3V RF-Transmitter Utilizing Fairchild's HiSeC Encoder

Fairchild
Application Note 1021

INTRODUCTION

In the recent years short range unlicensed RF links have become very popular for a wide variety of applications in the field of remote control, data exchange, keyless entry systems, and others. The most popular frequency in Europe for these links is 433.92 MHz.

In order to get the approval for a license-free operation the systems have to comply to the European I-ETS 300 220 regulations. They are essentially valid all over Europe. The main parameters for portable class Ia transmitters are:

Frequency range	433.05 MHz–433.79 MHz
Temperature range	-10°C to +55°C
Radiated power	ERP ≤ 10 dBm (Note 1)
Harmonics	ERP ≤ -36 dBm (Note 1)

Note 1: I-ETS 300 220 specification: substitution method, dipol reference, main directivity.

Systems for the specified use are high volume products and therefore they are very cost sensitive. So the design goal has to be a product which is as cost-effective as possible.

Small hand held SAW stabilized single stage AM-transmitters with an integrated antenna can be economically realized. They use a minimum of components, have no tuning points, and are easy to produce.

The availability of new encoders for 3V operation—as the FSC HiSeC encoder—allow the realization of power effective transmitters which are operated by a single 3V lithium cell.

The transmitter described in this application note is considered to be a guide which can be used as a basis for the customer's own development. The customer has to adapt the circuit to his technical and production requirements.

BASIC OSCILLATOR DESIGN

There are many design topologies in which oscillators can be configured. The more popular ones are the Pierce and the Colpitts circuits which are shown as simplified block diagrams in Figure 1 and Figure 2.

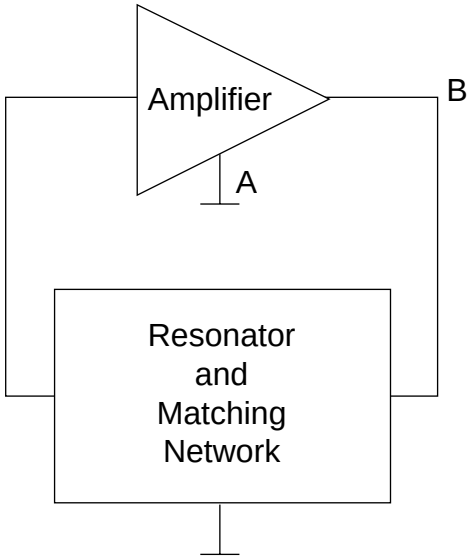
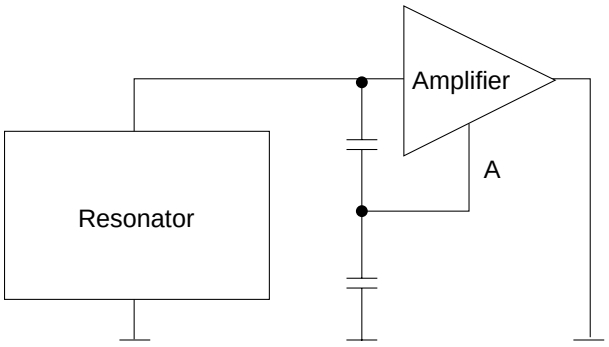
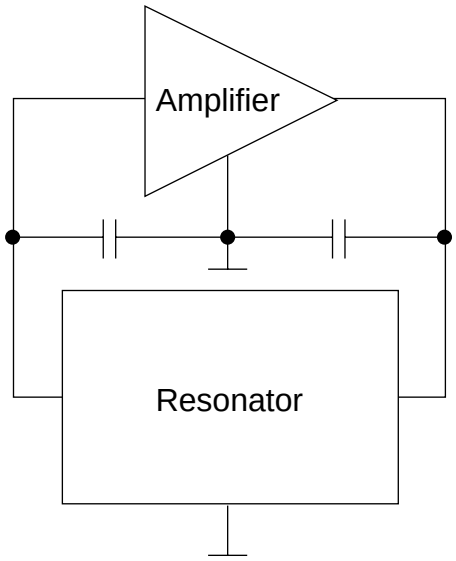


FIGURE 1. Pierce Oscillator

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a.



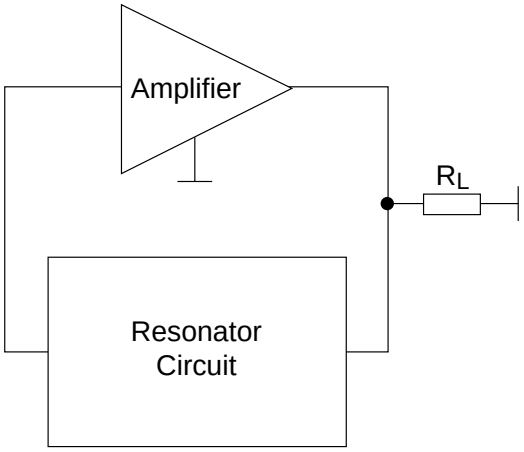
b.

