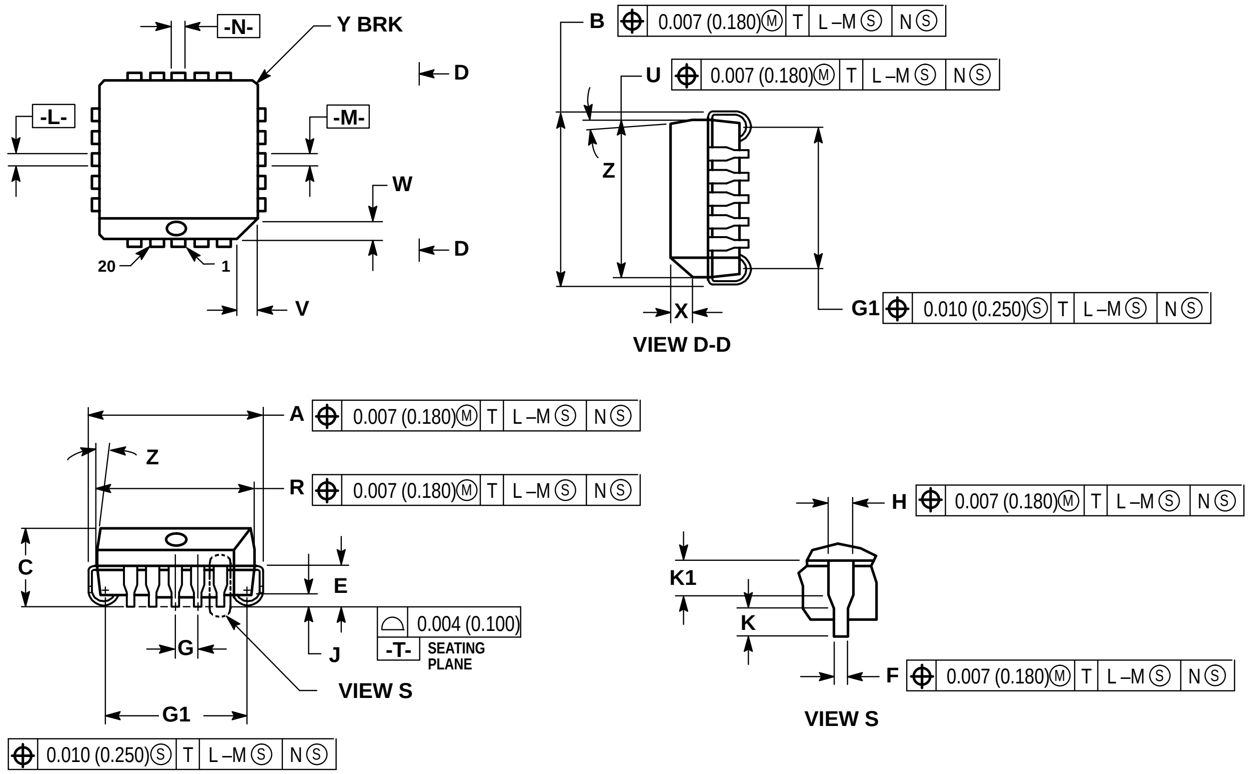


NOTES:
1. DIMENSIONING AND TOLERANCING PER ASME Y14.5M, 1994.
2. DIMENSIONS ARE IN MILLIMETERS.
3. DIMENSION D AND E DO NOT INCLUDE MOLD PROTRUSION.
4. MAXIMUM MOLD PROTRUSION 0.15 PER SIDE.
5. DIMENSION B DOES NOT INCLUDE MOLD PROTRUSION. ALLOWABLE DAMBAR PROTRUSION SHALL BE 0.127 TOTAL IN EXCESS OF THE B DIMENSION AT MAXIMUM MATERIAL CONDITION.

	MILLIMETERS	
DIM	MIN	MAX
A	1.35	1.75
A1	0.10	0.25
B	0.35	0.49
C	0.18	0.25
D	4.80	5.00
E	3.80	4.00
e	1.27 BSC	
H	5.80	6.20
h	0.25	0.50
L	0.40	1.25
θ	0°	7°

OUTLINE DIMENSIONS

FN SUFFIX
PLASTIC PLCC PACKAGE
CASE 775–02
ISSUE C



NOTES:
1. DATUMS -L-, -M-, AND -N- DETERMINED WHERE TOP OF LEAD SHOULDER EXITS PLASTIC BODY AT MOLD PARTING LINE.
2. DIM G1, TRUE POSITION TO BE MEASURED AT DATUM -T-, SEATING PLANE.
3. DIM R AND U DO NOT INCLUDE MOLD FLASH. ALLOWABLE MOLD FLASH IS 0.010 (0.250) PER SIDE.
4. DIMENSIONING AND TOLERANCING PER ANSI Y14.5M, 1982.
5. CONTROLLING DIMENSION: INCH.
6. THE PACKAGE TOP MAY BE SMALLER THAN THE PACKAGE BOTTOM BY UP TO 0.012 (0.300). DIMENSIONS R AND U ARE DETERMINED AT THE OUTERMOST EXTREMES OF THE PLASTIC BODY EXCLUSIVE OF MOLD FLASH, TIE BAR BURRS, GATE BURRS AND INTERLEAD FLASH, BUT INCLUDING ANY MISMATCH BETWEEN THE TOP AND BOTTOM OF THE PLASTIC BODY.
7. DIMENSION H DOES NOT INCLUDE DAMBAR PROTRUSION OR INTRUSION. THE DAMBAR PROTRUSION(S) SHALL NOT CAUSE THE H DIMENSION TO BE GREATER THAN 0.037 (0.940). THE DAMBAR INTRUSION(S) SHALL NOT CAUSE THE H DIMENSION TO BE SMALLER THAN 0.025 (0.635).

	INCHES		MILLIMETERS	
DIM	MIN	MAX	MIN	MAX
A	0.385	0.395	9.78	10.03
B	0.385	0.395	9.78	10.03
C	0.165	0.180	4.20	4.57
E	0.090	0.110	2.29	2.79
F	0.013	0.019	0.33	0.48
G	0.050 BSC		1.27 BSC	
H	0.026	0.032	0.66	0.81
J	0.020	—	0.51	—
K	0.025	—	0.64	—
R	0.350	0.356	8.89	9.04
U	0.350	0.356	8.89	9.04
V	0.042	0.048	1.07	1.21
W	0.042	0.048	1.07	1.21
X	0.042	0.056	1.07	1.42
Y	—	0.020	—	0.50
Z	2°	10°	2°	10°
G1	0.310	0.330	7.88	8.38
K1	0.040	—	1.02	—

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ASIA / PACIFIC: Motorola Semiconductors H.K. Ltd.; 8B Tai Ping Industrial Park,
51 Ting Kok Road, Tai Po, N.T., Hong Kong. 852–26629298

Appendix D

DataSheet for the PP3 not available at Present.