FIGURE 2. Colpitts Oscillator

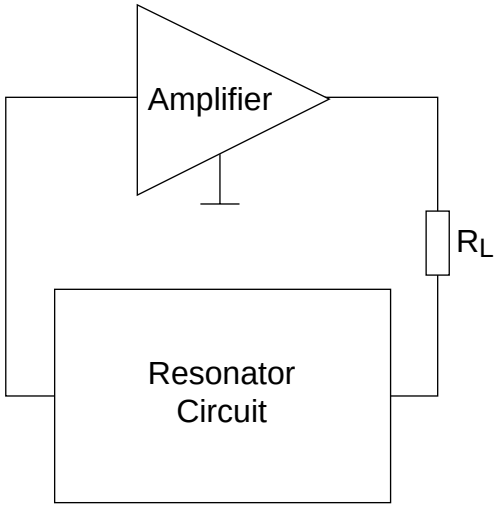
Rearranging the Colpitts oscillator of Figure 2 by changing the ground to point A results in a structure as shown in Figure 2.

Since the divider capacitors C1 and C2 can be seen as part of the amplifier or the resonator, both structures of Figure 2 and Figure 2 are identical. The only difference can be found in the grounding of either point A (Pierce) or B (Colpitts).

The load can be coupled to the circuit either across the amplifier output or between the amplifier output and the input of the resonator as can be seen from Figure 3.



a.



b.

FIGURE 3. Pierce Oscillator

Though the arrangement of Figure 3 is more common with the Colpitts configuration, it will be used in this design of a Pierce oscillator.

The feedback path typically contains a frequency selective network which in the case of a SAW oscillator is a surface acoustic wave resonator. It can be either a one- or a two-port device. In the latter case the common electrode of the SAW resonator has to be grounded.

The basic criteria for oscillation are:

— Open loop gain greater than unity at resonance.

— Phase shift within the loop must be either 0 or 360 degrees at resonance.

Because of the difficulties in calculating and measuring open loop parameters (the open ports must be terminated correctly with the circuit parameters!) it is more practical to consider the oscillator as a one port device which generates a negative resistance $R_0 < 1$ to the load R_L.

The network will oscillate when

$$R_C = R_0 || R_L \leq 0. \quad (1)$$

The resonant frequency occurs where the phase is

$$\phi = \phi_0 + \phi_L = 0. \quad (2)$$

In practice the oscillator will be designed for a loop gain of approximately 10 dB to maintain oscillation under all operating conditions. This value corresponds to the load resistance R_L being three times the negative resistance magnitude of R_0. The gain margin will be reduced to zero as soon as oscillation builds up and the increasing voltage brings the amplifier into its nonlinear region and into compression. Excessive gain should be avoided because it results in rich harmonics whose radiation is difficult to suppress.

With a practical transmitter design additional, partly diverging aspects have to be considered. The efficiency should be as high as possible. For a given battery the maximum pulsed current is limited. So in order to get a good range of the radio system the radiated power (ERP) should be as high as possible. The main sources of power loss are collector efficiency, matching and filter loss, and antenna efficiency.

The radiated power of the harmonics has to be low to comply with the regulations.

The transmitter parameters have to be stable against variations of temperature, battery voltage, and circuit component values.

Degradation in ERP, unacceptable frequency shift, and negative influence on the modulation characteristic under the influence of hand or metal proximity to the antenna shall be kept to a minimum.

DC BIAS CONSIDERATIONS

The purpose of a good bias design is to select the proper quiescent point and hold the quiescent point constant over variations in transistor parameters, battery voltage V_{CC}, and temperature. A resistor bias network can be used with good results if the circuit allows a voltage drop of approximately 1V across an emitter resistor. This results in a good stabilization against variations in DC current gain hfe and temperature. In a high efficient single stage transmitter this voltage drop can not be accepted because it lowers the usable voltage V_{CE}.

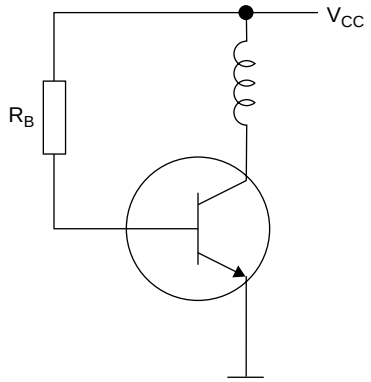


FIGURE 4. Simple Bias Network

The bias network used in the demonstration transmitter is the simplest possible realization (Figure 4). For a given base resistor RB the collector current can be easily calculated as

$$I_C = \frac{V_{CC} - V_{BE}}{R_B} \bullet h_{fe} \quad (3)$$

The small emitter resistor used in the demonstration circuit has been neglected in this calculation. Its purpose is to give some RF decoupling to the encoder circuit; the small DC voltage drop can be ignored. As can be seen from (3), as long as $V_{CC} >> V_{BE}$, there is only a minor temperature dependence found in I_C due to the temperature coefficient of V_{BE} which is approximately $\Delta V_{BE}/\Delta T \approx -2$ mV/K. However, I_C is directly proportional to the DC current gain h_{fe}. Unfortunately, this is a transistor parameter which can vary by quite some degree so in most practical cases R_B has to be adapted to the hfe of the transistor used.

If this “tuning” shall be avoided a much more complicated bias network has to be used. The suggested circuit (Figure 5) makes use of the combination of a (small) voltage drop across the bypassed emitter resistor R_E and a base voltage V_B which compensates for the temperature dependence of V_{BE} and the variation of V_{CC} by the use of two diodes.

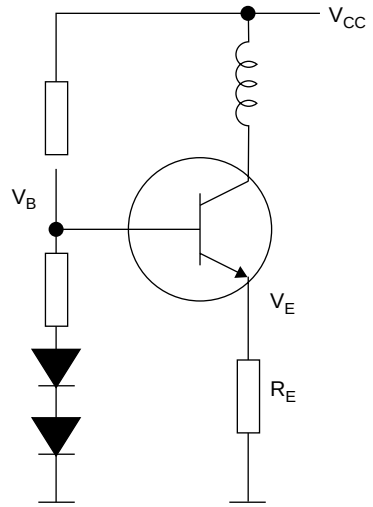


FIGURE 5. Proposed Bias Network for Improved Bias Stabilization

A careful design of this network for a particular transistor results in a good stabilization of the quiescent point (I_C) over variations in transistor parameters, temperature, and battery voltage.

It should be pointed out that a base network with a low resistance (compared to the input resistance of the transistor) requires a much more stable quiescent point to assure a stable oscillator function. In this case the limiting and stabilization of the RF amplitude occurs primarily by the decreasing RF input resistance with increasing base emitter voltage. In the case of a high resistive base network (as in the circuit of Figure 5) the stabilization of the RF amplitude is primarily caused by a rectifying process over the base emitter diode which results in a lower bias current I_C and hence a lower transconductance and power gain of the transistor with increasing amplitude. This automatic stabilization is quite effective and compensates for some parameter variations.

ANTENNA DESIGN

Usually space is the limiting factor for the design of a high efficient antenna in small hand-held transmitters. In case of the proposed demonstration PCB layout the area covered by the antenna has the dimensions of 17mm by 28mm.

Within this area an electrical (capacitive) antenna could be realized. It gives good results concerning easy matching and high efficiency. However, the efficiency decreases significantly if the antenna is covered by the hand of the operator at some short distance. This loss can easily reach -10 dB. Theoretical calculations show a slightly better and more stable efficiency for a magnetic (loop) antenna covering the same board area under the same influence of hand proximity.

The following calculation of the efficiency of the loop antenna used on the demonstration board shall show some rules on how to design such an antenna successfully.

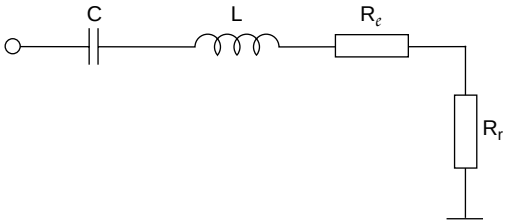


FIGURE 6. Equivalent Circuit of the Tuned Loop Antenna

As can be seen from Figure 6 the equivalent circuit of the tuned antenna comprises the following components:

The loop inductance L. It can be calculated according to [1]. With l being the loop circumference and d the conductor diameter we get

$$L = \frac{2 \bullet \ell}{cm} \bullet \ln\left(\frac{\ell}{d} - 1.6\right) nH$$

$$\approx 17 \bullet \ln\left(\frac{86}{2} - 1.6\right) nH \quad (4)$$

$$= 64\ nH$$

The radiation resistance R_r depends on the loop area A and the wavelength λ [1]:

$$R_R = 31200 \bullet \left(\frac{A}{\lambda^2}\right)^2 \Omega$$

$$= 31200 \bullet \left(\frac{16 \times 27}{690^2}\right)^2 \Omega \quad (5)$$

$$= 26\ m\Omega$$

Unfortunately, the loss resistance R_l in practice is much larger than the radiation resistance. R_l comprises two components: inductor loss R_{li} and the series loss resistance R_{lc} of the tuning capacitor C.

At 434 MHz the skin effect determines the loss resistance R_{li}. Based on the conducting surface height of $\delta = 3.2\ \mu m$ for copper, the width of d = 2mm for the antenna conductor, and $\sigma = 58 \bullet 10^6$ S • m/mm being the conductivity of copper, the inductor resistance is [1]:

$$R_{\ell i} = \frac{\ell}{d} \bullet \frac{1}{2 \bullet \delta \bullet \sigma}$$

$$= \frac{86}{2} \bullet \frac{1}{2 \bullet 3.2\mu \bullet 58 \bullet 10^6} \Omega \quad (6)$$

$$= 115\ m\Omega$$

The calculated value is based on a copper surface without oxidation and without tin.

The different values of σ and δ for an oxidized or tinned surface raise the resistance up to $R_{l\,i} \approx 170\ m\Omega$. This value applies to a conductor being tinned on one side. So its obvious not to use tin in the top side of the antenna. It is rather convenient to cover the antenna conductor by solder resist.

In order to get the antenna into resonance at 433.92 MHz, a series capacitor is applied. A further capacitor is needed if some impedance matching has to be realized. This second capacitor does not influence the following calculation of the matching loss.

Both capacitors have losses which can be summarized in a quality factor Q. For good capacitors of size 0805 a typical value is Q ≈ 200.

The equivalent series loss resistance is

$$R_{\ell c} = \frac{|X_C|}{Q} = \frac{|X_L|}{Q} = \frac{64nH \bullet 2\pi \bullet 434MHz}{200} = \frac{175\Omega}{200} = 875m\Omega \quad (7)$$

The total loss resistance of the tuned antenna is:

$$R_e = R_{el} + R_{ec} \quad (8)$$

$= 115\ m\Omega + 875\ m\Omega = 990\ m\Omega.$

This yields to an antenna efficiency of

$$\zeta = \frac{R_r}{R_r + R_c} \approx \frac{26m\Omega}{990m\Omega} = 0.026\ (-158dB) \quad (9)$$

The efficiency of the magnetic antenna described above will now be compared to the efficiency of an electrical antenna covering the same board area of 17mm by 28mm.

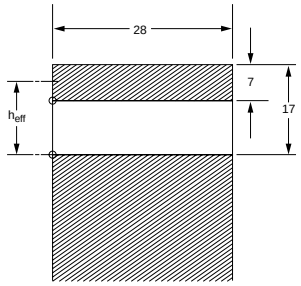


FIGURE 7. Structure of the Electrical Antenna

For the antenna structure as outlined in Figure 7 we find an equivalent circuit as of Figure 8.

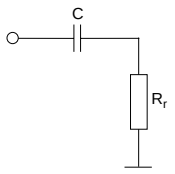


FIGURE 8. Equivalent Circuit of the Electrical Antenna

The antenna capacitance can be assessed to be

$C \approx 1\ pF\ (-j360\Omega).$ (10)

The radiation resistance will be [1]:

$$R_r = 160 \bullet \left(\frac{\pi \bullet h_{eff}}{\lambda}\right)^2 \Omega = 160 \bullet \left(\frac{\pi \bullet 13.5}{690}\right)^2 \Omega = 604\Omega \quad (11)$$

The antenna will be in resonance with a series inductor L:

$|X_L| = |X_C| = 360\Omega,$ (12)

$L = 140\ nH.$

Small inductors usable for this application show a realistic quality factor of Q ≈ 50 (e.g. a SIEMENS "Simid 01" or a Coilcraft 1206CS device). Its loss resistance RI has to be added to the circuit:

$$R_\ell = \frac{|X_L|}{Q} = 7.2\Omega \quad (13)$$

If some additional transformation is required a capacitor has to be added to form a L-C matching network. This additional component does not have too much influence on the calculation of the antenna efficiency because the Q of the capacitor is much higher than the Q of the inductor.

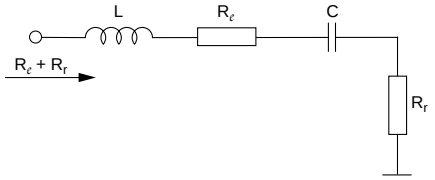


FIGURE 9. Equivalent Circuit of the Tuned Electrical Antenna

The resultant antenna efficiency becomes

$$\zeta = \frac{R_r}{R_r + R_\ell} = 0.077\ (-11.1dB) \quad (14)$$

Again, the effect of the losses in the matching components is obvious.

The efficiency is quite better than that found for the magnetic antenna. The additional loss of up to 10 dB by the close approximation of the hand or a finger to the electrical antenna gives comparable results for both, the electrical and the magnetic antenna.

ENCODER INTEGRATION

The Fairchild Semiconductor HiSeC encoder (part NM95HS01) is a very unique device well suited for car security and keyless entry systems. It shows some interesting features which allow the realization of very adaptable products.

One key feature certainly is the clock generator which can be realized by a simple and inexpensive RC configuration. Some other features which will be used in this design are the LED, TX, and $\overline{\text{RFEN}}$ outputs.

The LED pin is dedicated to a LED which is connected to V_{CC} via a current controlling resistor. The LED output is active low during the transmission of a pause between the data frames of the encoded bit stream.

If the encoder is programmed for a code without pauses between the data frames (which avoids problems in recovering the DC level at the data slicer in the receiver) the LED can be connected to the $\overline{RFEN}$ pin which is low during transmission of a pause and a frame.

The transmit output (TX) is a configurable logic level output which modulates the transmitter with the encoded bit stream.

The $\overline{RFEN}$ pin is active low during signal transmission. Its purpose is to provide power to the RF circuit only during transmission. Thus, battery life is increased by avoiding the leakage current of the RF transmitter to discharge the battery during off state.

The output characteristic of the pull-down transistor shows a voltage drop of typically 0.2V at a sink current of 8 mA. This voltage drop has to be considered when designing the DC bias network.

For the many other unique features of the HiSeC encoder please refer to the extended relevant documentation.

TRANSMITTER DESIGN EXAMPLE

The basic configuration of Figure 10has been chosen for the realization of the practical transmitter.

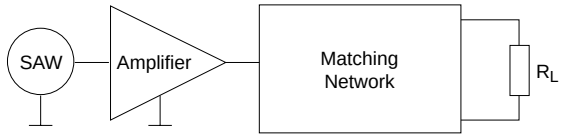


FIGURE 10. Basic Transmitter Circuit

The power the transmitter can be designed for is dictated by the capacity of the battery being used. It should have appropriate proportions for use in a small hand held transmitter.

Because of its widespread acceptance a CR2032 type 3V lithium battery is used. It can handle a pulse current of 8 mA, still maintaining a good capacity of approximately 150 mAh.

The transistor BFS17P is a good choice for this application. It is economical and shows good parameters at a collector current of 8 mA.

The SAW resonator is of the type SIEMENS R2632. It is a two port device with a nominal insertion shift of 180° at resonance. It can be replaced without major effect by a RFM type RP1303 device.

A linear approach is used to simulate the transmitter. The key to utilizing linear RF simulation involving nonlinear devices is to characterize them under their operating conditions. The large-signal parameters of the transistor used are not available so the simulation will make use of the small-signal parameters.

The rest of the oscillator loop circuitry is easily designed using linear models for the network elements.

The simulation has to take into account a parasitic inductance with each capacitor. It is the inductance of the mechanical structure of the component, its wiring, and any necessary via. Following the rule of 1 nH per mm total length the inductance easily reaches 5 nH even with 0805 size components.

The simulation itself makes use of the above mentioned one port negative impedance technique. When looking to the collector of the circuit a complex reflection coefficient $|\Gamma|$ greater than unity will be seen. The impedance at this point is the load resistance (transistor output and input of the feedback network) in parallel with a negative resistance caused by the positive feedback of the amplifier. This negative resistance has an absolute value which is the load resistance divided by the loop gain.

The result of a successive optimization process varying the values of the circuit components will be a $|\Gamma| > 1$ at the collector of the transistor at resonance. Varying the frequency, $\Gamma(\omega)$ shall be a circle symmetrical to the x-axis at the extended Smith chart (which includes some area of return gain $|\Gamma| > 1$).

The effective resistance at the collector terminal can be taken from the reflection coefficient at the point where its phase is 0° (for $\Gamma > 1$) or 180° (for $\Gamma < -1$).

The oscillator circuit of Figure 10 has been simulated within the frequency span of 431 MHz to 435 MHz. As a result of an interactive simulated tuning process a maximum value of $\Gamma = +11$ dB at $\phi = 0°$ has been achieved. This value is equivalent to a resistance of $R_C = -90\Omega$.

A rough estimation of R_L can be made by considering the voltage and power level at the collector of the transistor.

Operation in a quasi linear mode at an RF saturation voltage of $V_{CES} \approx 1.2V$ gives

$$V_{CE} = \frac{1}{\sqrt{2}} (V_{CE} - V_{CES})$$

$$= \frac{2.3 - 12}{\sqrt{2}} V \qquad (15)$$

$$= 0.78V$$

Assuming a collector efficiency of $\zeta_C = 0.25$ the RF power will be

$$P_C = V_{CE} \bullet I_C \bullet \zeta_C \qquad (16)$$

$$\approx 2.3V \bullet 8\ mA \bullet 0.25$$

$$= 4.6\ mW.$$

R_L can be calculated to be

$$R_L = \frac{V_{CE}^2}{P_C} \approx \frac{0.78V^2}{4.6mW} = 130\Omega \qquad (17)$$

From (1) R_0 can be calculated as $R_0 = -53\Omega$. It is near the aspired value of $|R_0| = R_L/3$.

The procedure of circuit optimizing applied with the circuit simulation can be very elegantly transferred to the process of fine-tuning the practical circuit on the board.

Connecting a network analyzer to the collector of the oscillator transistor allows a very effective optimization of the circuit. The components shall be tuned to get a circle for $\Gamma(\omega)$, symmetrical to the x-axis at the extended Smith chart. The intersection of the impedance circle with the real axis is located at R_C.

The transmitter design example was built using component values determined with the linear RF simulator optimizer.

The components on the demoboard transmitter (Figures 11, 12, 13, 14) have been fine-tuned for

— high output power,

— harmonic radiation below regulations,

— stable oscillation while detuning the antenna by nearby obstacles,

— temperature stability of oscillation and frequency,

— stable oscillation with variation of the battery voltage,

— insensitivity to component tolerances.

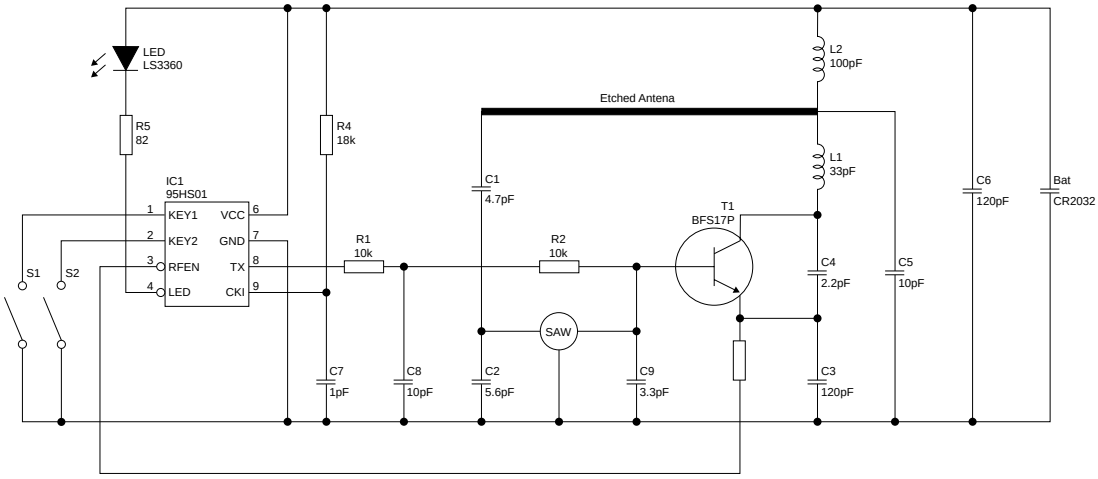


FIGURE 11. Demonstration Transmitter Circuit

Harmonic radiation has been located to be not only by the antenna itself but to a great extent by the wiring at the collector and the base port of the transistor. Both ports should be shortened for the harmonics by placing short wired capacitors there. The circuit of the demoboard (Figures 11, 12, 13, 14) even makes use of a series resonant circuit for the 2nd harmonic at the base of the transistor. Figure 11 shows the schematic of the final version of the complete transmitter.

POWER EFFICIENCY

Extended battery life and a maximum range make it desirable to get the most radiated RF power out of the applied DC power.

However, there are many sources of loss which limit the efficiency by quite some degree:

— Voltage drop across the battery.

— DC loss within the bias network.

— Collector efficiency.

— RF input driving power for the oscillator transistor.

— Matching and filtering loss.

— Antenna efficiency.

— Absorbing of RF power by obstacles near the antenna.

An estimation of the effective radiated power shall be made for the represented transmitter.

The battery data shows a typical voltage drop of 0.4V across the battery for a pulsed load current of 8 mA, so a usable supply voltage of V_{CC} = 2.6V is available. The DC voltage at the emitter of 0.3V reduces the available voltage across the transistor to V_{CE} = 2.3V. The DC base current can be neglected. The DC battery power of 3.0V • 8 mA=25 mW reduces to a power of 2.3V • 8 mA=18.4 mW available to the transistor.

Operation of the transistor at a low voltage V_{CE} results in a strong influence of the saturation voltage ($V_{CES} \approx$ 1.2V) on the AC behavior. At the quasi linear operating conditions a collector voltage of 800 mV and a collector efficiency of ζ = 0.25 can be expected. Therefore, the RF power available at the transistor output is

P_C = 18.4 mW • 0.25 = 4.6 mW. (18)

Driving power at the base of the transistor can be calculated from the transistor power gain of approximately 10 dB:

$$P_b = \frac{4.6mW}{10} \approx 0.5mW \quad (19)$$

Losses in the SAW resonator of -6 dB summarize to a total driving power of

$P_d = P_b \bullet 4 = 2$ mW. (20)

The power available to the antenna and its matching network becomes

$P = P_C - P_d$ =2.6 mW (4.1 dBm). (21)

Filter losses of -2 dB reduce the power delivered to the antenna to 2.1 dBm.

The calculated antenna efficiency of -15.8 dB results in a radiated power of -13.7 dBm.

Some possible directional gain of the antenna will be wasted by absorbing losses at obstacles in the vicinity of the antenna. So the expected radiated power will be

ERP = -13.7 dBm (43 μW). (22)

The calculated total efficiency of the RF transmitter becomes

$$\zeta = \frac{ERP}{P_B} = \frac{43\mu W}{3.0V \bullet 8mA} \quad (23)$$

$$= 1.8 \bullet 10^{-3} \text{ (-27dB)}$$

PCB LAYOUT OF TRANSMITTER

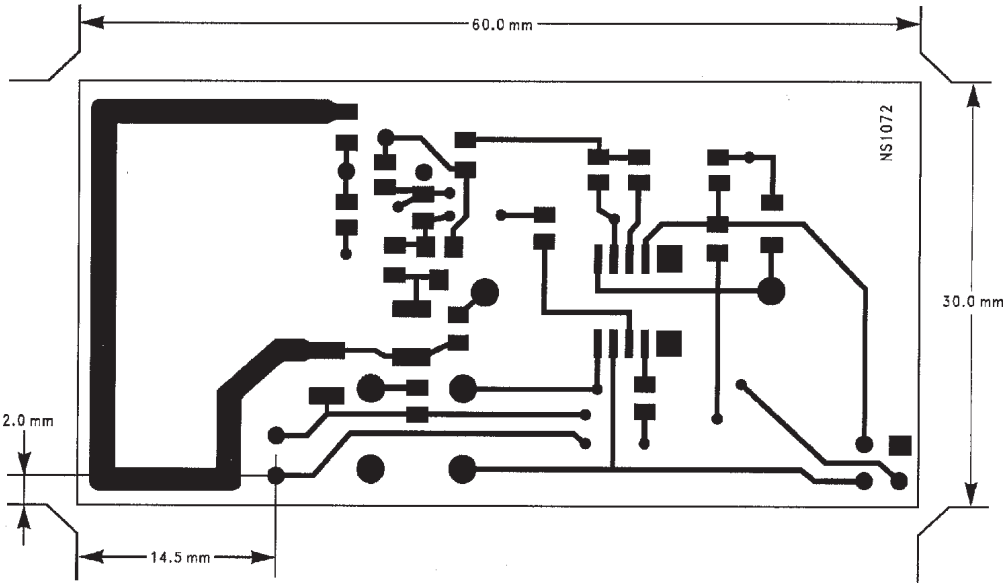


FIGURE 12. PCB Layout Top Layer, SMD Component Side

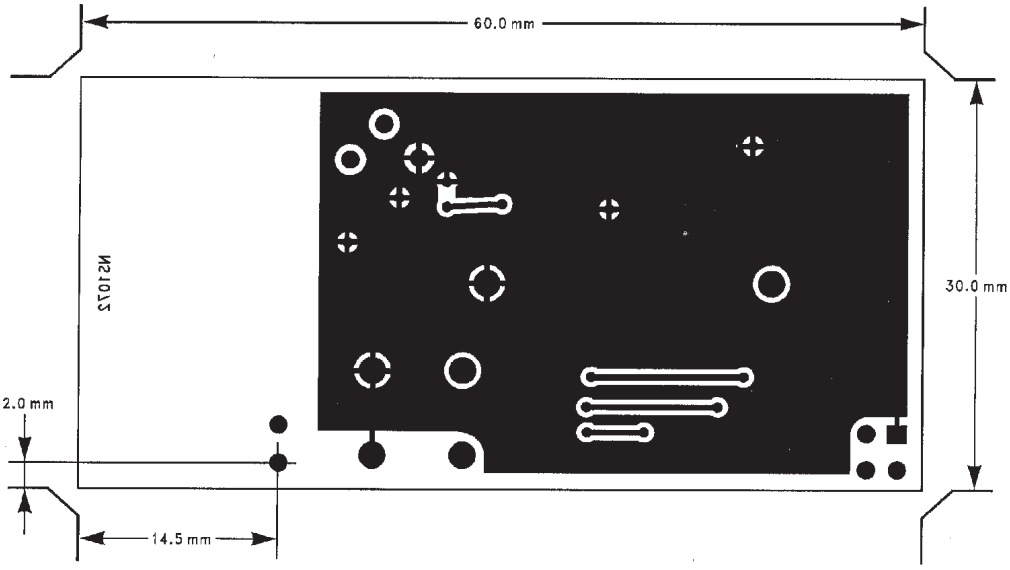


FIGURE 13. PCB Layout Bottom Layer, Wired Component Side

COMPONENT PLACEMENT

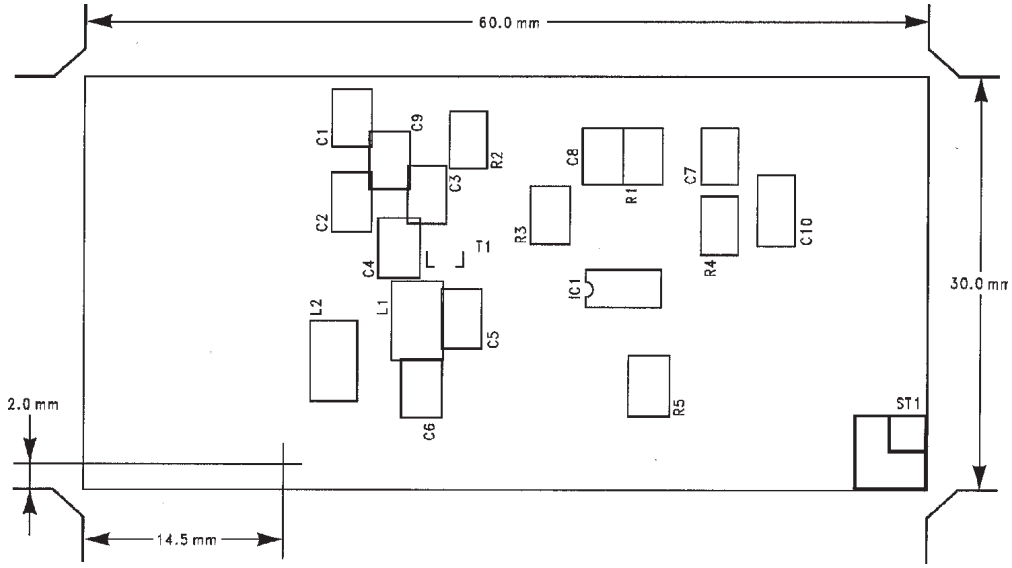


FIGURE 14. Component Placement of Transmitter. SMD Component Side

COMPONENT LIST

Circuit arrangement is according to Figure 11, placement as shown in Figure 14.

All chip components are of size 0805. Resistors have a manufacturing tolerance of 5%. Capacitors are of COG type, having a tolerance of 5%.

R1	10 kΩ
R2	10 kΩ
R3	12 kΩ
R4	18 kΩ
R5	82Ω
C1	4.7 pF
C2	5.6 pF
C3	120 pF
C4	2.2 pF
C5	10 pF
C6	120 pF
L1	33nH,S +M,Simid 0 1
L2	100nH,S +M,Simid 0 1
S1, S2	SKHL, ALPS
LED	LS3360
T1	BFS17P
IC1	NM95HS01, NS
SAW	R2632,S +M or RP1303, RFM

PERFORMANCE CHARACTERISTICS

Several samples of the final transmitter have been tested.

The following performance parameters have been achieved:

	Measured	Regulation I-ETS300220
Offset of operating frequency from SAW resonance	-18 kHz	N/A
Frequency shift with temperature		N/A
-20°C	-38 kHz	
-10°C	-18 kHz	
+55°C	-10 kHz	
+65°C	-22 kHz	
Frequency shift		N/A
capacitively loading the antenna by the hand extended metallic object at a distance of 5mm to the antenna	-5 kHz +10 kHz	
Radiated RF power	-12 dBm	≤ +10 dBm
Spurious emission		
2nd harmonic	-48 dBm	≤ -36 dBm
3rd harmonic	-50 dBm	≤ -30 dBm

	Measured	Regulation I-ETS300220
Power supply volt.		N/A
nominal min. operating	2.5V-3.3V 1.7V	
Operating current		N/A
RF transmitter, pulsed, peak total transmitter, average (depending on code format)	8.4 mA 4.5 mA	
Operating temp. nom.	-10°C ... +55°C	-10°C ... +55°C
Oscillator turn on time	80 µs	N/A
Oscillator turn off time	5 µs	N/A

CONCLUSION

It has been the purpose of this Application Note to show how an economic transmitter for applications in short range RF links in the 433.92 MHz band can be designed. Some general rules have been outlined and some specific problems have been highlighted in detail. This should help the customer to realize a product which is optimized to his requirements.

The development of the demoboard transmitter is based on the concept as described. It is a simple design of a single stage SAW based oscillator whose inductance of the tank circuit is partly realized as a printed loop antenna.

The transmitter is pulse modulated on and off by the coded pulse train generated by a HiSeC encoder. Pulse rates of up to 3 kbit/s can be handled. A radiated power of -12 dBm at a supply current of 8.5 mA peak out of a 3V lithium cell has been achieved.

[1]Meinke, Gundlach: “Taschenbuch der Hochfrequenztechnik”, Springer, Berlin 1986, Vol. 1.

[2]Zinke, Brunswig: “Lehrbuch der Hochfrequenztechnik”, Springer, Berlin 1990.

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